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An Ultra-Low Voltage Analog Front End for Strain Gauge Sensory System Application in 0.18 μm CMOS

Alexander EDWARD†(a), Nonmember and Pak Kwong CHAN(b), Member

SUMMARY This paper presents analysis and design of a new ultra-low voltage analog front end (AFE) dedicated to strain sensor applications. The AFE, designed in 0.18 μm CMOS process, features a chopper-stabilized instrumentation amplifier (IA), a balanced active MOSFET-C 2nd order low pass filter (LPF), a clock generator and a voltage booster which operate at supply voltage $(V_{dd})$ of 0.6 V. The designed IA achieves 30 dB of closed-loop gain, 101 dB of common-mode rejection ratio (CMRR) at 50 Hz, 80 dB of power-supply rejection ratio (PSRR) at 50 Hz, thermal noise floor of 53.4 nV/√Hz, current consumption of 14 μA, and noise efficiency factor (NEF) of 9.7. The high CMRR and rail-to-rail output swing capability is attributed to a new low voltage realization of the active-bootstrapped technique using a pseudo-differential gain-boosting operational transconductance amplifier (OTA) and proposed current-driven bulk (CDB) biasing technique. An output capacitor-less low-dropout regulator (LDO), with a new fast start-up LPF technique, is used to regulate this 0.6 V supply from a 0.8–1.0 V energy harvesting power source. It achieves power supply rejection (PSR) of 42 dB at frequency of 1 MHz. A cascade compensated pseudo differential amplifier is used as the filter’s building block for low power design. The filter’s single-ended-to-balanced converter is implemented using a new low voltage amplifier with two-stage common-mode cancellation. The overall AFE was simulated to have 65.6 dB of signal-to-noise ratio (SNR), total harmonic distortion (THD) of less than 0.9% for a 100 Hz sinusoidal maximum input signal, bandwidth of 2 kHz, and power consumption of 51.2 μW. Spectre RF simulations were performed to validate the design using BSIM3V3 transistor models provided by GLOBALFOUNDRIES 0.18 μm CMOS process.

key words: active-bootstrapped, analog front end, chopper-stabilized amplifier, continuous-time filter, energy harvesting, instrumentation amplifier, low-dropout regulator, low voltage, strain gauge transducer

1. Introduction

Recent interest in energy harvesting systems has sparked research in low voltage analog ICs. For example, solar cells powered systems benefit directly in terms of reliability and compactness in view of the fact that solar cells are traditionally stacked in series with a significant number to provide a high $V_{dd}$. The commonly used energy harvesting design [1] uses a boost converter to provide a local boosted $V_{dd}$ at the cost of switching noise at high frequency where most analog circuits have poor PSR. This paper, in contrast, provides a true low voltage analog solution in the design of low noise, low power, and high power supply/common-mode rejection energy harvesting sensory system that operates at only two stacked solar cells.

Conventional techniques such as cascoding and strong-inversion operation are a design concern in low $V_{dd}$ operation. Use of weak-inversion operation [2] reduces designer’s degree of freedom in optimizing amplifier’s input-stage transistors’ noise contribution. This unavoidably leads to a considerable increase of noise power spectral density (PSD). If the active load transistors are biased at the same inversion level with the input transistors, the PSD can be approximated to double compared to high $V_{dd}$ design’s noise PSD. In addition to the reduced signal swing in low $V_{dd}$ design, this forces designer to increase circuit’s current consumption so as to get satisfactory SNR.

Threshold voltage ($V_t$) reduction via forward-biased bulk [3] has been explored to increase amplifier’s input common-mode range (ICMR). This technique also allows for transistor operation in slightly higher inversion level compared to transistor with grounded bulk connection for the same gate-source voltage bias and quiescent current. Excessive forward bias, however, can reduce reliability and increase chances of turning on parasitic BJT which can couple its noise to the circuit.

With $V_{dd}$ of 0.6 V, the use of tail current source reduces the available headroom for the input and active load transistors. Tail current source-less design or pseudo-differential [4], although very attractive in low $V_{dd}$ design, has poor common-mode rejection (CMR) which is not adequate for instrumentation application. Cascading fully differential amplifiers can solve this problem at the cost of higher power consumption. Triode tail current source [5], [6], which utilizes feedback to suppress common-mode signal, can be used to retain the advantage of additional CMR capability of tail current source design. Nevertheless, the CMRR of a design using tail current source and current-mirror active load, which is approximately two-stage of gain, is still not adequate.

The objective of this work is to explore and to propose low voltage analog techniques applicable to the design of low noise instrumentation systems. Following this introduction, the AFE and its application are explained in Sect. 2.

The main highlight of the paper is a new differential difference amplifier (DDA) using the active-bootstrapped technique [7]. The OTA, which is the core building block of the technique, is implemented using the pseudo-differential technique suitable for low $V_{dd}$ operation. New low voltage CDB bias circuit is also proposed to reduce the noise contribution of the input stage’s active load transistors. Mathematical analysis, qualitative circuit description, and simulation

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results of the technique are discussed on Sect. 3.

Other supporting building blocks of the AFE: LDO, MOSFET-C filter, clock generator and voltage booster circuitry are qualitatively described in Sects. 4, 5, and 6 respectively including methods to reduce bandgap’s PSR contribution in an LDO and to improve CMRR in a pseudodifferential two-stage amplifier. The overall system’s simulation results and conclusion are given in Sects. 7 and 8.

2. Analog Front End Architecture & Specifications

An overview of the AFE is given in Fig. 1. Differential input signal from external strain gauge is amplified by a single-ended-output chopper-stabilized IA, splitted by a single-ended-to-balanced converter, and filtered by a balanced active MOSFET-C filter. An LDO regulates both the internal $V_{dd}$ of 0.6 V and the strain gauge’s bias voltage. An external 32.768 kHz tuning fork crystal, requiring two external pins, is used to generate a chopper clock for the IA. This clock is boosted by a two-stage charge pump to generate a tripled clock with swing of approximately 1.8 V. The charge pump also provides a stable 1.8 V voltage to bias the filter’s MOSFET resistors.

Strain gauge connected in wheatstone bridge configuration, shown in Fig. 1, yields a linear relationship between its output voltage and its sensed strain given by

$$V_{in} = \frac{\Delta R}{R} \frac{R}{R + R_{bias}} V_{dd}$$

(1)

where $R/(R + R_{bias})V_{dd}$ is the strain gauge’s bias voltage $V_{bias,s}$ and the change of resistance $\Delta R$ is related to the sensed strain by

$$\Delta R = \epsilon GF$$

(2)

where $\epsilon$ is the sensed strain and GF is the strain gauge’s gauge factor, that is pertaining to its sensitivity.

The differential output noise PSD of this configuration is simply $4kT R$. $R_{bias}$, which only contributes common-mode noise, influences input SNR since it sets the maximum signal’s magnitude as shown in Eq. (1).

In this design, 1 kΩ strain gauge is used. With $V_{bias,s}$ of 0.3 V, maximum strain of 30,000 $\mu$ and GF of 2, it consumes 300 $\mu$A of supply current, produces 0.18 mV of maximum output signal, and exhibits differential output noise voltage spectral density (VSD) of 4.07 nV/$\sqrt{Hz}$. Even when biased at low voltage, the strain gauge’s noise is not the limiting factor to the overall system’s SNR.

Table 1 shows some of the typical strain gauge’s parameters [8]. Mismatch in strain gauge’s nominal resistance creates unavoidable DC offset at the IA’s inputs. For example, choosing strain gauge with ±0.15% mismatch will yield 540 mV peak-to-peak output signal and 48 mV output offset for IA gain of 29.54 dB (30 V/V) which is marginally adequate to prevent output signal clipping. In this design, offset calibration is implemented by trimming the resistive voltage divider that biases the DDA’s $V_{−}$ terminal. The detailed implementation, similar to gain error calibration, is discussed briefly in the next section.

<table>
<thead>
<tr>
<th>Table 1 Omega SGD series strain gauge parameters.</th>
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</tr>
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<td>Gauge Factor</td>
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<tr>
<td>TC (Steel) (Ref=23°C)</td>
</tr>
<tr>
<td>Operating Temp.</td>
</tr>
<tr>
<td>Max. Strain</td>
</tr>
</tbody>
</table>

![Fig. 1 AFE’s block diagram.](image-url)
3. Active-Bootstrapped Chopper-Stabilized Instrumentation Amplifier

DDA [9] is an excellent choice to implement low voltage IA with reasonable power consumption. Compared to the three op-amp IA, the CMRR in a DDA-based IA does not depend on resistor matching. The current feedback IA [10], [11], although has excellent NEF, is not suitable for low $V_{dd}$ operation.

The output voltage of a DDA-based IA, shown in Fig. 2, is given by

$$V_{out} = \frac{A_{ol}}{1 + A_{ol}\beta} (V_{in} + V_{ox} + V_{bias}) + \frac{V_{in}}{2 \beta \text{CMRR}_{d}} (3)$$

where $A_{ol}$ is the DDA’s open loop gain, $\beta = R_{f2}/(R_{f1} + R_{f2})$ is the feedback factor, $V_{in}$ is the strain gauge’s output voltage, $V_{ox}$ is the combined DDA’s and strain gauge’s offset voltage, $V_{bias}$ is the bias voltage applied to DDA’s $V_{+}$ terminal to set the DC output voltage at half $V_{dd}$, and $\text{CMRR}_{d}$ is the differential-common-mode rejection ratio which quantifies rejection from the $(V_{+} - V_{-})/(V_{+} - V_{-})/2$ signal.

A DDA has two sources of gain error, the finite open loop gain $A_{ol}$ and the finite differential common-mode gain $\text{CMRR}_{d}$. While the error caused by finite $A_{ol}$ is dominant when the chosen closed-loop gain $1/\beta$ is high, the error caused by finite $\text{CMRR}_{d}$ is constant regardless of the closed-loop gain. Finite $\text{CMRR}_{d}$ is mainly caused by $g_{m}$ mismatch in the DDA’s input transistors. This in turn is caused by the tail current source mismatch. As they are biased in weak inversion, the matching is poor and unavoidable in low $V_{dd}$ operation.

Gain error can be digitally calibrated by implementing $R_{f1}$ or $R_{f2}$ using series resistor strings and switches. Monte Carlo simulations were done to measure the initial gain error, $\text{Gain}$, when the chosen closed-loop gain $1/\beta$ is constant regardless of the gains error. Assumed high impedance tail current source, one can rearrange all the small-signal node equations of the circuit in Fig. 3 (feedback network not shown) to obtain the signal flow graph illustrated in Fig. 4. $g_{m1}$, $g_{m2}$, and $g_{m3}$ are the transconductances of the input transistors, active loads transistors, and output-stage transistor respectively. $g_{mdm}$ and $g_{mcm}$ are the OTA’s differential-mode and common-mode transconductances. $R_{1} = (r_{oa})/2 || r_{o2}$ is the input-stage output resistance without active-bootstrap technique and $R_{2} = (r_{o3}) || r_{o4} || (R_{f1} + R_{f2})$ is the output resistance of the output stage.

By applying the Mason’s gain formula [13] to the signal flow graph shown in Fig. 4, one can see that the active-bootstrapped input stage is a three-stage amplifier. There are three feedforward paths from the inputs to the output: (i) $M_{b1}$ common-source, (ii) $M_{b1}$ common-source $\rightarrow$ OTA’s and supply rail respectively. The bias voltage can be easily obtained using a voltage divider from $V_{dd}$ which is shown in the offset trim block in Fig. 1.
inverting terminal inverting M2b, common-source, and (ii) M1x.c.
common-source non-inverting terminal common-source. Due to matching property, the positive and negative feedback loop gains, as indicated in Fig. 4, cancel each other. This leaves the three-stage feedforward gain paths intact.

By adding a common-source amplifier as an output stage, the rail-to-rail output swing capability is obtained. Since the IA is not designed for unity gain, Miller compensation is adequate in this design without any significant power drain.

The closed-loop transfer function of the active-bootstrapped DDA based IA is given by

$$A_c(s) = \frac{A_{ol}(s)}{1 + A_{ol}B_1 + \frac{p_1}{p_2p_3} + \frac{p_2}{p_1p_2p_3}}$$

where the open loop gain A_{ol} is given by

$$A_{ol} = (g_{m1}R_1g_{mdm}/g_{ncm})g_{m2}R_2$$

and the input-stage output resistance R1 is boosted by the OTA’s CMRR g_{mdm}/g_{ncm}. In this design, the gain-boosting OTA is realized using pseudo-differential technique. Therefore, one can assume that OTA’s common-mode gain is in the order of unity. In the case that OTA’s CMRR is high, the input-stage output resistance is boosted by factor of g_{mdm}R_{ota}g_{m2}R_1.

The zeros are given by

$$z_1 = -\frac{g_{m3}}{C_c}$$

$$z_2 = \frac{g_{mdm}}{C_{gd2a}} \left| \frac{2g_{m2}}{g_{cd2a} + g_{cd2b}} \right|; \ z_2z_3 = \frac{2g_{m2}g_{mdm}}{Y}$$

and the poles are given by

$$p_1 = \left( \frac{\beta g_{m1}1}{C_c} - \frac{g_{m3}}{C_c} \right)$$

$$p_1p_2 = \frac{\beta g_{m1}}{C_c} \left| \frac{g_{mdm}}{X/C_c} - \frac{g_{m3}g_{mdm} - g_{m2}}{g_{cd2a}} \right|$$

$$p_1p_2p_3 = \frac{\beta g_{m1}}{C_c} \left| \frac{g_{mdm}g_{mdm}}{Z/C_c} \right|\ \left| \frac{g_{m2}g_{mdm}}{X/C_c} \right|$$

$$p_1p_2p_3p_4 = \frac{\beta g_{m1}g_{m2}g_{mdm}}{XZ}$$

where

$$X = C_1C_2 + C_1C_1 + C_2C_1; \ Y = (C_L + C_2)C_2$$

$$Z = C_{ota}C_2 + C_{ota}C_{gd2a} + C_2C_{gd2a} + C_2C_{gd2b}$$

The following assumptions are used when obtaining the above expressions: (i) high single-stage amplifier gain (40 dB), (ii) small OTA’s differential transconductance, and (iii) $C_{gd2a} \approx C_{gd2b}$ is small.

Note that the underlined terms are pole-zero locations of a two-stage Miller compensated amplifier in closed-loop configuration. The modified pole-zero locations attributed by this technique will now be discussed.

Equation (10) indicates that RHP poles exist caused by the positive feedback loop. This effect can be minimized by ensuring the capacitive loading for the positive feedback loop is higher than that of negative feedback loop ($C_1 > C_2$). Very crude approximation for $p_2$ and $p_3$ are: $g_{m3}/C_L$ and $g_{mdm}/g_{ncm}/(C_{ota}C_2/C_c)$ respectively. This indicates potential peaking ($p_3 < p_2$) due to the small value of OTA’s transconductance and the large size of $C_{ota}$ (long channel device is used in $M_2$). Another design rule-of-thumb is to ensure $g_{mdm} || - g_{m2}$ to be positive to reduce the potential presence of RHP pole.

3.2 Noise Analysis

The input-referred noise PSD of the DDA is given by

$$\frac{\nu^2_{DDA}}{\Delta f} = 4kT \frac{\gamma}{\alpha} \frac{g_{m2}}{g_{m1}} \left( 4 + \frac{g_{m2}}{g_{m1}} \right)$$

$$+ \frac{\nu^2_{OTA}}{\Delta f} \left| \frac{1}{g_{m1}} \left[ 1 + \frac{2z}{p_1} \left( 1 + \frac{z}{p_2} \right) \right]^2 \right|$$

with zeros and poles given by

$$z_1 = \frac{1}{R_1(C_2 + C_{gd2a})}; \ z_2 = \frac{g_{m2}}{C_{gd2b}}$$

$$p_1 = \frac{g_{mdm}}{C_{gd2a}} \left| \frac{-2g_{m2}}{C_{gd2b} + C_{gd2b}} \right|; \ p_1p_2 = \frac{2g_{m2}g_{mdm}}{Z} \left| \frac{1}{g_{mdm}g_{mdm}} \right|\ \left| \frac{g_{m2}g_{mdm}}{X/Z} \right|$$

$$p_1p_2p_3p_4 = \frac{\beta g_{m1}g_{m2}g_{mdm}}{XZ}$$

where $\gamma/\alpha$ is a constant and is approximately 2/3 in long channel MOSFET.

For good CMRR, or gain error performance, both differential pairs are usually matched. This increases DDA’s input-referred noise VSD by a factor of $\sqrt{2}$ compared to that of an op amp.

The gain-boosting OTA contributes little to the overall DDA’s noise performance. It can be deduced from Fig. 5, OTA’s input noise voltage is only amplified twice whereas the input signal has to go through three stages of amplification. By the same reasoning, the OTA’s offset voltage can also be considered to be negligible.

3.3 Detailed Circuit Description

The complete schematic of the proposed DDA is shown in
Fig. 6. The Dynamic Threshold MOS (DTMOS) [14], [15] devices are used as the input transistors $M_{1a}$, $M_{1b}$, $M_{1c}$, and $M_{1d}$ for the DDA. The DTMOS is essentially a MOS transistor with shorted gate and bulk connection. As a result, the bulk transconductance ($g_{mb}$) directly adds to the gate transconductance ($g_m$), reducing noise PSD by approximately 20% ($g_{mb}/g_m \approx 0.2$).

To increase ICMR, the tail current sources $M_{8a}$ and $M_{8b}$ that can work in the triode region [5] are used in the design. As long as $M_{7a}$, $M_{7b}$, and $M_{8a}$ remain in saturation, a constant current is forced to flow in $M_{6a}$ and copied by $M_{5a}$ to bias the input transistors. Transistors $M_{7a}$ and $M_{7b}$ replicate the input transistors and sense the common-mode signal. This common-mode signal is amplified by $M_{6a}$ and fed back to $M_{7a}$ through its non-inverting terminal. The resulting negative feedback suppresses the common-mode signal. The same also goes for $M_{8b}$, $M_{9b}$, $M_{10}$, $M_{11}$, and $M_{12}$.

Small-signal analysis shows that the common-mode gain of an active-bootstrapped DDA using the feedback tail current source is given by

$$A_{cmDDA} = \frac{1}{2g_{mcm}R_{ota}} \left( 1 - \frac{g_m}{g_{m2f}} - \frac{g_m}{g_{m2f}g_{m3f}} \right) \quad (18)$$

The common-mode rejection is composed of factors $g_{m2f}$ associated with “folded” tail current sources $M_{6a}$ and $M_{8a}$ and $g_{m3f}$ associated with feedback’s loop gain. However, in this design the circuit is operated with $M_{5a}$, $M_{5b}$, $M_{9a}$, and $M_{9b}$ in triode region and one can only get a modest common-mode rejection of around $g_{m2f}$. Neglecting mismatch (assume $g_{mcm}R_{ota}$ around unity since pseudo-differential topology is used). Combining this and Eq. (6), one gets CMRR of approximately $2g_{m1}R_{1}(g_{m2f}R_{ota}g_{m2f}R_{ota})$ in the ideal case.

The unique feature of the proposed IA is the adoption of pseudo-differential technique in the design of gain-boosting OTA. This shows that the active-bootstrapped technique can also be used in $V_{dd}+V_{sat}$ supply operation. Besides boosting CMRR, this approach allows interfacing between the active-bootstrapped input stage and the PMOS common-source output-stage. One difficulty in using this approach is in the start-up phase of the amplifier since the bias current of pseudo-differential circuit depends on the input voltage. During start-up phase, the voltage at $M_{2a}$ and $M_{2b}$’s drain terminals are near $V_{dd}$ because of higher parasitic capacitance from there to $V_{dd}$ ($C_{gs}$ of $M_{3}$) than that to ground. In that condition, the PMOS input transistors of the gain-boosting OTA $M_{3a}$ and $M_{3b}$ are turned off. No current will flow through $M_{12a}$ and $M_{12b}$, and the bias currents from $M_{13a}$ and $M_{13b}$ will bias the OTA’s active loads $M_{13a}$ and $M_{13b}$. This in turn biases the main amplifier’s active loads $M_{2a}$ and $M_{2b}$ to discharge their parasitic drain capacitance. This large signal start-up loop is stable because of the positive sign of the OTA’s common-mode gain. Conventional OTA, however, is biased by tail current source which automatically biases the active load of the main amplifier. In that case, no start-up problem occurs.

To reduce noise contributed by the active loads $M_{2a}$ and $M_{2b}$, they are operated in moderate-inversion using the current-driven bulk technique. Although strong-inversion operation is preferred for lower noise, $M_{2a}$ and $M_{2b}$’s headroom is fixed by $V_{dd} - V_{sat}$. Therefore, their output resistance will drop and noise from the gain-boosting OTA starts to be noticeable when they are operated at strong-inversion (see Eq. (15)). The $\beta$ insensitive bias circuit in [3] is not used in this design due to limited headroom. As an alternative, a new feedback-based bias circuit is proposed.

In the proposed CDB biasing circuit, a reference transistor $M_{18}$ is biased to have the desired $V_{gs} \approx 0.45$ V for example. $M_{19}$ is designed to have lower aspect ratio than $M_{18}$, which implies higher inversion level or lower $g_m/I_d$. To

![Complete schematic of the proposed active-bootstrapped chopper-stabilized DDA.](image-url)
<table>
<thead>
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<th>Parameter</th>
<th>[18]</th>
<th>[10]</th>
<th>[19]</th>
<th>[11]</th>
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</tbody>
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### Fig. 7
DDA’s open and closed loop transfer function.

match both transistors’ \(V_{fsr}\), an amplifier formed by \(M_{20a}, M_{20b}, M_{31a},\) and \(M_{31b}\) is used to control the bias transistor \(M_{22}\). As a result, transistor with a higher inversion level can be designed using this technique.

To drive the relatively low 200 kΩ feedback resistors, a power efficient pseudo class AB output stage [12] is used. The cut-off operated transistor \(M_{15}\) implements a very large resistor to block the DC biasing whereas a coupling capacitor \(C_{bat}\) makes this configuration looks like a class AB amplifier during AC condition.

The second-stage is formed by a PMOS common-source amplifier. This peculiar arrangement limits ICMR if this amplifier is operated at high \(V_{dd}\). However, this is not of a concern in a fixed low \(V_{dd}\) operation. The ICMR of the proposed topology is given by

\[
\text{ICMR}_{\text{min}} = V_{dd} - V_{sg3} + V_{dsat1} - V_{sg1} \tag{19}
\]

\[
\text{ICMR}_{\text{max}} = V_{dd} - V_{dsat5} - V_{sg1} \tag{20}
\]

\[
\text{ICMR} = V_{sg3} - V_{sat1} - V_{dsat5} \tag{21}
\]

which indicates that ICMR is fixed regardless of \(V_{dd}\).

The chopper-stabilization technique [16] can be used in conjunction with the active-bootstrapped technique to reduce offset and flicker noise of the DDA. Since the differential-to-single-ended conversion is implemented in the input-stage, the modulator placement in [17] is used. Chopper-stabilization is also applied to the gain-boosting OTA to remove its offset and flicker noise.

### 3.4 DDA Simulation Results

The simulated open-loop transfer function of the DDA is given in Fig. 7. The open-loop gain and GBW are 123 dB and 500 kHz respectively for 50 pF in parallel with 200 kΩ load. Since the DDA is dedicated to provide the gain function and it will not operate at 0 dB of gain, one can save power by biasing the second-stage just enough to drive the feedback resistors, which are constrained by the noise requirements. At a gain of 29.54 dB, the phase margin is 85°. Other DDA performance parameters are shown in Figs. 8–11. Table 2 shows that the DDA ultra-low voltage performances are comparable with the reported works.

To verify the agreement between theoretical analysis and simulation results, Spectre pole-zero analysis was performed for the DDA under closed-loop condition. The simulated pole-zero locations, ignoring the doublets, were compared with Eqs. (5)–(12) using parameters estimated from Spectre DC operating point simulation. Shown in Table 3,
the theoretical analysis predicts the locations of poles and zeros with reasonable accuracy when considering the effects of gain-boosting OTA’s internal poles and zeros, β frequency degradation caused by finite DDA’s input capacitance, and complete quasi static capacitance modelling used by the simulator which includes the effects of $C_{dd}$.

In addition to the well known RHP zero caused by Miller compensation, this technique adds two complex zeros which is responsible for the peaking of the noise response in Fig. 8 at frequency around 400 kHz. The presence of the complex poles is harder to be seen since the complex poles and zeros damped each other in the closed-loop response, but nevertheless slight peaking is observed around 300 kHz in Fig. 8 followed by a notch caused by the complex zeros. Together, the theory predicts $-270^\circ$ of phase at high frequency after taking into account 1 RHP zero, 2 complex zeros, 2 complex poles, and 2 LHP poles which is true at frequency around the RHP zero (846.8 kHz which is at the highest frequency). The complete simulation, however, predicts more poles and zeros than that of the prediction of the theory. This is why the closed-loop frequency response goes beyond $-270^\circ$ of phase.

Transient response shown in Fig. 12 verifies that the design is stable. Note that the transient simulation was performed for the whole AFE and the IA is loaded by the filter in this case. The strain-gauge is modelled by resistors and some voltage controlled voltage sources to emulate the resistance changes (the input signal). The input signal glitches shown in Fig. 12 is caused by the input chopper modulator’s finite non-overlapping clock and finite strain gauge’s input resistances.

The noise responses of the DDA with and without chopper-stabilization simulated with PSS+PNOISE analyses are shown in Fig. 8. Chopper-stabilization reduces flicker noise from 365 nV/√Hz to 53.7 nV/√Hz at 1 Hz at the cost of peaking at the chopper frequency of 32.768 kHz. Peaking predicted by Eq. (15) can also be seen in the noise response.

With 14 μA current consumption (without bias circuit) and integrated input noise voltage of 537.2 nVrms (0.5–100 Hz or 64 Hz noise bandwidth), the NEF is calculated to be 9.7. For a differential pair operated in weak inversion, its theoretical minimum NEF is 2.9 [18]. Since a DDA has four input transistors, its theoretical minimum NEF is 5.8. If the active load transistors have the same noise contribution with that of the input transistors, this NEF again degrades with $\sqrt{2}$ to 8.2. The gain-boosting OTA, output stage, and feedback resistors consume a total of 4 μA. When compared with 10 μA input-stage current consumption, this additional current consumption degrades NEF by $\sqrt{1.4}$ to 9.7. This degradation of NEF is tolerable in view of low $V_{dd}$ operation.

The DDA’s open loop gain plotted against common-mode input voltage is shown in Fig. 9. The input transistors stay in saturation up to common-mode voltage of 0.2 V, which is more than sufficient in this application.

Frequency response of CMRR and PSRR is shown on Fig. 10. Note that the AC analysis overestimates low frequency CMRR based on assumption of perfect matching. In order to estimate the realistic CMRR performance metric, Monte Carlo simulations were conducted.

The 3σ Monte Carlo simulation results for DC open-loop gain, CMRR, PSRR, gain error, and offset voltage without chopper are 114 dB, 101 dB at 50 Hz, 80 dB at 50 Hz, 2%, and 0.4 mV respectively. CMRR can be expected to get better with the use of chopper-stabilization.
4. Composite Cascode Low-Dropout Regulator

The complete schematic of the LDO is given in Fig. 13. The core LDO is composed by a pass transistor $M_1$, feedback resistors $R_{I1}$ and $R_{I2}$, and a pseudo-differential OTA formed by $M_{2a}$, $M_{2b}$, $M_{3a}$, $M_{3b}$, $M_{4a}$, and $M_{4b}$. Composite cascode devices $M_{2a}$, $M_{2b}$, $M_{3a}$, and $M_{3b}$ are used as the OTA’s input transistors to create a low impedance node for cascode compensation. In this design, both transistors $M_{2a}$ and $M_{3a}$ are operated in saturation region [24]. This requires $M_{3a}$ to have larger aspect ratio than that of $M_{2a}$. For $M_{2a}$ to have gain of 20, for example, $M_{3a}$ should have 4 times the aspect ratio of $M_{2a}$ neglecting body effect. At this operating condition, $M_{2a}$ consumes only 60 mV of headroom. The same also goes for $M_{2b}$ and $M_{3b}$.

Besides stability at no-load condition, cascode compensation has good PSR response [25]. Nevertheless, since the LDO output is referred to the reference voltage, the bandgap’s supply noise can be dominant at low frequency.

To eliminate bandgap’s PSR contribution, an RC LPF can be constructed using MOSFET $M_{14}$ in cut-off region to implement a very high resistor [12]. Passive RC filtering is the proper choice to eliminate noise and to increase supply rejection of the bandgap in low power operation. However, the large time-constant required for the filtering implies slow start-up.

A new start-up circuit can be constructed whose function is to change $M_{14}$’s on-resistance during start-up. A key consideration for the design is not to let any parasitic capacitance connects the filter output capacitor $C_h$ to the supply rail as this degrades PSR. This is implemented by common-source amplifier $M_{17}$ loaded by diode connected transistor $M_{18}$ which senses the filter’s output voltage.

During start-up, the output voltage of $M_{17}$ is high and the inverter $M_{19}$ and $M_{20}$ is low. This will turn on and turn off transistor $M_{16}$ and $M_{15}$ respectively. The gate terminal of $M_{14}$ is effectively connected to $V_{dd}$, hence it is operated in triode region. During normal operation, $M_{16}$ and $M_{15}$ are off and on respectively. The gate terminal of $M_{14}$ is effectively connected to the bandgap’s output, hence it is operated in cut-off region.

In this design, the current-mode bandgap in [26] is used. Subthreshold-operated MOSFETs $M_5$ and $M_6$ are used to replace BJTs. Pseudo-differential OTA composed by $M_{8a}$, $M_{8b}$, $M_{9a}$, and $M_{9b}$ forms current mirrors with $M_5$ and $M_6$. This arrangement [27] eliminates biasing circuitry for the OTA.

The LDO was designed to provide maximum current of 500 $\mu$A, in which 300 $\mu$A is supplied to the strain gauge. With the addition of 100 $\mu$F output capacitor, the LDO is stable at no-load condition with phase margin of 60°. The LDO achieves 42 dB of PSR at a frequency of 1 MHz. The PSR response, shown in Fig. 14 shows low-frequency pole caused by the filter. The LDO consumes quiescent current of 5.58 $\mu$A, in which 3 $\mu$A is consumed by the bandgap and filter’s start-up circuit. Table 4 summarizes the overall LDO performance.

5. Balanced Active MOSFET-C Filter

A balanced MOSFET-C Tow Thomas Biquad, shown in Fig. 15, is used to implement a 2nd order LPF with 2 kHz of cut-off frequency. Active RC filter has the capability to process rail-to-rail input signal with good dynamic range, suitable for low $V_{dd}$ application. Balanced topology [28] is used to cancel even-order of non-linearity and noise from both $V_{dd}$ and $V_{ss}$. MOSFET resistors are used since a DC voltage is available from the charge-pump. With approxi-
approximately 1.8 V gate bias and 540 mV input signal, this scheme will yield resistors linear enough for this application. Tuning is not implemented in this design since variations in filter’s cut-off frequency will not affect SNR considerably.

As a building block for the filter, a pseudo-differential amplifier with common-mode feedforward (CMFF) [4] shown in Fig. 16 is used in this design. The amplifier uses composite cascode devices as the input-stage’s active loads and output-stage common-source amplifiers to implement a two-stage cascode compensated amplifier. In contrast with the composite cascode used in the LDO, the bottom transistors $M_{8a}$, $M_{8b}$, $M_{9a}$, $M_{9b}$, and $M_{10b}$ are operated in triode region. This yields slight improvement in GBW [29] by factor of $1 + g_{ds3}/g_{m2}$. However, the location of complex poles and peaking’s magnitude is lowered and increased by square root of the same factor respectively. This method reduces amplifier’s output-stage current consumption with stability maintained. Additional capacitors can be added at the amplifier’s outputs and gate terminals of $M_{8a}$, $M_{8b}$, $M_{9a}$, $M_{9b}$, and $M_{10b}$ to reduce peaking without any significant change in the biquad’s transfer function.

To implement single-ended-to-balanced conversion, a unity gain inverting amplifier is used. The amplifier, shown in Fig. 17, implements a two-stage common-mode cancellation to eliminate biasing for the output-stage current source. $M_{3a}$, $M_{3b}$, $M_{4a}$, and $M_{4b}$ sense the common-mode signal. This signal is inverted by $M_{5}$ and $M_{6}$ and amplified by $M_{7}$ to cancel the common-mode signal amplified by $M_{8}$. In contrast with its fully differential CMFF technique counterpart, this scheme does not require the CMFF circuitry to consume the same current as that of the input-stage for optimum cancellation. This technique can also be used to improve CMRR in an op amp with tail current source for non-inverting amplifier application.

Both amplifiers were designed for 0.3 V input common-mode voltage. To avoid overly large transistor associated with subthreshold operation, the proposed current-driven bulk bias circuit described in Sect. 3.3 is used. The amplifiers’ bias current is chosen to minimize filter’s output noise PSD after the filter’s cut-off frequency. The fully-differential amplifier consumes 14.3 $\mu$A in which the input-stage consumes 6.2 $\mu$A, the CMFF circuitry consumes 6.2 $\mu$A, the output-stage consumes 0.67 $\mu$A, and biasing consumes 0.1 $\mu$A. The single-ended output amplifier consumes 13.4 $\mu$A in which the input-stage+CMFF circuitry consumes 7.1 $\mu$A and the output-stage consumes 6 $\mu$A.

6. Clock Generator

A two-pin crystal oscillator in [2] generates 32.768 kHz chopper clock signal. This clock output is tripped by a two-stage charge pump in [30]. The charge pump also provides a stable DC output voltage to bias filter’s MOSFET resistors. Since the charge pump only drives transistors’ gate, its power consumption is negligible.

7. System Simulation Results and Discussions

The overall transfer function of the AFE, which is dominated by the filter, is shown in Fig. 18. The output noise VSDs with and without the chopper-stabilization of the IA are depicted in Fig. 19. The output noise VSDs of the AFE with chopper-stabilization are 3.917 $\mu$V/$\sqrt{\text{Hz}}$ at 1 Hz, 3.229 $\mu$V/$\sqrt{\text{Hz}}$ at 1 kHz and 146.7 nV/$\sqrt{\text{Hz}}$ at 10 kHz. The integrated output noise of the AFE from 1 Hz–1 MHz was simulated to be 171.7 $\mu$V. With differential output signal of 1.08 V peak-to-peak (0.382 Vrms), this corresponds to 67 dB SNR or sensitivity of 13.4 $\mu$Strain.

To verify the robustness of the design, Monte Carlo simulations were performed on the AFE. The average SNR, shown in Fig. 20, was found to be 65.6 dB. This slight deviation from the ideal condition can be attributed to gain error and mismatch which cause incomplete cancellation of $V_{bias}$’s and $V_{ld}$’s noise from the LDO. The calculation of SNR takes into account of the $V_{ld}$ spread caused by the process variations in bandgap’s output voltage. $3\sigma$ variation in $V_{ld}$ ranges from 0.54 to 0.66 V. Unfortunately, Monte Carlo process model does not include capacitor’s variation model. Nevertheless, even increase in filter’s cut-off frequency by $\pm50\%$ can be estimated to reduce SNR by only 1.76 dB.

The transient response of the AFE for a 2 kHz square-wave 18 mV peak-to-peak input signal is shown in Fig. 21.
The THD for 18 mV 100 Hz sinusoidal input signal was simulated using transient analysis and found to be 0.837%.

Comparison with the published AFEs dedicated to strain-gauge monitoring in Table 5 shows that the proposed design achieves low power operation without sacrificing noise performance. Although this design does not feature an ADC or RF transmitter, the power consumption of the analog signal conditioning circuitry (IA and filter) is much less than the previously published designs.

### 8. Conclusion

This work presents a new low voltage AFE for strain-gauge sensory system. A new active-bootstrapped IA with pseudo-differential gain boosting OTA and CDB biasing technique is proposed to achieve high CMRR, low noise, and rail-to-rail output swing capability with 0.6 V supply voltage. A low-voltage LDO with the proposed filtering technique is useful to improve supply rejection and noise performance of the overall system. The design of a balanced active MOSFET-C filter alongside with pseudo-differential amplifier as its building block is presented for low voltage filtering applications. A single-ended-output amplifier with a new two-stage common-mode cancellation technique is used to implement single-ended-to-balanced conversion. The design techniques presented here are also applicable to the design of other types of low noise, low voltage sensory systems.

### References


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