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Conv-SRAM: An Energy-Efficient SRAM with In-Memory Dot-product Computation for Low-Power Convolutional Neural Networks

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Abstract—This work presents an energy-efficient SRAM with embedded dot-product computation capability, for binary-weight convolutional neural networks. A 10T bit-cell based SRAM array is used to store the 1-b filter weights. The array implements dot-product as a weighted average of the bit-line voltages, which are proportional to the digital input values. Local integrating ADCs compute the digital convolution outputs, corresponding to each filter. We have successfully demonstrated functionality (> 98% accuracy) with the 10,000 test images in the MNIST hand-written digit recognition dataset, using 6-b inputs(outputs). Compared to conventional full-digital implementations using small bit-widths, we achieve similar or better energy-efficiency, by reducing data transfer, due to the highly parallel in-memory analog computations.

Index Terms—in-memory computation, dot-product, energy-efficient SRAM, machine learning, edge-computing, analog computing, convolutional neural networks, binary weights

I. INTRODUCTION

Artificial intelligence (AI) and machine learning (ML) are changing the way we interact with the world around us. Speech recognition [1] allows us to interact with “smart” devices more naturally using our voices. Facial recognition [2] enables using our faces to get access to devices in a more intuitive manner, instead of traditional passcodes. As we think about extending machine intelligence to more and more devices around us, in the “Internet-of-Things” (IoT), “edge-computing” i.e. computing on these edge devices vs. the “cloud” becomes increasingly important. There are a multitude of reasons for this. Firstly, “edge-computing” enables the devices to make fast decisions locally, without having to wait for the “cloud”. Secondly, it can significantly reduce the communication traffic to the “cloud”, by only sending the critical/relevant information and filtering out the rest of the massive amount of data the edge-devices may collect. Furthermore, “edge-computing” helps in improving the security of the data by keeping it local (within the devices), rather than having to send sensitive information to the “cloud”. While “edge-computing” promises significant benefits for IoT devices, it also has certain requirements. The circuits to run the compute algorithms must be very energy-efficient to extend the battery-life of these IoT devices, most of which have a very limited energy budget and would be “always-ON”. Additionally, in many applications, the local decision-making has to be done in real-time (e.g. self-driving cars), to make them practical.

Convolutional neural networks (CNN) provide state-of-the-art results in a wide variety of AI/ML applications, ranging from image classification [3] to speech recognition [1]. However, they are highly computation-intensive and require huge amounts of storage. Hence, they consume a lot of energy when implemented in hardware and are not suitable for energy-constrained applications e.g. “edge-computing”.

CNNs typically consist of a cascade of convolutional (CONV) and fully-connected (FC) layers (Fig. 1) [4], [5], with some non-linear layers in between (not shown in the figure). The CONV layers extract different features of the input and the FC layers combine these features to finally assign the input to one of the many pre-determined output classes. For each of the CONV/FC layers, there is a set of 3-D filters \( W_k \), which are applied on the 3-D input feature-map (IFMP) to that layer and generate its 3-D output feature-map (OFMP). Each 3-D filter/input consists of multiple 2-D arrays, each of which corresponds to a different “channel” (1 to \( C \)). When a 3-D filter \( W_k \) is applied on the input \( X \), an element-wise multiplication is performed, followed by an addition of the partial products to compute the convolution output \( Y_k \). For CONV layers the 3-D filter is applied on the shifted input to compute the next element in the 2-D OFMP. Each individual filter corresponds to a different channel in the 3-D OFMP. Therefore, the fundamental operation for both the CONV and FC layers can be simplified to a dot-product or a multiply-and-accumulate (MAC) operation, as shown in equation (1).
Embedding computation inside memory has two significant benefits. Firstly, data transfer to/from the memory is greatly reduced, since the filter-weights are not explicitly read and only the computed output is sent outside the memory. Secondly, we can take advantage of the massively parallel nature of CNNs to access multiple memory addresses simultaneously. This is because we are only interested in the result of the computation using the memory data and not the individual stored bits. Therefore, a much higher memory bandwidth can be achieved with this approach, overcoming some of the major limitations posed by the conventional “von-Neumann bottleneck”.

This paper is organized as follows. Section II explains the concept of memory-embedded convolution computation as voltage averaging in SRAMs. Section III presents the overall architecture. Section IV highlights the key contributions of this work, compared to prior in-memory computing approaches. Section V discusses the details of the different circuitry involved in the embedded convolution computation. Section VI presents the measurement results. Finally, concluding remarks are discussed in section VII.

II. CONCEPT OF SRAM-EMBEDDED COMPUTATION

The basic operation involved in evaluating convolutions ($Y$) for CNNs is the dot-product of the 3-D IFMP ($X$) and the filter-weights ($W$), as shown in equation (1). It can be rewritten by flattening the 3-D tensor into a 1-D vector to obtain equation (2), where the 2-D subscripts ($x, y$) have been omitted for simplicity.

$$Y_{x,y,k} = \sum_{c=1}^{C} \sum_{j=1}^{R} \sum_{i=1}^{R} W_{i,j,c,k} \times X_{x+i,y+j,c}$$

(1)

$$Y_k = \sum_{i=1}^{R \times R \times C} W_{k,i} \times X_i$$

(2)

Equation (2) can be further simplified for the case of binary filter-weights ($w_i$’s) to get equation (3a), where $\alpha_k$ is the common coefficient for the $k^{th}$ filter. If $\alpha_k$ is expressed as a ratio of two integers ($M_k/N$: number of elements added per dot-product in 1 clock cycle), then we get equation (3b).

$$Y_k = \alpha_k \sum_{i=1}^{R \times R \times C} w_{k,i} \times X_i, \quad w_i \in (+1, -1)$$

(3a)

$$Y_k = \frac{M_k}{N} \sum_{i=1}^{R \times R \times C} w_{k,i} \times X_i, \quad M_k, N \in I^+$$

(3b)

Now, if we separate out the scaling factor of $M_k$ (which can be incorporated after computing the entire dot-product), we get the expression for the effective convolution output ($Y_{OUT}$) as:

$$Y_{OUT,k} = \frac{1}{N} \sum_{i=1}^{R \times R \times C} w_{k,i} \times X_{IN,i}$$

(4)

where $X_{IN}$ is the effective convolution input, i.e. scaled version of the original input $X$, limited to 7-b (includes 1-b sign). For energy-efficient computation with multi-bit values inside the memory, equation (4) has to be implemented in the analog domain, as shown in equation (5).
$V_{Y_{\text{AVG}},k} = \frac{1}{N} \sum_{i=1}^{R \times R \times C} w_{k,i} \times V_{a,i}$  \hspace{1cm} (5)

Fig. 3. Concept of embedded convolution computation as averaging in SRAMs for binary-weight convolutional neural networks.

The equivalence of equations (4) and (5), conceptually shown in Fig. 3, becomes apparent in 3 key steps. First, the digital inputs ($X_{IN}$'s) are converted into analog voltages ($V_a$'s) using digital-to-analog converters (DAC). Then, the analog voltages are multiplied by the corresponding 1-bit filter-weights ($w_i$'s), which are stored in a memory array. This is followed by averaging over $N$ terms to get the analog-averaged convolution output voltage ($V_{Y_{\text{AVG}}}$). These constitute the second step: multiply-and-average (MAV). Finally, in the last step, the analog-averaged voltage is converted back into the digital domain ($Y_{OUT}$) using an analog-to-digital converter (ADC), for further processing. It may be noted that if the 3-D filter size ($R \times R \times C$) is greater than $N$, the above-mentioned 3-step process is repeated multiple ($N_f$) times using $R \times R \times C'$ ($\leq N$) elements in each cycle, where $N_f = C/C'$. The partial outputs (from the ADC) can then be further added digitally (outside the memory) to get the final convolution output.

III. OVERALL ARCHITECTURE

Fig. 4 shows the overall architecture of the 16 Kb Conv-SRAM (CSRAM) array, consisting of 256 rows by 64 columns of SRAM bit-cells. It is divided into 16 local arrays, each with 16 rows. Each local array is meant to store the binary filter-weights ($w_i$'s) for a different 3-D filter in a CONV/FC layer. $w_i$ is stored in a 10T SRAM bit-cell as either a digital ‘0’ or a digital ‘1’, depending on whether its value is +1 or -1 respectively. The 10T bit-cell consists of a regular 6T bit-cell and 2 decoupled read-ports. Each local array has its analog averaging circuits (MAV's) and a dedicated ADC to compute the partial convolution outputs ($Y_{OUT}$'s). Sharing these circuits for 16 rows in a local array reduces the area overhead. The IFMP values ($X_{IN}$'s) are fed into column-wise DACs, which convert the digital $X_{IN}$ codes to analog input voltages on the global bit-lines (GRBL's). The GRBL's are shared by all the local arrays, implementing the fact that in CNNs each input is shared/processed in parallel by multiple filters. With this architecture, the 16 Kb CSRAM array can process a maximum of 64 convolution inputs and compute 16 convolution outputs in parallel.

Fig. 5(a) shows the simulated test error-rates for the MNIST dataset with the LeNet-5 CNN, consisting of 2 CONV layers (C1, C3) and 2 FC layers (F5, F6). The number of bits to represent the IFMP/OFMP values are varied from 8 to 4. Lower bit-width helps in reducing the area/power costs of the DAC and ADC circuits involved for the convolution computations. However, as seen from Fig. 5(a), the error rate starts to increase steeply for <7-b. Hence, 7-b is chosen as the target bit-width for the DAC/ADC circuits. With 7-b (including the sign bit) the voltage resolution needed on a 1 V scale is 1 LSB = $1/2^7 \approx 15.6$ mV. Next, the effect of the averaging factor (‘$N$’) on the test error-rate is observed. A high value of ‘$N$’ would decrease the area/power overhead of the ADC by amortizing it over more MAV operations per clock-cycle. However, higher ‘$N$’ can also degrade the computation accuracy due to increased quantization by averaging. This is more critical for CNN layers with smaller filter sizes. As shown in Fig. 5(b) for layer F6, with a 3-D filter size of 120, the error-rate steeply increases as ‘$N$’ is varied from 15 to 120. For the other 3 layers of LeNet-5, the 2-D filter size is $5 \times 5$. Hence, a minimum $N = 25$ is required to fit at least one full filter channel per CSRAM row. We chose $N = 64$ to fit 2 channels for $5 \times 5$ filters, without sacrificing much on the error rate.

The number of rows ($N_{\text{rows}}$) per local array in the CSRAM...
determines the unit capacitance ($C_{LBL}$), which is used for all the analog operations required for the in-memory convolution computation. For every column in a local array, there is a corresponding MAV$_a$ circuit. Hence, a higher value of $N_{rows}$ would decrease the area-overhead of MAV$_a$, by amortizing it over multiple rows. It also reduces variation of the $C_{LBL}$ value, which helps in improving accuracy of the computations. However, a high value of $N_{rows}$ means a high $C_{LBL}$, which translates to increased energy costs. It would also lead to less throughput for a given SRAM size, since fewer outputs would be computed per cycle. Therefore, $N_{rows} = 16$ is chosen as a trade-off. It may be noted that with $N_{rows} = 16$, the thermal noise ($\frac{K T}{C}$) is $< 1$ mV, which is well below 1 LSB = 15.6 mV.

IV. KEY CONTRIBUTIONS OF THIS WORK

While there are a few different approaches [13]–[17] for in/near-memory computing, the proposed architecture has some key contributions, which provide significant benefits over prior work. The first key feature of our approach is the robustness to SRAM bit-cell $V_t$ variations. SRAM bit-cells use near-minimum transistor sizes available in a given CMOS process and hence, suffer from transistor mismatch and variation. For example, if we consider the discharge current ($I_{cell}$) through an SRAM bit-cell (shown in Fig. 6) we can observe that it has a significant spread from its mean value ($\sigma \approx 30\% \mu$). Now, when $I_{cell}$ is used to modulate the analog voltage ($V_a$) on the bit-line [13]–[15], [17], there is a wide variation in the $V_a$ value and it cannot be controlled very well. This compromises the computation accuracy and extra algorithmic techniques might be required to compensate for that. [13] uses the ‘AdaBoost’ technique, in which the results of many weak classifiers are combined to get a more accurate final result. However, this would lead to an increase in the number of computations and the energy required. [15] proposed an on-chip training to compensate for chip-to-chip variations. However, this would incur energy and timing penalty required to re-train the network corresponding to every single chip. In our approach (Fig. 6), the analog voltage ($V_a$) is directly sent to the bit-lines using global DACs at the periphery. Since the global DACs can be upsized, with their area being amortized over multiple rows (256 in this case), the variation due to it is significantly less compared to that of the bit-cell. Furthermore, the SRAM bit-cell is only used to multiply $V_a$ by the 1-b filter weight ($w_i$) stored in it, using full signal swing locally. That means, the purpose of the SRAM bit-cell is to discharge one of its local bit-lines to 0, it is not used to control $V_a$. Hence, given enough time for the worst-case bit-cell discharge, the computation accuracy does not suffer from local bit-cell $V_t$ variations.

The second key feature of our approach is the improvement of the dynamic voltage range for the analog computations without disturbing any bit-cell. In the conventional approach (with 6T SRAM bit-cells) [13], [14], [17], where multiple word-lines (WL) are activated for the same bit-line, there might be a situation where one of the accessed bit-cells in that column is in pseudo-write mode (Fig. 7). This is because multiple activated bit-cells in that column can discharge a bit-line to a very low voltage, which could overwrite the data stored ($Q_k = \lceil 1 \rceil$) in the disturbed bit-cell. Hence, the bit-line voltage range has to be limited to prevent any write-disturb. In our approach, 10T bit-cells are used which de-couple the read and the write ports, to prevent any write-disturb. Furthermore, each bit-cell is read independently in parallel without sharing any bit-lines. And hence, the discharge on one bit-line cannot affect another accessed bit-cell. Thus, we can utilize a wide voltage range (close to full-rail) for the analog computations, without disturbing any bit-cell. It may be noted that, although a 10T bit-cell has more transistors than a 6T, it can be designed using smaller sized devices, compared to a 6T bit-cell. This is because, unlike a 6T, a 10T bit-cell does not have conflicting sizing requirements to achieve high margins for both read and write operations. In addition, due to the limitations of 6T bit-cells for in-memory analog computations, network augmentation, i.e. larger sized neural networks, might be required to compensate for lower computation accuracy. Larger neural networks translate to increased storage requirement for filter-weights on-chip and hence, increased SRAM size. Therefore, overall, our 10T bit-cell based in-memory architecture is not necessarily higher area than a 6T based design.

The third key feature of our work, which distinguishes it from other “in-memory” computing approaches [14], [16], is the use of the inherent bit-line capacitances in the SRAM array to implement the computations. This precludes the need for extra area-intensive capacitors, which would be otherwise required at the SRAM periphery [14] to implement some of the analog computations.

Finally, this work supports multi-bit resolution for the inputs and outputs of the dot-products, compared to [13] (output: 1-b) and [16], [17] (both input/output: 1-b). This helps in achieving higher classification accuracy for a neural-network of a given size.

All the key features, described above, make our proposed architecture scalable i.e. multiple CSRAM arrays can operate in parallel to run larger neural networks.
V. CIRCUITS FOR THE 3-PHASE CONV-SRAM OPERATION

A. Phase-1: DAC

During the first phase of the Conv-SRAM (CSRAM) operation the digital convolution input \(X_{IN}\) is converted into an analog voltage \(V_a\) using a column-wise digital-to-analog converter (GBL_DAC). The analog voltage is used to pre-charge the global read bit-line (GRBL) and the local bit-lines (not shown in Fig. 4) in the SRAM array. Each GRBL is shared by all the 16 local arrays and hence, they get the same value of the analog pre-charge voltage. This implements the fact that in a given CNN layer (CONV/FC) each input is same value of the analog pre-charge voltage. This implements the convolution operation the digital convolution input \(X\).

Fig. 7 shows the comparison of the conventional and proposed approaches on write-disturb issue of SRAM bit-cells during compute mode.

Fig. 7. Comparison of the conventional and proposed approaches on write-disturb issue of SRAM bit-cells during compute mode.

Fig. 8. Schematic of the column-wise GBL_DAC circuit, showing the digital-to-analog converter (bottom-left) and time-to-analog converter (top-left). Also shown are the timing signals and operation waveforms for 2 input codes (right).

Fig. 8. Schematic of the column-wise GBL_DAC circuit, showing the digital-to-analog converter (bottom-left) and time-to-analog converter (top-left). Also shown are the timing signals and operation waveforms for 2 input codes (right).

The analog voltage is used to pre-charge the global read bit-line (GRBL) and the local bit-lines (not shown in Fig. 4) in the SRAM array. Each GRBL is shared by all the 16 local arrays and hence, they get the same value of the analog pre-charge voltage. This implements the fact that in a given CNN layer (CONV/FC) each input is processed simultaneously by multiple filters. Furthermore, all the 64 column-wise GBL_DACs operate in parallel and can send a maximum of 64 analog inputs to the CSRAM array in one clock cycle.

Fig. 8 shows the schematic of the proposed GBL_DAC circuit. It consists of a cascode PMOS stack biased in the saturation region to act as a constant current source. The GRBL circuit. It consists of a cascode PMOS stack biased in the saturation region to act as a constant current source. The

The circuit is shared by all the 16 local arrays and hence, they get the same value of the analog pre-charge voltage. This implements the fact that in a given CNN layer (CONV/FC) each input is processed simultaneously by multiple filters. Furthermore, all the 64 column-wise GBL_DACs operate in parallel and can send a maximum of 64 analog inputs to the CSRAM array in one clock cycle.

To design the pulse-widths of the 8 timing signals, we need to express \(X_{IN}\) in terms of its 2 components:

\[
X_{IN,dec} = 8 \times k_A + k_B,
\]

\[
k_A = \text{Decimal}(X_{IN}[5 : 3]),
\]

\[
k_B = \text{Decimal}(X_{IN}[2 : 0])
\]

where \(k_A\) and \(k_B\) are the decimal values for the 3 MSBs and the 3 LSBs of \(X_{IN}\) respectively. Since \(k_A\) and \(k_B\) can have any integer values from 0 to 7, the pulse-widths of the timing signals \((TD's)\) are chosen as:

\[
t_{TD_{9k}} = 8 \times k t_0 + k t_0 = 9 \times k t_0 \quad k \in (0, 1, \ldots, 7)
\]

where \(t_0\) is the minimum time resolution. A delay-line architecture, with a controllable unit delay of \(t_0\), is used to generate 64 time-delayed signals from the input clock. Then the appropriate signals are combined using NOR gates to generate the \(TD's\). This is done at the global level and the generated \(TD's\) are buffered and routed to all the GBL_DACs. A higher value of \(t_0\) reduces the non-linearities from the timing generation circuitry, at the cost of increased clock cycle time.

To understand how the 2-phase charging technique works, let us consider two \(X_{IN}\) values of 24 and 63, as shown in Fig. 8. For \(X_{IN} = 24 = 8 \times 3 + 0, k_A = 3\) and \(k_B = 0\). Hence, \(TD_{23}\) or \(TD_{27}\) is used in phase A and \(TD_0\) is used in phase B, to select the pulse-width of the \(ON\) timing signal. Similarly, for the code \(X_{IN} = 63 = 8 \times 7 + 7\), both \(k_A\) and \(k_B\) are 7 and hence, \(TD_{63}\) is used in both the charging phases.

In addition to the linearity aspect of the \(DAC\) transfer function, this architecture also performs better in terms of de-
vice mismatch, compared to binary-weighted PMOS charging DACs [13]. This is because, here, the same PMOS stack is used to charge the global bit-line for all input codes, rather than having to use smaller PMOS devices for small input values. Furthermore, the pulse-widths of the globally generated timing signals have less variations typically, compared to those arising from local \( V_C \)-mismatch in the PMOS devices [13].

It may be noted that, a one-time calibration is required to set the maximum value of the analog pre-charge voltage for the maximum input code \( (X_{IN,max}) \). The maximum pre-charge voltage should be kept lower than the supply voltage \( V_{DD} \) arising from local timing signals. During calibration, all DACs are fed the same input value of \( X_{IN,max} \). In a given clock cycle, first, the GBL_DAC precharges the \( GBL \) to an analog voltage \( (V_a) \). Then, \( V_a \) is compared to an externally provided reference voltage \( V_{ref} \) (typically kept at 1 V in this work). The comparison is done by the column-wise sense-amplifiers (SA), which are already present for normal read-out of the SRAM. All the 64 SAs operate in parallel and use the same \( V_{ref} \) to provide 64 comparison outputs simultaneously. \( V_{bias} \) is monotonically increased from 0 V until majority of the SAs (> 50%) flip their outputs (‘1’ to ‘0’), at which point the calibration is achieved. In this work, a 5 mV step-size is used to tune \( V_{bias} \).

### B. Phase-2: Multiply-and-Average

The second phase of the Conv-SRAM operation involves the multiplication of the analog input voltages \( (V_i) \) with the 1-b filter weights \( (w_i) \) and averaging over \( N \) values. This multiply-and-average (MAV) operation is done in parallel for all the 16 local arrays, each storing the \( w_i \)’s for a different 3-D filter when running a CONV/FC layer.

\[
V_{Y_{AVG,k}} = \frac{1}{N} \sum_{i=1}^{N} w_{k,i} \times V_{a,i}, \quad 0 \leq k \leq 15, \quad N \leq 64
\]

\[
V_{Y_{AVG}} = V_{P_{AVG}} - V_{n_{AVG}}
\]

Fig. 9 shows the details for the MAV operation for one local array. It starts by turning on the read word-line \( (RWL) \) for the selected row in the local array. This leads to discharging of one of the local bit-lines \( (LBLT, LBLF) \) in each column, depending on the \( w_i \) stored in the corresponding 10T bit-cell. A positive \( w_i \) (+1) is stored as a digital ‘0’ and a negative \( w_i \) (−1) as a digital ‘1’. It may be noted that, the local bit-lines have been pre-charged to the same analog voltage \( (V_{a,i}) \) as its corresponding global bit-line \( (GRBL) \) during phase-1. Therefore, at the end of weight evaluation, the difference between the local bit-line voltages represents the product of the analog voltage \( (V_{a,i}) \) and the 1-b weight \( (w_i) \). For example, the bit-cell in the ‘0th’ column stores a −1 and hence, \( \Delta V_{LBL,0} = V_{LBLT,0} - V_{LBLF,0} = -V_{a,0} \).

The weight multiplication/evaluation step is completed by turning off the \( RWL \). After that, the appropriate local bit-lines are shorted together horizontally to evaluate the average. The positive and negative parts of the average as obtained on two separate voltage rails, \( V_{P_{AVG}} \) and \( V_{n_{AVG}} \), respectively. This is implemented by the local MAV circuits, which pass the voltages of the \( LBLT \)’s and \( LBLF \)’s to either the \( V_{P_{AVG}} \) or the \( V_{n_{AVG}} \) voltage rails, depending on the sign of the input \( X_{IN} \). If the input for the particular column is positive \( (X_{IN} > 0) \), then \( ENP \) is on, otherwise \( ENN \) is on. \( ENP \) and \( ENN \) are digital control signals which are globally routed and shared column-wise by all the 16 local arrays. The switches controlled by \( ENP \) and \( ENN \) are implemented using NMOS pass transistors, since the final \( V_{P_{AVG}} \) and \( V_{n_{AVG}} \) voltages would be closer to 0 V than \( V_{ref} \). On the other hand, the switches controlled by \( PCHG \) use CMOS transmission gates, since they need to pass a wide range of voltages from 0 V to \( V_{ref} (~1 \text{ V}) \), during phase-1 (DAC pre-charge).

The fully-differential nature of the averaging architecture helps in mitigating many common-mode noise issues, e.g. clock coupling noise from the control switches, capacitance variation of the local bit-lines and the voltage rails due to different process corners, etc. This helps in improving the accuracy of the dot-product computations with our approach.

It may be also noted that during this phase, when the SRAM bit-cell is actually used for weight-evaluation, the time required does not have a huge variation. Fig. 10 shows the simulated local bit-line discharge time \( t_{dis,LBL} \) in the slowest process corner (SS). As seen from the figure, even the \( 6\sigma \) value of \( t_{dis,LBL} \) is merely 500 ps, which is much smaller than the total clock period (≈ 100 ns). This shows that bit-cell \( V_i \) variations do not dominate the overall computation time. The longer clock period is justified due to the highly parallel processing in the compute mode.
to track the local bit-line capacitance better in the presence of process and temperature variations. The SA has a standard StrongARM latch-type architecture. PMOS devices are chosen for the input differential pair of the SA, since the common mode voltages of $V_{P\text{AVG}}$ and $V_{n\text{AVG}}$ signals are expected to be closer to the GND rail. The logic block provides the timing signals for the charge-sharing ($PCH_R, EQ_P, EQ_N$) and the SA comparison ($SA\_EN$), using the globally provided timing signals ($\phi_1, \phi_2$). It also has a counter to count the number of cycles it takes to finish the ADC operation and that provides the digital output of the dot-product computation.

Fig. 12 also shows the waveforms for a typical CSH_ADC operation. It starts by sending a $SA\_EN$ pulse from the ADC logic block to the SA. The SA compares $V_{P\text{AVG}}$ and $V_{n\text{AVG}}$, and sends its outputs ($SAO_P, SAO_N$) to the ADC logic block. The first comparison determines the sign of the output, e.g. for the case shown in Fig. 12, $Y_{OUT}$ is positive since $V_{P\text{AVG}}$ is higher than $V_{n\text{AVG}}$. After the first comparison, the lower of the 2 voltage rails ($V_{n\text{AVG}}$) is integrated by the SA and sends the reference local bit-line ($BL\text{N}_{\text{ref}}$), using the equalize signal ($EQ_N$ in this case). The reference bit-line, which replicates the local bit-line capacitance, was pre-charged during the SA comparison using the $PCH_R$ signal to $V_{\text{ref}}$ (1 V in this work). Therefore, the time to charge the equalize/integrate operations, along with the SA comparison, is $\approx \frac{V_{\text{ref}}}{V_{P\text{AVG}} - V_{n\text{AVG}}}$, where $N$ is the number of SRAM local columns that were averaged. The pre-charge and equalize/integrate operations, along with the SA comparison, are continued until the lower voltage rail ($V_{n\text{AVG}}$) exceeds the higher one ($V_{P\text{AVG}}$). When this happens, the SA outputs flip indicating the end-of-conversion ($EOC$). After this, no more timing pulses are generated. A counter in the ADC logic block counts the number of equalize pulses ($EQ_N$) it takes to reach $EOC$ and that generates the digital value of the convolution/dot-product output ($Y_{OUT}$), which is +4 for the example shown in Fig. 12.

It may be noted that $Y_{OUT}$ is directly affected by the SA offset voltage ($V_{OS}$), which can degrade the overall computation accuracy due to incorrect ADC outputs. To address this issue, we propose a simple 2-cycle offset-cancellation technique, using a flipping mux at the input of the SA (Fig. 13). During the first/even cycle of this 2-cycle period, $FLIP = '0'$. Hence, $V_{P\text{AVG}}$ and $V_{n\text{AVG}}$ are passed to the positive and negative input terminals of the SA respectively. Therefore,
\[ Y_{OUT,0} = ADC(V_{YAVG,0} - V_{OS}) \]

The output in this cycle is exactly same as in the conventional case (without the flipping mux). However, during the next odd cycle, the inputs to the SA are flipped by setting \( FLIP = '1' \). Hence, a differential voltage of \( (V_{nAVG} - V_{pAVG}) = -V_{YAVG} \) is applied to the input of the SA. To get the correct polarity at the output of the ADC, another negation is applied by the ADC logic block. This results in \( Y_{OUT,1} = -ADC(-V_{YAVG,1} - V_{OS}) = ADC(V_{YAVG,1} + V_{OS}) \), as compared to \( Y_{OUT,1} = ADC(V_{YAVG,1} - V_{OS}) \) for the conventional case. Finally, we add the \( Y_{OUT} \) s for the 2 consecutive cycles to accumulate the partial results for the convolution. With our proposed 2-cycle approach the effect of \( V_{OS} \) is inherently canceled since:

\[
Y_{OUT} = Y_{OUT,0} + Y_{OUT,1} = ADC(V_{YAVG,0} + V_{YAVG,1})
\]

On the other hand, for the conventional case the effect of \( V_{OS} \) adds up since:

\[
Y_{OUT} = ADC(V_{YAVG,0} + V_{YAVG,1} - 2 \times V_{OS})
\]

and this makes the accumulation result further inaccurate. It may be noted that, the benefits of this offset-cancellation technique comes without any extra timing and power penalty, as long as an even number of cycles are required to finish a full convolution computation. This can be easily expected for most CNNs.

VI. MEASURED RESULTS

The 16 Kb CSRAM array was implemented in a 65-nm LP CMOS process. The die photo in Fig. 14 shows the relative area occupied by the key blocks. The bit-cell array (ARY) along with its peripheral circuitry occupies 73.1\% of the total CSRAM area, 8.2\% is occupied by the GBL_DACs, 8.6\% by the local MAV\(_a\) circuits, 7.3\% by the CSH_ADCs and the rest by global timing circuits. The test-chip summary is shown in Table I.

Fig. 12. Architecture (left) for the charge-sharing based ADC (CSH_ADC) for 1 local array of the Conv-SRAM and typical waveforms (right) for the digital output (\( Y_{OUT} \)) computation for the convolution (dot-product) operation.

Fig. 14. Die micro-photograph in a 65-nm CMOS process.

A. Circuit Characterizations

Fig. 15 shows the measured transfer function of the 6-b GBL_DAC. With \( V_{ref} = 1 \) V and \( t_0 \approx 250 \) ps, a 5-b voltage resolution is effectively achieved in measurements. Hence, we set the LSB of \( X_{IN} \) to ‘0’. To estimate the DAC analog output voltage \( (V_a) \), \( V_{GRBL} \) for the 64 columns of the CSRAM are compared to an external voltage \( (V_{ref}) \) by column-wise SAs, as explained before for DAC calibration (Section V-A). For each \( X_{IN} \), the \( V_{ref} \) at which more than 50\% of the SA outputs flip is chosen as an average estimate of \( V_a \). As mentioned before, an initial one-time calibration is needed to set \( V_{a,max} = 1 \) V for \( X_{IN} = 31 \) (max. input code). The supply voltage for the DACs is fixed at 1.2 V, to keep the PMOS...
DAC at ADC with and without the offset-cancellation (OC) technique, for two Videal 250 MHz ADC, (mV) ADC, is much lower than 5 MHz ADC, all ADCs, all

TABLE I
TEST-CHIP SUMMARY

| Technology   | 65-nm |
|--------------|-------|
| CSRAM size   | 16 Kb |
| CSRAM area   | 0.063 mm² |
| Array org.   | 256×64 (10T int-cells) |
| # column DACs| 64    |
| # row ADCs   | 16    |
| Max. # MAVs  | 64×16 |
| Supply volts | 1.2 V (DAC), 0.8 V (array), 1 V (rest) |
| Clock freq.  | 5 MHz |
| ADC clock freq. | 250 MHz |

Fig. 15. Measured transfer function of GBL_DAC at $V_{dd, D AC} = 1.2$ V, with $V_{ref} = 1$ V and $t_0 \approx 250$ ps.

stack in them operating in the saturation region (as a constant current source). It can be seen from Fig. 15 that there is a good linearity in the DAC transfer function with $DNL < 1$ LSB. Since the SAs have NMOS input-pair, low values of $V_a$ cannot be properly estimated. Hence, the characterization is done till $X_{IN} = 16$ (or $V_a \approx 500$ mV).

Fig. 16. Measured transfer function and energy consumption of CSH_ADC at $V_{dd, ADC} = 1$ V, $V_{dd, AR Y} = 0.8$ V and $f_{ADC} = 250$ MHz.

The effect of the offset-cancellation (OC) technique for the SA (in the CSH_ADC) is also characterized, as shown in Fig. 17 for two different input codes. It can be clearly seen that the OC helps in reducing the variation of the $Y_{OUT}$ values, leading to a better computation accuracy for the dot-products/convolutions.

B. Test Case: MNIST Dataset

To demonstrate the functionality for a real CNN architecture, the MNIST hand-written digit recognition dataset is used with the LeNet-5 CNN [18]. As shown in Fig. 18, LeNet-5 consists of 2 CONV layers (C1, C3) and 2 FC layers (F5, F6). In addition, there are 2 sub-sampling or max-pooling layers (S2 and S4, following layers C1 and C3 respectively) and a non-linear ReLU layer (R5 after layer F5). Only the CONV/FC layers, which involve majority of the computations, are implemented on-chip by the CSRAM array. The non-linear layers are implemented in software. Fig. 19 shows the test setup used to automatically run the 4 CONV/FC layers, one...
after the other, on the test-chip. Data is transferred back and forth between MATLAB (running on a host PC) and the test
chip, via an FPGA board. Table II shows the detailed mapping of the 4 CONV/FC layers to the CSRAM array to compute the convolutions. Let us first consider layer C3. It has a filter size of $5 \times 5$, with 6 input channels and 16 output channels (number of 3-D filters). Each of the 16 3-D filters are mapped to one of the 16 local arrays in the CSRAM. Since each row in the local array has 64 bit-cells, a maximum of $2 \left( \frac{64}{28} \right)$ input channels can fit per row. Therefore, 3 ($= \frac{6}{2}$) rows are required in each local array to fit the entire 3-D filter. In every clock cycle, 50 ($= 5 \times 5 \times 2$) $X_{IN}$’s are sent through a buffer (shift-registers) to the CSRAM array to compute 16 partial convolution outputs. Thus, the CSRAM array processes $50 \times 16 \times 2$ operations ($1 \text{ MAV} = 2 \text{ OPs}: 1 \text{ multiply} + 1 \text{ add/average}$) per clock cycle. For layer F5, the entire filter cannot fit at once in the CSRAM array (due to its limited 16 Kb size in the test-chip). Hence, the entire process, explained above, is repeated multiple times to finish all the computations. However, having multiple CSRAM arrays operating in parallel can easily alleviate this problem, by fitting all the filter weights together on-chip.

![Diagram](image)

**Fig. 19.** Test setup for automatically running the 4 CONV/FC layers of LeNet-5 CNN on Conv-SRAM, for a given input image ($28 \times 28$).

Fig. 20 shows the measured error rate for the 10,000 test images in the MNIST dataset, with the 4 CONV/FC layers being successively implemented on-chip. 3 different chips are measured, each experiment is repeated multiple times, and the average value of the error rate is reported. We tested 2 different versions of LeNet-5: with and without Batch-Normalization (BN) layers preceeding the CONV/FC layers. Without BN layers (‘v1’) we achieve a classification error rate of 2.5% after all the 4 layers. The error rate is improved to 1.7% by using the BN layers (‘v2’). This is mostly because BN normalizes the convolution inputs for every layer, with a mean around 0 and also limits the maximum value of the inputs. Hence, after input quantization to 6-b, its features are better preserved compared to an un-normalized input distribution. The measured error rate, which is close to the expected value from an ideal digital implementation, shows the robustness of the CSRAM architecture to compute convolutions. The error rate for the MNIST dataset is improved by 8.3% compared to prior work on in/near-memory compute [13], [16], where a 10% error rate was achieved. Next, we tested functionality at a lower voltage setting of $V_{dd, DAC} = 1 \text{ V}$ and the rest of the circuits operating at $V_{dd} \text{ (rest)} = 0.8 \text{ V}$, with a clock period of 400 ns. The maximum DAC pre-charge voltage ($V_{a,max}$), corresponding to the maximum input code, is calibrated to 0.8 V. Hence, the magnitude of 1 LSB is $\sim 32 \text{ mV}$ for the previous case with $V_{a,max} = 1 \text{ V}$). Fig. 21 shows the measured error rate for this set of voltages. Due to reduced analog voltage precision, the error rates are slightly higher, with ‘v1’ achieving 3.4% and ‘v2’ achieving 1.9% for the MNIST test dataset.

![Graph](image)

**Fig. 20.** Measured error rate for the 10K images in the MNIST test dataset using LeNet-5 CNN, with and without BN, at $V_{dd} = V_{a,max} = 1\text{ V}$.

The distributions of the partial convolution outputs from the ADC ($Y_{OUT}$’s) are shown in Fig. 22, for all the 4 CONV/FC layers. For each of these layers, $Y_{OUT}$ has a mean around $\approx 1 \text{ LSB}$, which justifies the use of the serial ADC topology to compute it. Fig. 23 shows the distributions of the convolution inputs ($X_{IN}$’s) for the 4 layers. $X_{IN}$’s have been properly

---

**Table II**

| Parameters | C1 | C3 | F5 | F6 |
|-----------|----|----|----|----|
| 3-D Filter size | $5 \times 5 \times 1$ | $5 \times 5 \times 6$ | $5 \times 5 \times 16$ | $1 \times 1 \times 120$ |
| # 3-D Filters | 6 | 16 | 60 | 10 |
| # Local ARY’s used | 6 | 16 | 15$^1$ | 10$^1$ |
| # IFMP channels/row | 1 | 2 | 2 | 30$^2$ |
| Rows, Col.s/local ARY | 1, 25 | 3, 50 | 8, 50 | 4, 30$^3$ |
| # cols for AVG ($N$) | 32 | 50 | 50 | 32 |
| # operations/cycle | $25 \times 6 \times 2$ | $50 \times 16 \times 2$ | $50 \times 15 \times 2$ | $30 \times 10 \times 2$ |

$^1$ Repeated 8 times to cover all the 120 filters
$^2$ 15 columns and 8 rows mapping used at $V_{DD} = 0.8 \text{ V}$
$^3$ 10% error rate was achieved. Next, we tested functionality at a lower voltage setting of $V_{dd, DAC} = 1 \text{ V}$ and the rest of the circuits operating at $V_{dd} \text{ (rest)} = 0.8 \text{ V}$, with a clock period of 400 ns. The maximum DAC pre-charge voltage ($V_{a,max}$), corresponding to the maximum input code, is calibrated to 0.8 V. Hence, the magnitude of 1 LSB is $\sim 32 \text{ mV}$ for the previous case with $V_{a,max} = 1 \text{ V}$). Fig. 21 shows the measured error rate for this set of voltages. Due to reduced analog voltage precision, the error rates are slightly higher, with ‘v1’ achieving 3.4% and ‘v2’ achieving 1.9% for the MNIST test dataset.
scaled and quantized to 6-b (including sign bit) before being sent to the CSRAM array to compute the convolutions. As seen from the figure, all the layers have a high proportion of 0’s for the $X_{IN}$’s. This helps in reducing the GBL_DAC energy to convert and send them to the columns of the CSRAM array.

Fig. 24 shows the overall energy consumption of the CSRAM array for running the different layers of LeNet-5, with $V_{dd,DAC} = 1.2$ V, $V_{dd,ARY} = 0.8$ V, $V_{dd}$ (rest) = 1 V and $f_{clk} = 5$ MHz. Of the 4 CONV/FC layers in LeNet-5, the energy consumptions while running layers C1 and F6 are lower than that of layers C3 and F5. This is because layers C1 and F6 do not fully utilize the entire CSRAM array, due to their small filter sizes. However, that also translates to a lower energy-efficiency for these layers (Table III), since the energy is amortized over fewer MAV operations. A higher array utilization in layer C3 (all 16 local arrays) helps in achieving an energy-efficiency of 38.8 TOPS/W (‘v1’), by consuming 41.3 pJ of energy for $50 \times 16 \times 2 = 1600$ operations.

Whereas, for ‘v2’ (with BN), layer F5 achieves the best energy-efficiency of 40.3 TOPS/W, utilizing 15 of the 16 local arrays. Fig. 24 also shows the energy breakdown for the 3 major components: GBL_DAC, ARY+MAVa and CSH_ADC. The energy for the GBL_DACs is limited by the bit-precision requirement for representing the IFMP values. Whereas, the energy for the ARY, MAV and CSH_ADC circuits can be scaled down by scaling their supply voltages while sacrificing speed. Fig. 25 shows the measured energy consumption of the CSRAM array, with $V_{dd,DAC} = 1$ V, $V_{dd}$ (rest) = 0.8 V and $f_{clk} = 2.5$ MHz. The reduced supply voltages help in decreasing the energy consumption, leading to better energy-efficiency numbers (Table IV).

**TABLE III**

| Type ($) | C1 | C3 | F5 | F6 |
|----------|----|----|----|----|
| v1: without BN | 14.8 | 38.8 | 38.8 | 24.3 |
| v2: with BN | 14.7 | 33.5 | 40.3 | 23.2 |

$^*$ 1 MAV = 1 multiply + 1 average = 2 OPs, with 6-b inputs and 1-b weights

Recent hardware implementations [5], [9], [16], [19]–[21] for NNs have focused on reduced bit-precisions to achieve higher energy-efficiency. Table V presents comparison with...
prior work, both conventional digital [5], [9], [19], [20] and in-memory approaches [14], [16]. It should be noted that, while [5], [9], [16], [19], [20] are full systems, the main focus of this work was to demonstrate in-memory computation capability for CNNs. Hence, ours does not include the energy for IFMP/OFMP memories. However, for the MNIST dataset with LeNet-5 CNN, we estimate [22] those to have only small contributions to the overall energy-efficiency per MAV operation, due to the high parallelism supported by our in-memory approach. Furthermore, as seen from Fig.s 23 and 22, both the inputs ($X_{IN}$’s) and the partial outputs ($Y_{OUT}$’s) have a high proportion of ‘0’’s. Hence, in future work, data-dependent memory architectures e.g. 8T SRAMs, [23], [24] can be used to store and access the inputs/outputs. [23], [24] take advantage of data properties to significantly reduce memory-access energy, which would be highly useful here. Compared to [5], [9], we achieve $>27\times$ improvement in energy-efficiency, due to the massively parallel in-memory analog computations. Our work achieves similar energy-efficiency numbers as [19] (considering a simplified technology scaling model), while using 6-b for IFMP/OFMP, compared to 1-b in [19]. Whereas, we achieve similar classification accuracy as [19] on MNIST, using $\sim 37\times$ less MAC/MAV operations per classification. Our numbers are also comparable to the energy-efficiency of [20] (not quoted for MNIST), which uses 1-b for weights. Next, we compare our results to an in-memory mixed-signal computing approach [13], which implements a support vector machine (SVM) algorithm with 45 binary-classifiers, for a 10-way MNIST digit classification. We achieve similar energy-efficiency as [13], while improving the classification accuracy from 90% to 98%. This is because, we mitigate the problem of degraded computation accuracy, caused by SRAM bit-cell variation. Although [13] can run at a higher speed, it only supports 1-b output resolution, compared to 6-b in our case. When compared to a near-memory approach [16], which uses only 1-b for IFMP/OFMP, we still achieve $8.5\times$ improvement in energy-efficiency. This is because, our approach exploits high parallelism of accessing multiple memory addresses simultaneously, without the need to sequentially and explicitly read out the data (filter weights) from the memory. We also achieve a higher classification accuracy compared to [16], because of using 6-b for inputs/outputs. Finally, our

### TABLE V

**Comparison with prior work on low bit-width hardware implementations of ML algorithms**

| Metric | This work | ISSCC’17 [9] | JSAC’17 [5] | ISSCC’17 [13] | JSAC’18 [14] | ISSCC’18 [16] | CICC’18 [19] | ISSCC’18 [20] |
|--------|-----------|-------------|------------|--------------|--------------|--------------|--------------|--------------|
| Tech. (nm) | 65 | 28 | 40 | 130 | 65 | 65 | 28 | 65 |
| Voltage (V) | 1 | 0.715 | 0.8 | - | 1 | 0.55 | 0.66 | 0.65 |
| Computation Mode | In-Memory, mixed-signal | Digital | Digital | In-Memory, mixed-signal | In-Memory, mixed-signal | Near-Memory, Digital | Digital |
| ML Algo. | CNN (4-layer) | FC-DNN (4-layer) | CNN (5-layer) | SVM$^1$ | k-NN$^2$ | FC-DNN (12-layer) | CNN (5-layer) | CNN |
| ML Dataset | MNIST | MNIST | MNIST | MNIST | MNIST | MNIST | MNIST | FER2013 |
| # MAC(V)’s per classification | 406.8K | 334.3K | 406.8K | 3.65K$^1$ | 16.4K | 768.1K | 15M | - |
| Classification Accuracy | 98.3% (1V) | 98% (0.8V) | 98.36% | 98% | 90% | 92%$^2$ | 90.1% | 97.4% |
| # bits for IFMP/OFMP | 6 | 8 | 6 | 5/1 | 8 | 2 | 1 | 16 |
| # bits for Weights | 1 | 8 | 4 | 1 | 8 | 1 | 1 | 1 |
| SRAM Size (KB) | 2 | 1024 | 144 | 2 | 16 | 102.1 | 328 | 256 |
| Peak Throughput (TOPs) | 8 (1V) | 4 (0.8V) | 10.7 | 102 | 57.6$^1$ | 10.2 | 380.2 | 90 | 368.6 |
| Peak Energy Efficiency (TOPs/W) | 40.3$^4$ (1V), 51.3$^3$ (0.8V) | 1.86 (0.345)$^5$ | 1.75 (0.663)$^5$ | 11.51$^4$ (46.0)$^5$ | 1.94 | 6.0 | 130 (24.12)$^5$ | 50.6 |

$^1$ SVM: Support Vector Machine with 45 binary classifiers, each with 81 inputs, i.e. $81 \times 45$ MACs per 10-way classification
$^2$ k-NN: k-Nearest Neighbor, only 4-output classes (out of 10) were demonstrated with 100 test images
$^3$ # assume 2 operations (OPs) for 1 MAC (1 mult. + 1 acc.), similar to a MAC (1 mult. + 1 acc.)
$^4$ Does not include energy to access IFMP/OFMP memories
$^5$ Assuming a 65-nm implementation and Energy (Tech.)

### TABLE IV

**Measured energy-efficiency* (TOPs/W) for the CONV/FC layers of LeNet-5 CNN, at $V_{dd} = 0.8$V**

| Type | C1 | C3 | F5 | F6 |
|------|----|----|----|----|
| v1: without BN | 19.7 | 49.6 | 48.9 | 17.0 |
| v2: with BN | 19.6 | 51.3 | 49.6 | 16.5 |

* 1 MAV = 1 multiply + 1 average = 2 OPs, with 6-b inputs and 1-b weights
work also achieves better classification accuracy than prior in-memory approach [14], although it supports 8-b weights. This is because we reduce the effect of bit-cell variation when evaluating the weights. In addition, our approach benefits from more parallelism, by supporting 16 different dot-product computations per array per cycle, compared to 1 for [14].

VII. CONCLUSIONS

This paper presents an SRAM-embedded convolution (dot-product) computation architecture for running binary-weight neural networks. We demonstrated functionality with the LeNet-5 CNN on the MNIST hand-written digit recognition dataset, achieving classification accuracy close to digital implementations and much better than prior in-memory approaches. This is made possible by our variation-tolerant architecture and also the support of multi-bit resolutions of input/output values. Compared to conventional digital accelerator approaches using small bit-widths, we achieve similar or better energy-efficiency, by overcoming some of the major limitations of memories in traditional computing paradigms. This is because our architecture can significantly reduce data transfer by running massively parallel analog computations inside the memory. The results indicate that the proposed energy-efficient, SRAM-embedded dot-product computation architecture could enable low power ML applications (e.g. “always-ON” sensing) for “smart” devices in the Internet-of-Everything.

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