A CMOS Synchronized Sample-and-Hold Artifact Blanking Analog Front-End Local Field Potential Acquisition Unit With ±3.6-V Stimulation Artifact Tolerance and Monopolar Electrode-Tissue Impedance Measurement Circuit for Closed-Loop Deep Brain Stimulation SoCs

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Abstract—In this paper, an analog front-end (AFE) local-field potential (LFP) acquisition unit with real-time stimulation artifact removal is proposed and verified for closed-loop deep brain stimulation (DBS) applications. The proposed acquisition unit is called the synchronized sample-and-hold stimulation artifact blanking (SSAB) AFE LFP acquisition unit. Both right-leg-driven (RLD) circuit and monopolar electrode-tissue impedance (ETI) measurement circuit associated with the AFE amplifier are also proposed. During closed-loop stimulations, the artifact removal is realized through the SSAB-IPC by blanking the AR-CCIA with a clock synchronized to the stimulation-enabled signal and holding the amplifier at its state before stimulation through a sample-and-hold operation. After stimulation, the acquisition unit can quickly recover from the holding state back to the LFP recording state to reduce the discontinuity in LFP recording. The proposed acquisition unit was fabricated in 0.18-um CMOS technology. With the RLD circuit, the measured CMRR is 124 – 145 dB in the signal bandwidth. The fabricated monopolar ETI measurement circuit has a measurement error less than 8.3% with an extra power consumption of 2.65 µW.

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to control DBSs. The stimulation currents or voltages at C2 are coupled to C1 and C3 and generate stimulation artifact voltages. The artifact voltages on a pair of electrodes can be characterized by both common-mode artifact voltage (CMAV) and differential-mode artifact voltage (DMAV) as indicated in Fig. 1. Both CMAV and DMAV become smaller at a pair of electrode contacts farther from the stimulated contact. But they are all much larger than the LFPs to be recorded by AFE amplifiers [5]. Without appropriate artifact removal methods, the AFE amplifiers could be easily saturated by CMAV and DMAV during the stimulation period and stable LFP biomarkers could not be obtained for closed-loop DBSs. As a result, stimulation artifact removal from LFP recording is critically required and remains a research challenge in closed-loop DBS systems, especially for real-time processing.

So far, many artifact removal methods in AFE bio-signal recordings have been reported [5], [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16]. All the reported methods can be divided into four categories as filtering, template subtracting, blanking, and selective sampling. In the filtering method [5], [6], [7], [8], analog or digital filters are used to remove stimulation artifacts. It is only suitable for high-frequency stimulation and LFPs after filtering may be distorted in the frequency domain [9]. In the template subtracting method [10], [11], [12], [13], the artifact templates are estimated and subtracted from the LFPs in digital domain [10], [11] or front-end stage [12], [13] for artifact cancellation. The main difficulty of template subtracting method is that a small estimation error may induce large residual distortion in the recovered LFP signal. Offline operation is required to improve the estimation accuracy under different stimulation parameters. Thus, the template subtracting method may not be suitable for real-time closed-loop DBSs [9]. In the blanking method, AFE amplifiers are reset and disconnected from electrodes during the whole stimulation period [14] or every stimulation pulse [15]. The major drawbacks of blanking method are the long recovery time to recording after stimulation and the discontinuity in the recorded data [16].

Two artifact removal methods [9], [16] have been proposed recently. They are classified as the selective sampling method. In [9], the LFP sampling clock of analog-to-digital converter (ADC) is synchronized with the stimulation pulse so that the stimulation artefacts are never sampled. Thus, a true uninterrupted artefact-free LFP recording could be achieved. In [16], an irregular sampling method was proposed, where the artifact peaks were detected from the raw digitized recordings. Then the samples within the contaminated period of stimulation pulses were excluded and replaced with the interpolation of the samples prior to and after the stimulation artifact duration. Both proposed methods could real-time remove artifacts in digital domain. In [9] and [16], however, a large CMAV may cause output saturation in the front-end stage during signal sensing. This limits the artifact tolerance range.

With the advancement of semiconductor technologies, more and more research effort has been devoted to the development of closed-loop neuro-modulation system on a single chip as the system-on-chip (SoC). For example, a closed-loop neuro-modulation SoC for epileptic seizure control has been reported [14], [17], where AFE ECoG (electrocorticography) acquisition unit with amplifier and ADC, stimulators, bio-signal processor, and wireless power/data telemetry are integrated on a chip.

To realize a real-time closed-loop DBS SoC, the artifact removal method should be implemented in integrated circuit (IC) and integrated on-chip with AFE LFP acquisition unit. Among the reported artifact removal methods in AFE acquisition units [5], [6], [7], [8], [9], [10], [11], [12], [13], [14], [15], [16], only [8], [12], [13], [14], [15] have been realized in ICs or SoCs. The filtering method was adopted in [8] whereas the template subtracting method in [12] and [13]. The blanking method was used in [14] and [15]. As mentioned earlier, the above three artifact removal methods [8], [12], [13], [14], [15] have drawbacks and limitations in real-time closed-loop DBS SoCs.

In this paper, a CMOS AFE LFP acquisition unit with real-time stimulation artifact removal, called the synchronized sample-and-hold artifact blanking (SSAB) AFE LFP acquisition unit, is proposed. In the proposed SSAB AFE acquisition unit, the artifact removal is realized by blanking the first stage of AFE amplifier with a clock synchronized to the stimulation-enable signal and holding the amplifier at its state before stimulation through a sample-and-hold operation. This operation called the SSAB operation, is performed during the CCS. Thus, stimulation artifacts can be blocked from entering the amplifier and the amplifier can be recovered quickly back to LFP recording. The proposed SSAB operation can also be applied in CVS as well. The proposed SSAB AFE LFP acquisition unit can remove both CMAV and DMAV with a fast recovery time and a high signal linearity. No filters or extra digital signal processing units are required. Thus, it can facilitate real-time closed-loop DBS SoCs.

To eliminate large common-mode interferences including 50/60 Hz AC power-line noise and electromagnetic interference (EMI) coupled to the patient’s body through the skin, a right-leg drive (RLD) [18] circuit is designed and integrated with the proposed SSAB AFE amplifier to increase its common-mode rejection ratio (CMRR) so that the electrical regulation can be satisfied.
In closed-loop DBSs, it is required to measure the electrode-tissue impedance (ETI) before electrical stimulation to determine the suitable stimulation currents or voltages. Since monopolar CCS or CVS is adopted in DBSs, it is more appropriate to perform the ETI measurement in monopolar operation through only one electrode. So far, the reported ETI measurement are of bipolar operation through two electrodes [19], [20], [21]. In [20] and [21], the ETI measurement circuits can share part of the AFE amplifier to save both chip area and power dissipation. In the proposed SSAB AFE LFP acquisition unit, a monopolar ETI measurement circuit is designed and integrated with the SSAB AFE amplifier to share part of amplifier circuit. Moreover, a semi-empirical method is used in impedance calculation to reduce measurement errors.

The proposed SSAB AFE LFP acquisition unit with monopolar ETI measurement circuit and RLD circuit was designed and fabricated in 0.18-µm CMOS technology. The measurement results of fabricated chip have shown that it can blank out stimulation CMAV and DMAV up to ±3.6 V with 150 mV electrode DC offset (EDO) between electrodes. In addition, the measured total harmonic distortion (THD) of recovered sinewave signal under 130 Hz stimulation frequency is only 1.12%. With the designed RLD circuit, the measured CMRR is larger than 124 dB in the signal passband of 0.28 – 117 Hz. Moreover, the designed monopolar ETI measurement circuit has a measurement error less than 8.3% and a total 2.65-µW extra power consumption.

This paper is organized as follows. Section II describes the architecture of the proposed multi-channel SSAB AFE LFP acquisition unit along with multi-channel monopolar biphasic CCS circuits and a bio-signal processor (BSP). The circuit design details are described in Section III. The experimental results are presented in Section IV. Finally, a conclusion is given in Section V.

II. SYSTEM ARCHITECTURE

The architecture of the proposed SSAB AFE acquisition unit with monopolar ETI measurement circuit and RLD circuit is shown in Fig. 2 where 16-channel AFE acquisition units, 16-channel monopolar biphasic CCS circuits [22], and one BSP are also demonstrated. All the operating clocks and control signals of the acquisition unit are generated by the BSP. The proposed SSAB AFE acquisition unit consists of 16-channel AFE amplifiers, a 16-to-1 multiplexer (MUX) for channel selection, a switched-capacitor amplifier (SC-Amp) [23], and a 10-bit hybrid capacitor switching scheme successive-approximation-register analog-to-digital converter (SAR ADC) [24]. Each channel of AFE amplifier contains a SSAB-input protection circuit (SSAB-IPC), an auto-reset capacitively-coupled instrumentation amplifier (AR-CCIA) with a CCIA and a modified auto-reset unit (ARU) redesigned from that in [14], a 16th order RC-high-pass filter (RC-HPF), two buffers (BUF), and a programmable switched-capacitor low-pass filter (SC-LPF). The SSAB operation is performed through the SSAB-IPC to blank stimulation artifacts with control signals generated by the SSAB-IPC control. The AR-CCIA has a gain of 34 dB whereas the programmable SC-LPF provides an extra programmable gain of 26/36/46 dB for patient-specific LFP signal acquisition. The 1st-order RC-HPF is used to eliminate the output DC offset of AR-CCIA and other low frequency biomedical interferences. A pair of source-follower buffers [21] are placed after RC-HPF to isolate the loading effect of SC-LPF. The filtered and amplified LFPs are sampled with 4-kHz sampling rate and stored in the holding capacitors of SC-LPF [23]. After channel selection of MUX, a SC-Amp with 64-kHz sampling rate is used to drive the capacitor array of SAR ADC.

In the modified ARU shown in Fig. 2, two input nodes are connected to the output nodes of AR-CCIA to detect its output differential voltage. The reset operation called the auto-reset operation, is automatically performed when the differential output voltage of AR-CCIA reach 1.5 V or 0.3 V to avoid the saturation of AFE amplifier caused by unexpected charge accumulation. The modified ARU can also be operated by the control clock CLKA RU from the BSP to receive the control signal RST_AFE and perform the reset operation called the power-on-reset operation as requested by the BSP. The three output control signals from the modified ARU for IPC and pseudo-resistors in CCIA/RC-HPF are operated at the same time. When pseudo-resistors are reset during both power-on-reset operation and auto-reset operation, it takes about 567 mS to recover back to stable LFP recording operation.

In the monopolar biphasic CCS of Fig. 2, stimulation currents IDNO and ICAT are delivered through the stimulus driver to the output of stimulating channel connected with an electrode contact. The discharge circuit at the output of stimulating channel is used to discharge the electrode through MDIS_N and MDIS_P after CCS so that the residual charges can be reduced. During CCS, MDIS_N and MDIS_P are off. The control signals of MDIS_N and MDIS_P are generated by the discharge control with both discharge signal STI_DIS and stimulation enable signal STI_EN generated by the BSP. The input voltage Vip (Vin) of ith SSAB-IPC is connected to the drain of MDIS N and MDIS P (VREF) to sense LFP signals indirectly from the nth non-stimulating electrode contact through MDIS P with MDIS N off. To avoid device over stressing issue, the power supplies of stimulus driver are ±6 V and those of both SSAB-IPC control and discharge control are ±3 V. Other analog and digital circuits are powered with two separated +1.8-V supplies AVDD and DVDD, respectively, for noise isolation.

Although the stimulus driver is supplied with ±6 V, the maximum voltage across the gate/source/drain nodes of I/O devices MP A, MP B, MNA, and MNB is limited to ±3.6 V to avoid gate dielectric overstress problem, which could occur if the across voltage is larger than 3.96 V. Thus, the CCS current is limited to 3.6 (1.2) mA with the ETI of 1 (3) kΩ.

In the proposed monopolar ETI measurement circuit shown in Fig. 2, a constant current generator [25] is designed to generate a nearly process-independent constant current for measurement. A current switching circuit is designed with the ETI measurement control circuit to switch the current via the electrode contact to electrode-tissue interface under
measurement. The generated voltage on the electrode contact is measured by the SSAB AFE acquisition unit to obtain the ETI value. Thus, the required extra power dissipation is small. After measurement, the electrode contact is connected to ground to discharge the residual charges. The monopolar ETI measurement circuit is enabled by the control signal IMP_MODE and operated by the clock CLKIMP at 250 Hz.

In the RLD circuit, the common-mode sensing circuit (CSC) is used to sense the common-mode signals at the differential outputs of one AR-CCIA as shown in Fig. 2. The sensed common-mode signals are amplified by the low-pass gain stage with a low bandwidth so that the common-mode signals generated by AC power-line noise and EMI can be amplified and sent to the RLD electrode to form a negative feedback path. Thus, the common-mode signals can be further suppressed to boost the CMRR of 16-channel acquisition unit.

III. CIRCUIT DESIGN

The circuit design and operation of SSAB-IPC, AR-CCIA, monopolar ETI measurement circuit, and RLD circuit are described below.

A. Synchronized Sample-and-Hold Artifact Blanking-Input Protection Circuit (SSAB-IPC)

The circuit schematic diagrams of SSAB-IPC and AR-CCIA with the monopolar biphasic CCS circuit are shown in Fig. 3(a) whereas the corresponding waveforms of SSAB-IPC, discharge, and stimulation control signals are in Fig. 3(b). As shown in Fig. 3(a), the SSAB-IPC consists of PMOS (NMOS) switch MPA/MNA for artifact blanking with MPB/MNB as dummy switches at the reference node for transient reduction, two grounding switches MG, and two holding switches MH. MPA/MPB, MSA/MNB, and MG are implemented with I/O devices. Since the maximum stimulation voltage is limited to ±3.6 V in the CCS operation, these I/O devices can avoid gate dielectric overstress problem during stimulation. The holding switches MH are designed with small size nominal devices to reduce the charge injection effect.

During the monopolar biphasic CCS with the stimulation enable signal STI_EN high at 1.8 V, a negative stimulation current is applied to the stimulation channel when the negative stimulation control signal STI_CAT is high, followed by a positive current with the positive stimulator control signal STI_ANO high. Corresponding to negative and positive stimulation currents, the coupled artifact voltages $V_{\text{ARTI}}$ in other non-stimulating channels are negative and positive, respectively. As shown in the top waveform of Fig. 3(b), the total voltage $V_{\text{NE}}$ in the non-stimulating channel during CCS is expressed as $V_{\text{ARTI}} + V_{\text{LFP}} + V_{\text{EDO}}$, where $V_{\text{LFP}}$ is the voltage of LFP and $V_{\text{EDO}}$ is the electrode DC offset voltage between non-stimulating channel and reference channel. $V_{\text{ARTI}}$ is much larger than $V_{\text{LFP}} + V_{\text{EDO}}$. The designed SSAB-IPC can tolerate ±3.6-V $V_{\text{NE}}$.

To blank the stimulation artifact voltages and record LFPs in the non-stimulating channel, the SSAB operation is performed with the SSAB-IPC control signals shown in Fig. 3(b). To minimize the nonlinearity of SSAB operation, the SSAB-IPC control signals are designed with suitable timing sequence. During CCS with STI_EN high, $V_{\text{IPC-H}}$ is fallen to 0 V at the same time to turn off both MH devices so that the input voltage $V_{\text{LCCIA}} = V_{\text{NE}} = V_{\text{LFP}} + V_{\text{EDO}}$ sampled before $V_{\text{IPC-H}}$ dropped to 0 V can be held at both nodes C and D on the two input capacitors $C_{\text{IN}}$. Moreover, all the node voltages of AR-CCIA are also frozen and held at this moment. The holding time is as long as that of STI_EN high. This realizes
the SSAB operation. After a 15-nS delay from the falling edge of \( V_{\text{PC,H}} \), \( V_{\text{PC,G}} \) is risen to 3 V to turn on both M_{G} devices and keep both nodes A and B grounded for the elimination of transient coupling. After a 15-nS delay from the rising edge of \( V_{\text{PC,G}} \), the gate voltage \( V_{\text{PC,P}} \) (\( V_{\text{PC,N}} \)) of M_{PA}/M_{PB} (M_{NA}/M_{NB}) is risen (fallen) from −3 V (+3 V) to 0 V as shown in Fig. 3(b) to turn off both switches. After 10 \( \mu \)S from the rising edge of \( \text{STI}_{\text{EN}} \), CCS starts and \( V_{\text{AR}} \) appears in the non-stimulating channel. When a negative \( V_{\text{NE}} \) appears at the left (drain) node of M_{PA}, it remains off to block the negative voltage. When a positive \( V_{\text{NE}} \) appears at the left (source) node of M_{PA}, it is turned on to pass \( V_{\text{NE}} \) to the off M_{NA} so that the positive \( V_{\text{NE}} \) is blocked.

When \( \text{STI}_{\text{EN}} \) is dropped to 0 V and CCS is ended, \( V_{\text{PC,P}} = −3 \) V and \( V_{\text{PC,N}} = 3 \) V after a 30-nS delay to turn on M_{PA}/M_{PB} and M_{NA}/M_{NB}. Then \( V_{\text{PC,G}} = 0 \) V after a 15-nS delay to turn off M_{G}. After another 15-nS delay, \( V_{\text{PC,H}} = 1.8 \) V to turn on M_{H} in the SSAB-IPC and the AR-CCA is in the LFP recording operation. With the SSAB operation, the recovery time of AR-CCA from the SSAB operation to the LFP recording operation is quite short as 230 \( \mu \)S. Thus, the discontinuity in the recorded data can be minimized.

To realize the SSAB operation with a small loss on LFP recording, the stimulation-enable signal \( \text{STI}_{\text{EN}} \) generated by the BSP is used to synchronize both stimulator and SSAB-IPC. \( \text{STI}_{\text{EN}} \) rises at the time \( t_{\text{gap1}} \) earlier than \( \text{STI}_{\text{CAT}} \) and falls \( t_{\text{gap2}} \) later than \( \text{STI}_{\text{ANO}} \), as shown in Fig. 3(b). Both \( t_{\text{gap1}} \) and \( t_{\text{gap2}} \) are designed to avoid the instant voltage overstressing caused by transient voltage peaking of simultaneous device ON/OFF switching. The duration \( t_{\text{STI}_{\text{EN}}} \) of \( \text{STI}_{\text{EN}} \) can be written as

\[
\text{STI}_{\text{EN}} = t_{\text{gap1}} + t_{\text{CAT}} + t_{\text{gap2}}
\]

where \( t_{\text{gap2}} \) is the inter-phase gap time interval between negative and positive pulses in biphasic CCS and \( t_{\text{CAT}} \) (\( \text{I}_{\text{ANO}} \)) is the negative (positive) stimulation pulse width. With \( t_{\text{gap1}} = t_{\text{gap2}} = 10 \mu \)S and \( t_{\text{CAT}} = \text{I}_{\text{ANO}} = 60 \mu \)S, \( t_{\text{STI}_{\text{EN}}} \) is designed to be 150 \( \mu \)S. For a longer stimulation with larger \( t_{\text{CAT}} \) and \( \text{I}_{\text{ANO}} \), \( t_{\text{STI}_{\text{EN}}} \) is larger and the discontinuity in the recorded data is increased.

As shown in Fig. 3(b), \( V_{\text{DIS,P}} = 0 \) V and \( V_{\text{DIS,N}} = 0 \) V during CCS to turn off both M_{DIS,P} and M_{DIS,N} in the stimulating channel. These two devices can tolerate the maximum stimulation voltages of \( ±3.96 \) V. After CCS, \( \text{STI}_{\text{EN}} \) is low and \( \text{STI}_{\text{DIS}} \) is high to make \( V_{\text{DIS,P}} = −3 \) V and \( V_{\text{DIS,N}} = 3 \) V. Thus, M_{DIS,P} and M_{DIS,N} are turned on for 2 mS to discharge the electrode. During the discharge time, the AR-CCA is still connected to the stimulating channel via its SSAB-IPC. But the recorded data in this time are not used by the BSP. In the non-stimulating channel, \( V_{\text{DIS,P}} = −3 \) V and \( V_{\text{DIS,N}} = 0 \) V to turn on M_{DIS,P} and turn off M_{DIS,N}. The SSAB-IPC is connected to the electrode contact through M_{DIS,P} and the AR-CCA is in the LFP recording operation. LFP signals from the electrode contact can pass through both M_{DIS,P} and SSAB-IPC to the input of the AR-CCA for signal amplification.

**B. Auto-Reset-Capacitive-Coupled Instrumental Amplifier (AR-CCA)**

The circuit structure of AR-CCA is shown in Fig. 3(a) where it consists of an inverter-based folded-cascode (IBFC)
operational amplifier (OP) [23], two feedback pseudo-resistors $R_{FB}$ [17], two input capacitors $C_{IN}$, two feedback capacitors $C_{FB}$, and three output capacitors $C_{L1}/C_{L2}/C_{L3}$. The T-network pseudo-resistor $R_{FB}$ is composed of three PMOS pseudo-resistors connected as a T-network for larger equivalent resistance as shown in Fig. 3(a). Thus, the amplifier can achieve a low cut-off frequency below 0.5 Hz with a smaller feedback capacitors $C_{FB}$.

The circuit of the IBFC OP is shown in Fig. 4 where the folded-cascode PMOS M11 and M12 are adopted. The increased voltage headroom of folded-cascode structure allows the folded-cascode PMOS M11 and M12 are adopted. The input-referred noise voltage $v_{n,AR−CCIA}$ of the AR-CCIA can be related to $v_{n,OP}$ of the IBFC OP by [17]

$$v_{n,AR−CCIA} = \frac{C_{IN} + C_{FB} + C_p}{C_{IN}} \cdot v_{n,OP}$$

(3)

where $C_{IN}$ is the input (feedback) capacitor shown in Fig. 3(a) and $C_p$ is the device input capacitance of OP. Large gate area of M3-M6 in IBFC OP can suppress the flicker noise, but increase $C_p$ and $v_{n,AR−CCIA}$. Thus, proper sizing for M3-M6 was performed to minimize $v_{n,AR−CCIA}$.

The loading capacitors $C_{L1}/C_{L2}/C_{L3}$ are used to limit the bandwidth of IBFC OP and form a low-pass corner of 3.26 kHz. The capacitor $C_{L3} = 6$ pF at differential output increases equivalently 2C$_{L3}$ to Vop$_{pre}$ and Von$_{pre}$. Two capacitors $C_{L1} = C_{L2} = 5$ pF are loaded at each side of IBFC OP for stability requirement of CMFB. The three loading capacitors are realized by MiM capacitors.

The 1st-order RC-HPF connected to the output nodes of IBFC OP is implemented by a T-network PMOS pseudo-resistor $R_{HP}$ and a capacitor $C_{HP}$ as shown in Fig. 3(a). The function of HPF is to filter the output offset voltage of IBFC OP, mainly generated by two T-network pseudo-resistors $R_{FB}$. Assuming perfect match in differential paths, the total differential voltage gain $A_{TD}$ of both AR-CCIA and HPF can be derived as

$$A_{TD}(s) = \frac{V_{O,HPF}(s)}{V_{I,CCIA}(s)} = \frac{C_{IN}}{C_{FB}} \cdot \frac{(s + R_{FB} C_{FB})(s + R_{HP} C_{HP})(1 + s C_{L1} C_{IN})}{(1 + s R_{FB} C_{FB})(1 + s R_{HP} C_{HP})(1 + s C_{L1} C_{IN})}$$

(4)

where $g_{mIBFC}$ is the transconductance of IBFC OP. Because $C_{HP}$ is 100 times of $C_{FB}$, we have $R_{HP} C_{HP} \gg R_{FB} C_{FB}$ and the high-pass cutoff frequency is dominated by $\frac{1}{2 \pi R_{FB} C_{FB}}$. As may be seen from (4), the low-pass cutoff frequency is $\frac{g_{mIBFC} C_{FB}}{2 \pi C_{IN}}$. To obtain the mid-band gain of 34 dB, $C_{IN} = 5$ pF and $C_{FB} = 100$ pF are chosen. The resultant high-pass (low-pass) cutoff frequency is 0.28 Hz (3.26 kHz). To enhance gain accuracy, both $C_{IN}$ and $C_{FB}$ are realized by MiM capacitors.

Fig. 5 shows the circuits of two 2-stage source-follower buffers (BUFs) where the first (second) stage is composed of a PMOS (NMOS) source-follower with a PMOS (NMOS) current source. They are used to isolate the loading effect of SC-LPF on the 1st-order RC-HPF. The sizes of PMOS $M_{B2}/M_{B6}$ and NMOS $M_{B3}/M_{B7}$ in PMOS and NMOS source followers, respectively, are designed to achieve nearly equal gate-source voltages and threshold voltages so that $V_{OP,BUF} = V_{OP,PRE}$.

The input-referred noise density $v_{n,OP}$ of IBFC OP can be derived by calculating output noise voltage $V_{n,OP}$ of IBFC OP first, and then referring back to the input nodes ($V_{IN,pre}$, $V_{IN,pre}$) by dividing the open-loop gain of IBFC OP. Since the cascode devices contribute negligible noise at low frequency, only the noises of M3,4, M5,6, and M15,16 are considered. The derived $\frac{V_{n,OP}^2}{\Delta f}$ can be written as

$$\frac{V_{n,OP}^2}{\Delta f} = \frac{16}{3} K T \left( g_{m3,4} + g_{m5,6} + g_{m15,16} \right) \left( g_{m3,4} + g_{m5,6} \right)^2 + 2 \left( \frac{K_p g_{m3,4}^2}{\omega_{op} \cdot f} + \frac{K_p g_{m5,6}^2}{\omega_{op} \cdot f} + \frac{K_p g_{m15,16}^2}{\omega_{op} \cdot f} \right) \left( g_{m3,4} + g_{m5,6} \right)^2$$

(2)

where $\Delta f$ is the bandwidth of amplifier, $g_{m,i,j}$ is the transconductances of MOS devices Mi and Mj in differential paths, K is the Boltzmann constant, T is the absolute temperature, $K_p$ ($K_p$) is the transconductance parameter of PMOS (PMOS), $C_{oxN}$ ($C_{oxP}$) is the gate oxide capacitance per unit area of NMOS (PMOS), $f$ is the signal frequency, and $W_{i,j}$ ($L_{i,j}$) is the width (length) of MOS devices Mi and Mj. As compared with the conventional folded-cascode amplifier, the IBFC OP can achieve better noise performance because of higher input transconductance ($g_{m3,4}+g_{m5,6}$). To reduce the input-referred noise and increase the gain of OP, a large current of 855 nA is designed for M3-M6 whereas a small current of 50 nA for M11-M16 to obtain small $g_{m15,16}$ and high output resistance. All the 4 input transistors M1-M4 are operated in the subthreshold region to achieve a high transconductance efficiency. The input-referred noise voltage $v_{n,AR−CCIA}$ of the AR-CCIA can be related to $v_{n,OP}$ of the IBFC OP by [17]
**C. Right-Leg-Driven (RLD) Circuit**

The schematic of RLD circuit is shown in Fig. 6(a) whereas its testing RC model in Fig. 6(b). As may be seen from Fig. 6(a), the RLD circuit has a common-mode sensing circuit (CSC) to sense the common-mode voltage $V_{OUT_{CSC}}$ at the output nodes of AR-CCIA. $V_{OUT_{CSC}}$ is sent to the low-pass gain stage consisted of an OP, a feedback pseudo-resistor realized by the PMOS $M_{RFB_{RLD}}$ with an on-chip bias voltage $V_{B5} = 0.65$ V, and a feedback capacitor $C_{FB_{RLD}} = 1.9$ pF. The pseudo-resistance of $M_{RFB_{RLD}}$ is 840 MΩ. With $M_{RFB_{RLD}}$ and $C_{FB_{RLD}}$, the low-pass gain stage has a low-pass corner frequency $1/(2\pi \cdot R_{FB_{RLD}} C_{FB_{RLD}}) = 100$ Hz to amplify the common-mode signals generated by AC power-line noise or EMI. With the +1.8-V power supply, $R_S = 36.2$ kΩ is used to limit the output DC current of OP to be less than 50 μA to prevent from irrecoverable tissue damage if the body of patient is grounded [18]. An off-chip 47-μF DC blocking capacitor $C_S$ is placed in series with $R_S$ to isolate the output common-mode voltage from the body potential.

Through $C_S$, the amplified common-mode signals are sent to the RLD electrode to form a negative feedback path. This leads to the degradation of common-mode signals and the increase of CMRR.

In Fig. 6(b), the RLD output voltage $V_{IN_{RLD}}$ is connected to the equivalent RC circuit of RLD electrode-tissue interface whereas CH1-16 and $V_{REF}$ are connected to their equivalent RC circuits of electrode-tissue interfaces. $V_{50/60}$ is used as the AC power-line noise voltage whereas the LFP signal voltage $V_{LFP}$ is connected between nodes A and B. Through the testing RC model of Fig. 6(b), the performance of RLD circuit can be simulated or tested. The testing RC model is referred from the international standard IEC80601-2-26 for medical electrical equipment. The model in Fig. 6(b) with $V_{LFP} = 0$ and the common-mode signal voltage $V_{cm}$, can be used for CMRR simulation or measurement.

The CSC circuit is shown in Fig. 7 where a differential source-coupled input pair is used to sense the output common-mode voltage of $V_{op_{pre}}$ and $V_{on_{pre}}$ and compare with $V_{CM} = 0.9$ V. Through the folded-cascode structure with current-mirror load $M_{R9-14}$, $V_{OUT_{CSC}}$ is generated and sent to the low-pass
Fig. 9. Timing diagram of operational clocks and output signal waveforms in the monopolar ETI measurement circuit.

Fig. 10. Chip photograph of the fabricated 16-channel SSAB AFE LFP acquisition unit with one channel as test key design, as highlighted by red dotted lines in a closed-loop SoC.

gain stage of RLD circuit. The closed-loop common-mode gain $A_{RLD}(s)$ can be derived from the circuits of Figs. 6(a), 6(b) and 7 as

$$A_{RLD}(s) = \frac{A_{CCIA}}{1 + A_{CCIA}(gm_{R3} + gm_{R6})} \times \frac{R_{FB,RLD}}{1 + sR_{FB,RLD}C_{FB,RLD}}$$

$$\approx \frac{1}{(gm_{R3} + gm_{R6})} \frac{R_{FB,RLD}}{1 + sR_{FB,RLD}C_{FB,RLD}}$$

(5)

where $A_{CCIA}$ is the common-mode gain of the AR-CCIA and $R_{FB,RLD}$ is the resistance of pseudo-resistor $M_{RFB,RLD}$. The mid-band gain of $A_{RLD}(s)$ is approximately equal to

$$1/[(gm_{R3} + gm_{R6})R_{FB,RLD}]$$

which is very small. This leads to a large CMRR of the AR-CCIA with RLD circuit.

Fig. 11. The measured frequency response of the fabricated SSAB AFE LFP amplifier with (a) programmable mid-band gains and (b) programmable low-pass corner frequencies.
Fig. 12. The measured input-referred noise of the fabricated SSAB AFE LFP amplifier.

Fig. 13. The measured CMRR of the fabricated SSAB AFE LFP acquisition unit with and without the RLD circuit.

D. Monopolar Electrode-Tissue Impedance (ETI) Measurement Circuit

The proposed monopolar ETI measurement circuit is shown in Fig. 8 where two channels of SSAB AFE amplifiers are demonstrated for impedance measurement. The equivalent RC circuits of electrode-tissue interfaces at two DBS electrode contacts of channels CH1/CH2 (ground contact V_{REF}) are also shown at the input nodes of SSAB_IPCs where the double layer capacitance C_{dl} (C_{dlg}) in parallel with the charge transfer resistance R_p (R_{pg}) is connected in series with the tissue resistance R_t (R_{tg}). In the ETI measurement control, a 6-bit counter with the clock CLK_IMP of 250 Hz and the control signal IMP_MODE, a 5-to-32 decoder, and a CSG (current switch grounding) generator is used to generate control signals IMP_Ctrl1, IMP_Ctrl2, ..., and IMP_Ctrl16. Clock buffers [20] are connected to the decoder output to generate D0, D2, ..., D30, and CSG for switching of impedance measurement current I_IMP = 260 nA from ground to each electrode contacts. Thus, the I_IMP switching can be faster with small transient voltages [20].

Fig. 9 shows the timing diagram of operational clocks and output signal waveforms in the monopolar ETI measurement circuit. When IMP_MODE is enabled, all SSAB AFE amplifiers perform the reset operation. Then CSG = 1.8 V to switch I_IMP to ground via M_{S0} whereas IMP_Ctrl1/IMPCtrl2 and D0/D2 are high to turn on switches M_P and M_N of two SSAB_IPCs, respectively, to discharge residual charges for 24 mS at all electrode-tissue interfaces. With a 4-mS elapse time, IMP_Ctrl1 and D0 are high. M_{S1} is turned on to switch I_IMP from ground via switches M_P and M_N of SSAB-IPC1 and stimulator discharge MOS device M_{DIS_P1} to both CH1 and ground electrode-tissue interfaces of channel CH1 for 4 mS. The generated voltage is amplified by the SSAB AFE amplifiers. The amplified output voltage V_{O_HPF1} at the output nodes of 1^{st}-order RC-HPFs are shown in Fig. 9. Using the measured voltage and I_IMP, the total impedance amplitude R_T can be obtained. Neglecting the effect of R_p (R_{pg}), the amplitude of electrode-tissue interface impedance R_{ETI} is dominated by C_{dl} and C_{dlg}. Thus, R_T

Fig. 14. The measured input impedance of the fabricated SSAB AFE LFP acquisition unit.

Fig. 15. (a) Measured signal waveforms of AR-CCIA and SC-Amp output voltages and (b) measured power spectrum of rebuilt SC-Amp output voltage in the fabricated SSAB AFE LFP amplifier.
can be expressed as
\[ R_T = R_{ETI} + R_s \cong \frac{1}{2\pi C_{dl}} + \frac{1}{2\pi C_{dlg}} + R_s \] (6)
where \( R_s \) is total series resistance of the channel resistances of \( M_{DIS_P}, M_P, \) and \( M_N \). Since \( R_s \) is known, it can be subtracted from \( R_T \) to obtain \( R_{ETI} \).

After impedance measurement at CH1, \( \overline{DO} \) is high until the end of ETI impedance measurement to discharge CH1 so that the residual charges generated during impedance measurement can be discharged. CSG is also high to switch \( I_{IMP} \) back to ground. After \( \overline{DO} \) high for 4 mS, \( IMP_{Ctrl2} \) and \( D2 \) are high to start the similar ETI measurement on CH2. This operation is repeated until the ETI measurement on CH16 is completed.

IV. EXPERIMENTAL RESULTS

The proposed SSAB AFE LFP acquisition unit has been designed and integrated in a closed-loop SoC which was fabricated in TSMC 0.18-\( \mu \)m CMOS technology. The chip photograph is shown in Fig. 10 where the 15-channel SSAB AFE LFP amplifiers with one channel as the test key design of AFE amplifier is highlighted with red dotted lines. The whole chip area is 20 mm\(^2\) where each channel of SSAB AFE LFP acquisition unit occupies 0.285 mm\(^2\). The power supplies AVDD (1.8 V), DVDD (1.8 V), and ±3 V of acquisition unit are on-chip generated. Electrical tests were performed to verify the functions of the proposed SSAB AFE LFP acquisition unit.

The measured frequency response of the fabricated SSAB AFE LFP amplifier is shown in Fig. 11. The measured mid-band programmable gains of the AFE acquisition unit are 34.4/59.5/68.9/79.4 dB and the high-pass corner frequency is at 0.28 Hz as shown in Fig. 11(a). The measured programmable low-pass corner frequencies are at 41.9/117/467/940 Hz as shown in Fig. 11(b). The measured input-referred noise integrated from 0.28 Hz to 117 Hz is 1.95 \( \mu V_{\text{rms}} \) with the noise floor at 160 nV\( \text{rms/}\sqrt{\text{Hz}} \) as shown in Fig. 12. The measured total power consumption is 5.92 \( \mu W \) per channel. The noise efficiency factor (NEF) [26] is 6.88.

Fig. 13 shows the measured CMRR of the SSAB AFE LFP acquisition unit with and without the RLD circuit. The testing RC model as shown in Fig. 6(b) is connected to CH1-CH16 and \( V_{\text{REF}} \) of AFE amplifiers and \( V_{\text{IN,RLD}} \) of RLD circuit to form the feedback loop for common-mode signals. The measured CMRR is 73.9 – 95.2 dB and 124 – 145 dB in the AFE bandwidth without and with the RLD circuit, respectively. It is verified that with the RLD circuit, CMRR at 60 Hz is larger than 120 dB as required by the electrical regulation.

The input impedance of AFE amplifier should be much larger than the ETI to avoid signal degradation. Fig. 14 shows the measured input impedance of the fabricated SSAB AFE LFP acquisition unit where the measured input impedance is 160 M – 6.87 G\( \Omega \) in the pass-band 0.28 Hz to 117 Hz. The input impedance is much larger than the ETI of DBS electrode, which is typically 1 – 3 k\( \Omega \).

In the linearity measurement, the measured total harmonic distortion (THD) of the fabricated SSAB AFE LFP amplifier is 0.89 % under a pure sinusoidal signal input at 27.34375 Hz and 1.6-V\( \text{dd} \) differential output peak-to-peak voltage. To verify the artifact blanking capability of the proposed SSAB AFE LFP amplifier, a 27.34375-Hz sine wave mixed with 150mV EDO and ±3.6 V artifact waveforms under 60-\( \mu \)S 130-Hz biphasic CCS pulses, is fed to the input of fabricated acquisition
Fig. 18. (a) Pre-recorded human LFP raw data without stimulation artifacts where the zoom-in figure is also shown in the right. (b) LFPs mixed with VSTIM of ±3.6 V, 130 Hz, and 60 µS width as LFPs with stimulation artifacts where the zoom-in figure is shown in the right. (c) Rebuilt LFPs after the SSAB operation of the fabricated unit where the zoom-in figure is shown in the right. (d) Power spectrum of the LFP raw data in (a) (e) Power spectrum of the LFPs with stimulation artifacts in (b). (f) Power spectrum of the amplified LFPs when DBS is OFF and no stimulation artifact appears. (g) Power spectrum of the amplified LFPs with stimulation artifacts when DBS is ON.

TABLE I
THE PERFORMANCE SUMMARY AND COMPARISON OF THE PROPOSED SSAB AFE LFP ACQUISITION UNIT

<table>
<thead>
<tr>
<th></th>
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<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Application</td>
<td>LFP, ECoG</td>
<td>LFP, AP</td>
<td>LFP, AP</td>
<td>ECoG</td>
<td>ECoG</td>
</tr>
<tr>
<td>Supply Voltage (V)</td>
<td>1.8, 3, –3, 6</td>
<td>2, –1.8</td>
<td>1.2</td>
<td>1.8, 3, 6, 9, 12</td>
<td>0.5, 2.5</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>16</td>
<td>16</td>
<td>12</td>
<td>16</td>
<td>64</td>
</tr>
<tr>
<td>Bandwidth (Hz)</td>
<td>0.28 – 41.9 / 117 / 467 / 940</td>
<td>2 – 3k</td>
<td>0.12 / 0.15 / 0.3 – 5k</td>
<td>0.59 – 117</td>
<td>1 – 1k</td>
</tr>
<tr>
<td>Gain (dB)</td>
<td>34.4 / 59.5 / 68.9 / 79.4</td>
<td>33.6</td>
<td>25.7</td>
<td>49.1 / 59.4 / 67.9</td>
<td>60</td>
</tr>
<tr>
<td>CM</td>
<td>±3.6</td>
<td>2</td>
<td>650m</td>
<td>10</td>
<td>2.5</td>
</tr>
<tr>
<td>DM</td>
<td>70m</td>
<td>3 m</td>
<td>N/A</td>
<td>1.1</td>
<td>N/A</td>
</tr>
<tr>
<td>Recovery Time after Artifact Blanking (s)</td>
<td>230 µs</td>
<td>N/A</td>
<td>N/A</td>
<td>50</td>
<td>N/A</td>
</tr>
<tr>
<td>EDO Suppression Range (mV)</td>
<td>150</td>
<td>100</td>
<td>50</td>
<td>50</td>
<td>N/A</td>
</tr>
<tr>
<td>THD w/o stimulation (@0.8VDD output swing)</td>
<td>0.89%</td>
<td>N/A</td>
<td>0.015%</td>
<td>0.82%</td>
<td>&lt;1%</td>
</tr>
<tr>
<td>THD w/ stimulation</td>
<td>1.12%</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>Power per channel (µW)</td>
<td>5.92</td>
<td>100</td>
<td>2.8</td>
<td>3.26</td>
<td>2.98</td>
</tr>
<tr>
<td>NEF (AR-CCIA)</td>
<td>6.88</td>
<td>17.6</td>
<td>7.4</td>
<td>3.36</td>
<td>2.89</td>
</tr>
<tr>
<td>CMRR (dB)</td>
<td>73.9 – 95.2 (w/o RLD)</td>
<td>124 – 145 (w/ RLD)</td>
<td>130@60Hz</td>
<td>N/A</td>
<td>78</td>
</tr>
<tr>
<td>PSRR (dB)</td>
<td>69.4 – 71.2</td>
<td>74.1@60Hz</td>
<td>N/A</td>
<td>76</td>
<td>69.6</td>
</tr>
<tr>
<td>Input Impedance (Ω)</td>
<td>6.87G – 160M (DC – 100Hz)</td>
<td>N/A</td>
<td>1.6G – 20M (DC – 5kHz)</td>
<td>N/A</td>
<td>92M – 9.2M (DC – 2kHz)</td>
</tr>
<tr>
<td>Area per channel (mm²)</td>
<td>0.285</td>
<td>0.0322</td>
<td>0.069</td>
<td>0.22</td>
<td>0.0025 (with FPGA)</td>
</tr>
</tbody>
</table>

Fig. 15(a) shows the measured signal waveforms of AR-CCIA and SC-Amp output voltages. The waveforms of SC-Amp output voltage are re-sampled at 4 kHz and rebuilt as continuous waveforms for linearity analyses. The power spectrum of re-built SC-Amp output voltage signal is shown in Fig. 15(b). As seen from Fig. 15(b), the THD is only slightly increased from 0.89 % without stimulation artifacts to 1.12 % with artifacts. This is because that stimulation artifacts have been nearly blanked out by the SSAB operation with small residual artifacts at the AR-CCIA output. Moreover, the SSAB AFE LFP amplifier can hold the previous input signal on C_in unit and the output signal waveforms and their THDs are measured.
to reduce the recovery time after stimulation ends. The above measurement results have verified the function of SSAB AFE LFP acquisition unit.

The post-simulation results of recovery times in the proposed SSAB AFE LFP amplifier under the SSAB operation and the blanking method [14] are shown in Fig. 16. It can be seen from Fig. 16 that the recovery time of the proposed unit under the SSAB operation is 230 \( \mu \)S whereas that under the blanking operation [14] is as long as 1.1 S because of the long recovery time from AFE amplifier reset operation. Thus, the blanking method in [14] is not suitable for real-time closed-loop DBSs.

The measured THD values of fabricated SSAB AFE LFP amplifier under stimulation artifacts at different stimulation frequencies with a fixed pulse width of 60 \( \mu \)S and different pulse widths with a fixed stimulation frequency of 130 Hz are shown in Fig. 17(a) and Fig. 17(b), respectively. The fabricated SSAB AFE LFP amplifier can perform artifact blanking for stimulation frequencies from 25 Hz to 390 Hz at a fixed pulse width of 60 \( \mu \)S and stimulation pulse widths from 10 \( \mu \)S to 240 \( \mu \)S at a fixed stimulation frequency of 130 Hz. The worst-case THD value is 1.75%, which is acceptable in closed-loop neuro-modulation applications.

The fabricated SSAB AFE LFP acquisition unit is further tested by using the pre-recorded human LFPs mixed with stimulation artifacts. The pre-recorded human LFPs raw data are recorded by a commercial differential amplifier (Model 3000, A-M Systems, Sequim, USA). The pre-recorded human LFP raw data for 5 s. The raw LFPs \( V_{\text{LFP}} \) are mixed with stimulation artifacts generated by stimulation pulses \( V_{\text{STIM}} \) of \( \pm 3.6 \) V,130 Hz, and 60 \( \mu \)S width from 2 S to 4 S to generate LFPs with stimulation artifacts as shown in Fig. 18(b). The LFPs with stimulation artifacts were fed to the fabricated SSAB AFE LFP acquisition unit to blank the artifacts. The amplified output signals were rebuilt back to continuous signals with 4-kHz sampling rate as shown in Fig. 18(c). During the CCS stimulation duration, the large simulation voltage \( V_{\text{STIM}} \) is blanked and \( V_{\text{LFP}} \) is sampled-and-held at the input of the SSAB AFE LFP acquisition unit as shown in the zoom-in figure of Fig. 18(c). After the stimulation ends, the acquisition unit can quickly recover back to signal recording. Thus, the LFP signal linearity is not degraded too much. Fig. 18(d) shows the power spectrum of the raw LFPs, which has significant beta-band oscillations in 13 – 35 Hz as indicated by yellow shadow. Fig. 18(e) shows the power spectrum of LFPs with stimulation artifacts. Fig. 18(f) shows the power spectrum of the rebuilt LFPs at the output of SSAB AFE LFP acquisition unit when DBS is off during 0 S – 2 S and 4 S – 5 S and no stimulation artifact appears. Fig. 18(g) shows the power spectrum of the rebuilt LFPs with stimulation artifacts at the output of SSAB AFE LFP acquisition unit when DBS is on during 2 S – 4 S. It can be seen from both Figs. 18(e) and 18(g) that after the SSAB operation of the fabricated unit, the 130Hz stimulation frequency tone has an equivalent -50-dB degradation whereas the beta-band oscillations are amplified with 60-dB gain and can be clearly recovered.

The measurements results of the proposed SSAB AFE LFP amplifier and the performance comparison with the reported AFE amplifiers with artifact removal and realized in ICs or SoCs [8], [12], [14] and [15] are given in Table I. Up to \( \pm 3.6\text{-V} \) stimulation artifact voltage including both CMAV and DMAV can be totally removed with the proposed acquisition unit. With the RLD circuit, the CMRR of this work can be further boosted to be higher than 124dB, which is the highest among all others. The proposed acquisition unit achieves 150-mV EDO suppression and a large in-band input impedance of 6.87 G – 160 M\( \Omega \) that are the largest as seen in Table I. Moreover, the main advantage of proposed acquisition unit over other reported amplifiers in Table I is that it is suitable for the integration with a SoC for real-time closed-loop neuro-modulation without external controller or processor.

Fig. 19 shows the measured 4-channel output waveforms of SC-Amps in the fabricated acquisition unit with monopolar ETI measurement circuit under the operation of

**TABLE II**

The performance summary and comparison of the proposed monopolar ETI measurement circuit.

<table>
<thead>
<tr>
<th>Technology</th>
<th>This Work</th>
<th>BioCAS'18 [19]</th>
<th>BioCAS'19 [20]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supply Voltage (V)</td>
<td>1.8</td>
<td>1.8</td>
<td>1.8</td>
</tr>
<tr>
<td>Frequency (Hz)</td>
<td>125</td>
<td>1k</td>
<td>15.625</td>
</tr>
<tr>
<td>Injection Current (( \mu )A)</td>
<td>0.26</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>Modulation Type</td>
<td>Square</td>
<td>Square</td>
<td>Square</td>
</tr>
<tr>
<td>Area (mm(^2))</td>
<td>0.047</td>
<td>0.055</td>
<td>0.001</td>
</tr>
<tr>
<td>Max. Error (( % )</td>
<td>8.3% (DUT=2.4 k( \Omega ))</td>
<td>N/A</td>
<td>3.4% (DUT=46k( \Omega ))</td>
</tr>
<tr>
<td>Measured Electrode-Tissue Impedance Type</td>
<td>Monopolar (From electrode to GND)</td>
<td>Bipolar (Between two electrodes)</td>
<td>Bipolar (Between two electrodes)</td>
</tr>
</tbody>
</table>
\[ \text{IMP\_MODE} = 1 \] According to the measured waveforms, the peak-to-peak output voltages of CH1 – CH4 are 1.2473 / 1.24735 / 1.3711 / 1.2507 V, respectively. After the linear regression on the measured data and the calculation with (6), the measured monopolar impedance values of each ETI to ground are 2.186 / 2.186 / 2.572 / 2.196 kΩ, respectively, with 7% - 9% errors.

The performance summary of the proposed monopolar ETI measurement circuit and comparison with other reported bipolar ETI measurement circuits [19], [20] are given in Table II. It can be seen that the proposed circuit can measure the monopolar ETI impedance with the smallest extra power consumption of 2.65 μW.

V. CONCLUSION

In this paper, a SSAB AFE LFP acquisition unit with monopolar ETI measurement circuit and RLD circuit was designed and fabricated in 0.18-μm CMOS technology. In the proposed SSAB AFE acquisition unit, the artifact removal is achieved by blanking the first stage of AFE amplifier and holding the amplifier at its state before stimulation through a sample-and-hold operation. The measurement results of the fabricated SSAB AFE LFP acquisition unit have shown that CMAV and DMAV up to ±3.6 V under 60-μs 130-Hz biphasic CCS pulses with 150-mV EDO, can be removed through the SSAB operation and a sinewave with only 1.2% THD can be obtained. The measured total power consumption is 5.92 μW per channel and the NEF of AR-CCIA is 6.88. A shared RLD circuit was designed and integrated with the multi-channel SSAB AFE LFP acquisition unit, which can increase the CMRR to be larger than 124 dB in the passband. Moreover, a monopolar ETI measurement circuit is proposed and integrated with the acquisition unit. It has a measurement error less than 8.3% with a total 2.65-μW extra power consumption. From the measurement results, it is verified that the proposed SSAB AFE LFP acquisition unit is suitable for the integration of SoCs in real-time closed-loop DBS systems for neuro-modulation applications.

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REFERENCES


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