A Novel Architecture for 10-bit 40MSPS Low Power Pipelined ADC using Simultaneous Capacitor and Op-amp Sharing Technique

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A novel architecture for 10-bit 40MSPS low power Pipelined ADC using simultaneous capacitor and op-amp sharing technique

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Abstract : This work presents a low power 10-bit 40 MSPS Pipelined ADC with 1.8V supply voltage in a 180nm silicon based CMOS process. Simultaneous capacitor sharing and op-amp sharing technique is used between two successive stages of a Sample-and-Hold Amplifier (SHA) to reduce the power consumption. The memory effect in the proposed ADC is eliminated by a low input capacitance variable g_m op-amp. The differential and integral nonlinearity of the converter are within LSB. Simulation results show that the required Signal-Frequency-Dynamic range (SFDR) of 70dB, Signal-to-Noise-plus Distortion Ratio (SNDR) of 56.1dB and 9.02 Effective Number of Bits (ENOB) has been achieved with a 2MHz, $1-V_{p-p,diff}$ input signal while consuming only 7.3mW power from 1.8V supply.

Keywords: op-amp, capacitor sharing, power consumption, Dynamic comparator, Analog to Digital converter (ADC)

1 Introduction

With the increasing demand for the Application Specific Integrated Circuits (ASICs) there is a wide development of mixed-signal design, compatible with System-on-chip (SoC). Data converters have become an important block in SoC design for data communication and image processing applications [1]. Among the different ADC design pipelined ADCs are suitable for high speed, medium resolution and low power applications [2--4]. Since the accuracy decreases in the later stages of pipeline ADC, proper scaling of capacitors are required in pipelined ADC [5-7]

Sharing an op-amp between two consecutive stages can further reduce power consumption [8-10]. Furthermore switched-op-amp techniques [11-12] was proposed to reduce the power consumption of pipelined ADC. The maximum power is consumed in the first stage of the pipelined ADC [12].

To reduce the power consumption of ADC capacitor sharing technique was proposed [13-16]. To further reduce the power consumption, the front-end sample and hold circuit (S/H) is eliminated [17-18]. SHA introduces additional noise signal which is integrated to the analog signal and results in large power dissipation and occupies more die area. Also the S/H circuit is removed by integrating the SHA in first stage of pipeline ADC to reduce the power [19]. Hence it is essential for certain applications to simplify the pipelined ADC

by eliminating the supplementary blocks such as SHA circuit [20]. The aperture error is reduced by matching the delays that occur between the sampling networks of the sub-blocks including first MDAC and comparators [21]. In this paper, a novel SHA-less pipelined ADC using a combination of simultaneous capacitor and amplifier sharing with a sampling frequency of 40 MSPS is presented. To achieve low- power and high performance, the P-gain and N-gain boosted variable g_m op-amp and dynamic comparators have been designed.

2 Architecture of SHA-less Pipelined ADC

The proposed SHA-less Pipelined ADC architecture is shown in Fig.1. It mainly contains a SHA-less front-end followed by second stages in which two op-amps are shared in neighboring stages followed by 3-bit Flash ADC. In order to limit the overall conversion rate the bandwidth of the Switched Capacitor (SC) circuit is to be maximized. The closed-loop gain of the amplifier is 2 so that a large feedback factor and a low load capacitance is achieved. Since the capacitance load is very less and the feedback factor is large the bandwidth of the amplifier which is used in the intermediate stages can be increased. Fig. 2 shows the first and second stages of the ADC.

Fig. 1 A 10-bit SHA-less Pipelined ADC structure

Fig. 2 First and second stages of 10-bit Pipelined ADC

2.1 Operational amplifier

An op-amp is the main important sub-circuit of analog systems[22-25]. The common mode noise and even harmonics are reduced in the fully differential amplifier which employs a two-stage topology when compared with the single ended output op-amp. Also the linearity is increased in the fully differential amplifier. Fig. 3 shows the designed two-stage fully differential OTA which uses the gain boosted auxiliary amplifier. Here the fully differential structures are used in P gain boost and N gain boost amplifiers. With 1.8 V supply voltage the gain boosted amplifier is simulated initially to verify the desired specifications which was described by [26-28].

Fig. 3 Fully differential OTA

2.1.1 Design of the op-amp:

The op-amp is the most essential block in ADC[29-30]. The output impedance R_{out} is increased by the added gain stage A_{OTA} as given in the following equation:

$$R_{out} = (g_{m2} r_{o2} (A_{add} + 1) + 1) r_{o1} + r_{o2} \quad (1)$$

Further more ,the DC gain of the op-amp is improved in several orders of magnitude:

$$\begin{aligned} A_{0,tot} &= g_{m1} r_{o1} (g_{m2} r_{o2} (A_{nsw} + 1) + 1) \\ &= A_{OTA} g_{m1} g_{m2} r_{o1} r_{o2} \end{aligned} \quad (2)$$

Fig. 4 Cascoded gain stage with gain boosting technique

To improve the voltage gain of the folded cascode amplifier the output impedance of the circuit need to be increased as shown in Fig.4. The design procedure starts with the design of fully differential op-amp and the second step is to introduce the gain boosting amplifier to obtain the desired gain without affecting the bandwidth of the op-amp. To start with, the sizing of the main differential input pair of the transistors MN₀, MN₁ is selected using the desired phase margin and the gain bandwidth specifications. This op-amp design includes the design of differential inputs, differential outputs, folded cascode bias circuits with common mode feedback (CMFB) and gain boosting amplifiers.[31-33]

The feedback factor β is given by:

$$\beta = \frac{C_f}{C_1 + C_p + C_f} \quad (4)$$

The unity gain frequency is given by:

$$\omega_{unity} = g_m \frac{C_i + C_p + C_f}{C_o (C_i + C_p + C_f + (C_i + C_p) C_f)} \quad (5)$$

The settling time τ can be found by:

$$\tau = \left\{ C_p + C_i + C_o + \frac{(C_i + C_p) C_o}{C_f} \right\} / g_m \quad (6)$$

In Fig.5 the overdrive voltage of M10 is assigned high than that of input transistors M8 and M9 to boost the gain. The gain of this op-amp is mostly based upon the transistors M3, M5 and M8. The transistors M0 and M1 should be matched and kept wide enough to act as a resistor i.e. the overdrive voltage assigned is more compared to all.

Fig. 5 P gain boost amplifier

The bias circuit is used in this op-amp is current biasing circuit which is current source with one Silicon based NMOS for n-bias and current source with one Silicon based PMOS for p-bias. The current in the current source is scaled to give the bias voltage needed to make the circuit stable. Fig. 6 shows the N gain boost amplifier.

Fig. 6 N gain boost amplifier

2.1.2 Analysis of the op-amp

For the residue signal generated by the MDAC the op-amp loop-gain has to be at least 2^9 (54 dB). In the design of the op-amp the main stage is the folded cascode op-amp which produces high gain. The V_{cm} node is used to control the common mode bias voltage of the op-amp. Table 1 shows the performance summary of P Gain and N Gain Boost Amplifiers.

Table 1 Performance Summary of P Gain and N Gain Boost Amplifiers

Specifications	P Gain Boost Amplifier	N Gain Boost Amplifier
Supply Voltage (V)	1.8	1.8
Input Voltage (V)	0.5	0.5
Gain (dB)	23.99	18.217
Phase Margin (Deg)	72	73
Power Consumption (mW)	0.7	0.7

Simulation results show that the total DC gain of the op-amp is 86dB. Phase Margin value of the variable g_m op-amp is greater than 80° . Table 2 shows the performance summary of variable g_m op-amp.

Table 2 Performance summary of variable g_m op-amp

Specifications	Variable g_m Op-amp
Supply Voltage (V)	1.8
Input Voltage (V)	0.5
Gain (main amplifier)	45.03dB
Total Gain	86 dB
Gain Margin	65.5 dB
Phase Margin	$85-90^\circ$
Power Consumption (mW)	1.64
Common Mode Rejection Ratio (dB)	76.14 dB
Power Supply Rejection Ratio (dB)	54.97dB
GBW product (MHz)	230
Unity Gain Frequency(MHz)	7.3
Bandwidth(MHz)	486.3
Output swing	1.65V

3 Simultaneous Capacitor sharing and Op-amp Sharing MDAC

The capacitor C_{1f} is composed into two parts: C_{2f} and C_{2s} . During the phase Φ_1 , the analog input signal is sampled on the two capacitors C_{1f} and C_{1s} . The residue signal is generated from each stage during the phase Φ_2 and the residue signal is not provided to the sampling capacitors in the second stage of the pipelined ADC. The residue signal is detained on C_{1f} which is the feedback capacitor. Using the stored value on the capacitor C_{1f} , residue signal is produced. This stored value is provided to the sampling capacitors of the third stage C_{3f} and C_{3s} during the next phase Φ_1 . When the second stage is generating its residue signal, the next input signal need to be sampled are used [34-36]. The feedback capacitors named as C_{1fe} and C_{1fo} are used. The matching network is used in the initial stage of ADC. In the proposed pipelined ADC the simultaneous capacitor sharing and op-amp sharing technique is used. Fig.7 shows the simultaneous capacitor and op-amp sharing technique.

Fig.7 Simultaneous capacitor and op-amp sharing technique

3.1.1 Sampling Network for Stage1

The use of S/H circuit is avoided with the MDAC and the sub-ADC performing the sampling procedure [37]. In this work, the aperture error is minimized by matching the sampling network through cautious design of the aspect ratio (W/L) of the transistor. The gate source voltage (V_{gs}) is same for the sampling capacitance C_s and C_{com1} - C_{com4} in the sampling network. This is because the above mentioned sampling capacitance is connected to the same common mode voltage. Fig.8 shows the proposed sampling network for the initial stage. Fig.9 shows the timing diagram for the sampling network.

Fig. 8 Proposed sampling network for the first stage

Fig. 9 Timing Diagram for the sampling network

4 Results and Discussion

The SHA-less Pipelined ADC is simulated in a 180nm Silicon based CMOS process. Various sub circuits like dynamic comparator, op-amp, MDAC, Sub ADC and sub DAC was integrated for pipelined ADC[38-39]. Fig.10 shows the 1024 points FFT spectrum with the input frequency of 2 MHz.

Fig.10 Obtained FFT spectrum of pipelined ADC

Fig. 11 shows the layout of 10-bit SHA less pipelined ADC. The delay and the power consumption obtained after post layout simulation is 15.1ns and 15mW. C.

Fig. 11 The layout of the 10-bit SHA less Pipelined ADC

Table 3 shows the Performance summary of 10-bit SHA less pipelined ADC. The 10-bit SHA less pipelined ADC was designed in a 180nm Silicon based CMOS process and achieves 56.12 dB SNDR, 56 dB SNR, 70 dB SFDR, from a 1.8V supply voltage.

Table 3 Performance Summary of 10-bit SHA less Pipelined ADC

Technology	180nm
Resolution	10bit
Input Range	1V _{p-p}
Frequency Range(f_{in})	2MHz
Sampling frequency(f_s)	40MHz
DNL	+0.57/-1 LSB
INL	+0.74/-0.74LSB
SFDR	70dB
SNDR	56.12dB
ENOB	9.0289 bits
FOM	0.35pJ/step
Total Power	7.3mW
Conversion time	2.5ns
Area	0.63mm ²

Table 4 Comparison of the SHA-less pipelined ADC with some reported pipelined ADCs.

Reference	Technology (nm)	Supply Voltage (V)	Sample Frequency (MSPS)	Resolution (bits)	Power (mW)	SNDR (dB)	SFDR (dB)	ENOB	DNL/INL (LSB)	Area (mm ²)
[10]	180	1.8	40	10	23.4	59.53	71.67	9.40	0.28/0.32/ 0.62	0.63
[12]	180	1.8	30	10	21.6	57.41	65.93	9.10	0.57/0.8	0.70
[30]	180	1.8/3	80	10	69.0	-	72.78	9.29	+0.2/-0.25 +0.5/-0.5	1.85
[31]	180	1.8	50	10	12.0	56.20	72.70	9.03	0.39/0.81	0.86
[32]	180	1.8	30	10	21.6	58.50	66.10	-	0.3/0.46	1.85
[36]	180	1.8	200	10	56	61.07	-	9.85	-	1.44
Present work	180	1.8	40	10	7.30	56.10	70.00	9.02	+0.57/-1 / +0.7/-0.74	0.35

Table 4 shows the comparison of the 10-bit SHA-less pipelined ADC with some reported pipelined ADCs. Comparing the present work with the reported work at [27] 39.1% of power reduction has been achieved. When compared to the other pipelined ADCs reported in previous reported works this 10-bit SHA less Pipelined ADC has less power consumption.

5 CONCLUSION

A 40MSPS pipelined ADC which is suitable for low power has been described. By removing the front-end SHA, considerable power saving is obtained. The variable g_m operational transconductance amplifier used in the design has been designed with a gain of 45dB. Clocked dynamic comparator is also designed and the same is used to build sub-ADC's which generates the LSB and MSB of the each single stage. Simulation results of the pipelined ADC in a 180nm Silicon based CMOS process shows an SFDR of 70 dB, SNDR of 56.12 dB, ENOB of 9.02 bits and FOM of 0.35pJ/step while consuming only 7.3mW for a 2 MHz input signal. From the results it was shown that the designed ADC maintains a good dynamic performance and low power consumption suitable for SOC Digital TV Application.

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All authors whose names appear on the submission

- 1) made substantial contributions to the conception or design of the work
- 2) drafted the work or revised it critically for important intellectual content
- 3) approved the version to be published
- 4) agree to be accountable for all aspects of the work in ensuring that questions related to the accuracy or integrity of any part of the work are appropriately investigated and resolved.

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Figures

Figure 1

A 10-bit SHA-less Pipelined ADC structure

Figure 2

First and second stages of 10-bit Pipelined ADC

Figure 3

Fully differential OTA

Figure 4

Cascoded gain stage with gain boosting technique

Figure 5

P gain boost amplifier

Figure 6

N gain boost amplifier

Figure 7

Simultaneous capacitor and op-amp sharing technique

Figure 8

Proposed sampling network for the first stage

Figure 9

Timing Diagram for the sampling network

Figure 10

Obtained FFT spectrum of pipelined ADC

Figure 11

The layout of the 10-bit SHA less Pipelined ADC