An Integration-Type High-Speed Analog-to-Digital Converter

KAZUYUKI KONDO, MEMBER, IEEE, AND KENZO WATANABE, SENIOR MEMBER, IEEE

Abstract—An analog-to-digital (A/D) converter consisting of a switched-capacitor integrator, comparator, and control circuit is presented. The conversion process consists of voltage-to-frequency (V/F) conversion to determine the upper M bits of an N -bit representation of an analog input voltage and the subsequent voltage-to-time (V/T) conversion to determine the remaining lower $N - M$ bits. The total clock cycle required for N -bit resolution is $2^M + 2^{N-M}$ at most. The circuits for the V/F and V/T conversion share most of the components and thus the converter can be implemented with the minimum component count. Error analysis shows that a conversion accuracy higher than 12 bits can be expected from its CMOS monolithic realization. Prototype converters built using discrete components have confirmed the principles of operation.

I. INTRODUCTION

DUAL-SLOPE and charge-balancing analog-to-digital (A/D) converters are two typical integration-type converters commonly used in the instrumentation and measurement field [1]. Though different in the principle, i.e., the former is based on a voltage-to-time (V/T) conversion while the latter on a voltage-to-frequency (V/F) conversion, their conversion processes are both insensitive to component variations and also immune to power-line noise if the integration time is made equal to the period of the line frequency [2]–[4]. By virtue of these distinctive properties, excellent linearity and a resolution higher than 14 bits can be easily obtained. They require, however, at least 2^N clock cycles for N -bit resolution, and thus the conversion speed is too slow for many applications. To increase the applicability, an improvement in speed is highly desirable.

A well-acknowledged method for increasing the conversion rate of an integration-type A/D converter is the use of a multislope technique [5]. For 4-digit resolution, a conversion rate as high as 1600 samples/s is possible with this method [6], but a complicated circuit design is inevitable because of the precisely weighted current sources required for the high-speed time measurement.

The improvement in speed is also possible by performing a V/F conversion during the integration period of a dual-slope A/D converter. Based on this idea, an integrating-type high-speed A/D converter has been devel-

oped. The V/F conversion determines the upper M bits of an N -bit binary representation of an analog voltage. The remaining lower $N - M$ bits are determined by the subsequent V/T conversion. The total clock cycle required for N -bit resolution is thus $2^M + 2^{N-M}$ at most. The V/F and V/T converters involved share most of the components and thus the speed improvement is accomplished with a minimum component count.

II. CIRCUIT DESCRIPTION

Fig. 1 shows the circuit diagram of the integration-type high-speed A/D converter. Here, V_x is an analog voltage to be converted into a binary number with reference to the voltage V_r . The whole circuit is divided into three main blocks; the switched-capacitor integrator, the comparator, and the control circuit including M - and K -bit counters. The control circuit generates the digital signals to manage the converter operation. The timing is shown in Fig. 2. ϕ and $\bar{\phi}$ are the nonoverlapping two-phase clocks. ϕ_R is the reset pulse in synchronism with the ϕ clock. After resetting the integrator and the counters, the reset pulse initiates the A/D conversion. The state signal divides the conversion cycle into the two states; the V/F conversion state T_{VF} during which the upper M bits are determined by the charge-balancing technique, and the V/T conversion state T_{VT} during which the lower K bits ($K = N - M$) are determined by the single-slope technique. The operation in each state for unipolar and for bipolar conversion will be described in detail next.

A. Unipolar Conversion

1) *Operation in the T_{VF} State:* In this state, switches S_1 and S_2 are kept "off." The circuit configuration and the control signal for each switch applicable in this state are shown in Fig. 3. Op-amp A_1 and capacitors C_1 , C_2 , and C_3 form the offset-free, parasitic- and gain-insensitive integrator [7], [8]. In the $\phi = "1"$ phase, C_1 is charged, via S_4 , S_8 , and C_3 , to an input voltage V_x . For the moment, V_x is assumed to be positive. In the subsequent $\bar{\phi} = "1"$ phase, the charge stored in C_1 is transferred into C_2 via S_5 and S_6 . The output voltage V_o of the integrator thus increases by $C_1 V_x / C_2$ in every $\bar{\phi}$ clock. This charge accumulation continues until the integrator output V_o becomes positive and the comparator output (CMP) turns "1." When this happens, the M -bit counter is incremented and switch S_3 is activated, instead of S_5 , in the

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K. Kondo is with the Department of Electrical Engineering, Suzuka College of Technology, Shiroko, Suzuka 510-02, Japan.

K. Watanabe is with the Research Institute of Electronics, Shizuoka University, Hamamatsu 432, Japan.

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Fig. 1. The circuit diagram of the integration-type high-speed A/D converter.

Fig. 2. The timing diagram of digital control signals and the output waveform of op-amp A_1 .

Fig. 3. The converter configuration in the T_{VF} state (unipolar conversion).

subsequent $\bar{\phi} = "1"$ phase. Then, the charge $C_1(V_r - V_x)$ is extracted from C_2 and the integrator output becomes negative again, to resume the charge accumulation. This process of charge accumulation and extraction is repeated for 2^M clock cycles. Output voltage waveforms of the integrator and the comparator are also illustrated in Fig. 2.

Assume now that the charge extraction occurred m times during this state. Then, the total reference charge extracted from C_2 is mC_1V_r , while the total signal charge accumulated into C_2 is $2^MC_1V_x$. The charge $Q(t_M)$ residing in C_2 at the end of this state is then

$$Q(t_M) = C_2V_o(t_M) = C_1(2^MV_x - mV_r). \quad (1)$$

Since $|Q(t_M)| < C_1V_r$, we have

$$\begin{aligned} |V_x - (m_12^{-1} + m_22^{-2} + \dots + m_M2^{-M})V_r| \\ < 2^{-M}V_r \end{aligned} \quad (2)$$

where

$$\sum_{i=1}^M m_i2^{M-i} = m. \quad (3)$$

Equation (2) indicates that m_i ($i = 1, 2, \dots, M$), the content of the M -bit counter, is an M -bit binary representation of V_x . To increase the resolution further, the output voltage $V_o(t_M)$ of the integrator is quantized by means of V/T conversion in the subsequent T_{VT} state.

2) *Operation in the T_{VT} State:* In this state, switches S_4 and S_7 are kept "off." The circuit configuration and the control signal driving each switch are shown in Fig. 4. Op-amp A_1 and capacitors C_1 , C_2 , C_3 , and C_4 now form a differential integrator. The comparator, consisting of op-amp A_2 and D flip-flop, monitors the polarity of the integrator output to control the operation of the integrator and the $(M + K)$ -bit up/down counter. If the output voltage $V_o(t_M)$ of the integrator at the end of the preceding T_{VF} state is positive, i.e., $CMP = "1,"$ then the integrator deposits the charge C_1V_r onto C_2 , in every $\bar{\phi} = "1"$ phase, while drawing the charge C_4V_r from C_2 concurrently. The $(M + K)$ -bit counter then operates as an up counter. If $CMP = "0,"$ on the other hand, then the charge C_4V_r is deposited, in every $\bar{\phi} = "1"$ phase, while the charge C_1V_r is drawn concurrently. The counter then operates as a down counter.

Assume now that capacitors C_1 and C_4 consist of 2^K and $2^K + 1$ unit capacitors, respectively, connected in parallel; $C_1 = 2^K C_u$ and $C_4 = (2^K + 1)C_u$, where C_u denotes the unit capacitance. Then the net charge Q_n :

$$Q_n = C_uV_r \quad (4)$$

is deposited onto or drawn from C_2 in every $\bar{\phi} = "1"$ phase, depending on the polarity of $V_o(t_M)$. This process of charge deposit or withdrawal is repeated k times, as illustrated in Fig. 2, until the total quantum charge kC_uV_r counterbalances $Q(t_M)$. Thus we have

$$\begin{aligned} Q(t_M) &= \text{sgn}[V_o(t_M)]kC_uV_r \\ &= \text{sgn}[V_o(t_M)](k_12^{-1} + k_22^{-2} \\ &\quad + \dots + k_K2^{-K})C_1V_r \end{aligned} \quad (5)$$

where

$$\text{sgn}[V_o(t_M)] = \begin{cases} 1, & \text{if } V_o(t_M) \geq 0 \\ -1, & \text{otherwise} \end{cases} \quad (6)$$

and

$$\sum_{i=1}^K k_i2^{K-i} = k. \quad (7)$$

It follows, therefore, that k_i ($i = 1, 2, \dots, K$), the content of the K -bit counter, gives a K -bit representation of $Q(t_M)$.

Fig. 4. The converter configuration in the T_{VT} state (unipolar conversion).

Combining (1) and (5), we can obtain an $N(=M+K)$ -bit representation of an analog voltage V_x :

$$V_x = \left\{ m_1 2^{-1} + m_2 2^{-2} + \cdots + m_M 2^{-M} + \text{sgn}[V_o(t_M)](k_1 2^{-(M+1)} + \cdots + k_K 2^{-(M+K)}) \right\} V_r. \quad (8)$$

The total clock cycle required for the N -bit conversion is $2^M + 2^{N-M}$ at most, which becomes minimum when $M = N/2$. A conventional integration-type converter, on the other hand, requires 2^N clock cycles at least. Therefore, the speed improvement as high as $2^{(N/2)-1}$ is possible with the present architecture.

B. Bipolar Conversion

In the previous description of the converter operation, an input analog voltage V_x was assumed to be positive. Bipolar conversion is also possible by slightly modifying the control signals in the T_{VF} state. The circuit diagram in the T_{VF} state for bipolar conversion is shown in Fig. 5. The timing diagram of the control signals and the output waveforms of the comparator and integrator when $M = K = 3$ are shown in Fig. 6. The analog circuitry is the same as that shown in Fig. 3, but the four-phase clocks (ϕ_a, ϕ_b, ϕ_c , and ϕ_d in Fig. 6) are required for driving each switch.

In ϕ_a and ϕ_b phases, op-amp A_1 operates as a noninverting integrator to accumulate the charge $C_1 V_x$ onto C_2 . Op-amp A_2 and D flip-flop form the comparator to test the polarity of the integrator output V_o . If V_o is positive, then the output of the comparator CMP is set to "1." Otherwise, CMP is set to "0." In ϕ_c and ϕ_d phases, if $CMP = "1,"$ the reference charge $C_1 V_r$ is extracted from C_2 and the M -bit counter is incremented. If $CMP = "0,"$ on the other hand, the reference charge $C_1 V_r$ is accumulated in C_2 . This process of charge accumulation and extraction is repeated for 2^M clock cycles of the four-phase clock.

Assume that the reference charge extraction occurred m times in this state. Then the total extracted charge Q_{ext} is $m C_1 V_r$, while the total charge Q_{dep} deposited upon C_2 is $2^M C_1 V_x + (2^M - m) C_1 V_r$. The charge $Q(t_M)$ remaining in C_2 at the end of this state is then

$$Q(t_M) = C_1 V_o(t_M) = Q_{\text{dep}} - Q_{\text{ext}} = C_1 \{ 2^M (V_x + V_r) - 2m V_r \}. \quad (9)$$

Fig. 5. The converter configuration in the T_{VF} state (bipolar conversion).

Fig. 6. The timing diagram of digital control signals and the output waveform of the integrator (bipolar conversion).

Since $|Q(t_M)| < C_1 V_r$, we have

$$\left| V_x - (m_1 2^0 + m_2 2^{-1} + \cdots + m_M 2^{-M+1} - 1) V_r \right| < 2^{-M} V_r \quad (10)$$

where $m_i (i = 1, 2, \dots, M)$ is defined by (3).

The circuit configurations and operations in the T_{VT} state for bipolar conversion are the same as those for unipolar conversion, except that the charge $Q(t_M)$ can assume any value between $-C_1 V_r$ and $C_1 V_r$, and thus not a K -bit but $(K+1)$ -bit up/down counter is required for quantizing the full scale value $2C_1 V_r$ using the quantum charge $C_u V_r$. Assume now again that the quantum charge is deposited onto or drawn from C_2 k times, depending on the polarity of $Q(t_M)$, as illustrated in Fig. 6, until the total quantum charge $k C_u V_r$ counterbalances $Q(t_M)$. Thus we have

$$Q(t_M) = \text{sgn}[V_o(t_M)] k C_u V_r = \text{sgn}[V_o(t_M)] (k_0 2^0 + k_1 2^{-1} + \cdots + k_K 2^{-K}) \cdot C_1 V_r \quad (11)$$

where

$$\sum_{i=0}^K k_i 2^{K-i} = k. \quad (12)$$

Combining (9) and (11), we have

$$\begin{aligned} V_x &= \left\{ \frac{m}{2^{M-1}} + \text{sgn}[V_o(t_M)] \frac{k}{2^{K+M}} - 1 \right\} V_r \\ &= [m_1 2^0 + m_2 2^{-1} + \cdots + m_M 2^{-M+1} \\ &\quad + \text{sgn}[V_o(t_M)] \{k_0 2^{-M} + k_1 2^{-(M+1)} \\ &\quad + \cdots + k_K 2^{-(M+K)}\} - 1] V_r. \end{aligned} \quad (13)$$

Therefore, m and k give an $(M + K + 1)$ -bit offset binary representation of V_x . The total number of clock cycles required for $(M + K + 1)$ -bit conversion is $2^{M+1} + 2^K$ at most, since one cycle of the four-phase clock corresponds to two clock cycles and $Q(t_M)$ is quantized within 2^K clock cycles. Thus the conversion rate for a bipolar signal is the same as that for a unipolar signal.

III. CONVERSION ACCURACY

The switched-capacitor integrator is configured such that parasitic capacitances and the nonideal factors of op-amp such as the offset voltage and the finite open-loop gain have no effect upon its operation. Therefore, the main error sources of the present A/D converter are clock feedthrough and the inaccuracy of the capacitance ratio between C_1 and C_4 . Their effects upon the conversion accuracy will be estimated separately in this section. Only unipolar conversion is analyzed for simplicity, but the result holds also true for bipolar conversion.

A. Clock Feedthrough

Clock feedthrough accompanying each transition causes the offset charge to appear in C_2 in every cycle. The amount of the offset charge differs in the T_{VF} and T_{VT} states because the circuit configuration is different. For simplicity, however, it is assumed that the feedthrough charge Q_f is injected into node (A) in Fig. 1 in every clock cycle. Then, the offset charge accumulated in C_2 during the whole conversion period amounts to $(2^M + 2^{N-M})Q_f$. If this offset charge is less than the quantum $C_u V_r$, then the N -bit conversion is accurate down to its LSB:

$$2^M(2^M + 2^{N-M}) \leq C_1 V_r / Q_f. \quad (14)$$

This condition is most relaxed when $M = N/2$. The signal-to-noise charge ratio $C_1 V_r / Q_f$ as high as 1×10^4 can be obtained by accommodating the clock feedthrough cancellation scheme [9]. Then, a conversion accuracy higher than 12 bits can be expected.

B. Capacitance Ratio Error

The V/F conversion in the T_{VF} state is independent of the capacitance ratio [4]. Thus it affects only the lower K bits determined by the V/T conversion. Assume now that unit capacitors composing C_1 and C_4 have the mismatch

Fig. 7. The output voltage of op-amp A_1 (upper trace) and the state signal (lower trace) observed in the prototype 8-bit converter. Horizontal scale: 0.1 ms/div; vertical scale: 1 V/div (upper trace), and 20 V/div (lower trace).

ϵ . The maximum error in the quantum charge deposited onto or subtracted from C_2 during the T_{VT} state then amounts to $2^K(2^{(K/2)+1})\epsilon C_u V_r$. If this error is less than one quantum, then the K -bit conversion is accurate down to its LSB. This gives the accuracy condition

$$2^{(K/2)+1}(2^K - 1)\epsilon < 1. \quad (15)$$

The capacitance mismatch ϵ as small as 0.05 percent is attainable with presently available MOS technologies [10], [11]. Therefore, the accuracy of the V/T conversion is estimated to be 7 bits.

Summarizing the above discussions, we can conclude that the overall accuracy of the present converter is determined by clock feedthrough and a 12-bit resolution can be obtained by its CMOS monolithic realization. Choosing $M = K = N/2$ gives the best compromise between the conversion accuracy and speed. If, however, the application allows the slower speed, then decreasing K and increasing M correspondingly would be beneficial in reducing the chip area, because capacitors C_1 and C_4 occupy most of the area.

IV. EXPERIMENTAL RESULTS

To confirm the principles of operation, prototype unipolar and bipolar converters were breadboarded using discrete components. Op-amps and switches used were LF347 and MC14016, respectively. Capacitors were $C_1 = 2.63$, $C_2 = 2.81$, $C_3 = 11.2$, and $C_4 = 2.79$ nF. The reference voltage was set to 2 V.

Fig. 7 shows the integrator output and the state signal observed when the analog voltage $V_x = 0.41$ V was converted into 8-bit ($M = K = 4$) binary number by the unipolar converter. In the T_{VF} state, the reference charge is extracted 4 times (pointed by white arrows) and thus $m = 4$. In the T_{VT} state, the quantum charge is deposited 12 times, and thus $k = 12$. Since $V_o(t_M) < 0$, $(M + K)$ -bit counter operates as a down counter. Therefore, the digital output is 010 000 00-000 011 00 = 001 101 00. This corresponds to the analog voltage $V_x = (52/256) \times 2 = 0.406$ V, agreeing well with the applied voltage.

Fig. 8. The differential nonlinearity error of the 12-bit prototype A/D converter.

Fig. 8 shows the differential nonlinearity error of the 12-bit ($M = 8$, $K = 4$) prototype unipolar converter measured by a static test. The maximum error is less than 0.7 LSB, and thus no missing code appears. A prototype bipolar converter has also confirmed the principles of operation and the accuracy estimate.

The clock rate used in these experiments was 30 kHz. In CMOS realization, the clock frequency higher than 10 MHz can be used, and thus a conversion rate compared favorably with that of a CMOS successive approximation A/D converter is attainable with the present architecture.

V. CONCLUSIONS

An integration-type high-speed A/D converter using a switched-capacitor circuit was presented. This converter features a small device-count integrable onto a small chip area, a high accuracy made possible by the parasitic and gain insensitive configuration, and a conversion rate com-

parable to that of a successive approximation A/D converter. Therefore, it will find wide applicability in the instrumentation and measurement field.

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