Pseudo-Differential Time-Domain Integrator Using Charge-Based Time-Domain Circuits

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Abstract—This work proposes a pseudo-differential time-domain integrator using half-delay time-domain registers and adders relying on charge-based time-domain circuits. It is implemented using a 65-nm CMOS Technology and performs first order integration of time-domain information within the range of [4 ns, -4 ns] across temperature -40°C to 80°C. It consumes 740 μW with a supply voltage of 1.2 V at a 100 MHz clock frequency. A delay-locked-loop (DLL) based foreground calibration is used to compensate for process and temperature variations.

Index Terms—Time-domain amplifier, time-domain adder, time integrator, half-delay, delay-locked-loop (DLL)

I. INTRODUCTION

The down-scaling of modern CMOS technologies has benefited the mixed-mode data converters utilising time-domain signal processing with improved time resolution. Although the converters have achieved fast operation with high throughput, but those using traditional voltage-domain signal processing suffer from an increased $1/f$ noise and reduced dynamic range due to scaled supply voltage. In time-domain signal processing, the circuits use the time-difference or interval between two discrete events to represent and process information. The basic time-domain blocks such as time-domain amplifiers help to improve the resolution by amplifying the time differences whereas the time-domain registers, adders and integrators assist in filtering [1] and data conversion applications [2]–[4]. In popular practice, the time-domain information is integrated in the phase-domain by using voltage-controlled-oscillators (VCOs), but suffers from VCO non-linearity and harmonic distortions. Although these can be reduced by using the VCOs inside $\Delta\Sigma$ loops, the overall performance is limited by the VCO phase noise. The time-domain integration could also be performed using bi-directional gated delay lines (BDGDL) [5] but gets affected from timing skew errors. Alternate approaches use charge-based time-domain amplifiers or adders [6], [7]. There the time-domain information is processed by exponential discharging of capacitors, introducing harmonic distortion as a result of the non-linear discharging profile. On the contrary, the methods using constant-current charging profiles [8] provide better linearity and less harmonic distortion.

II. PROPOSED TIME-DOMAIN DIFFERENTIAL INTEGRATOR

The proposed time-domain differential integrator (Fig. 1) operates by accumulating time-intervals over the sampling

clock cycles similar to first order integration in discrete-time voltage-domain signal processing.

Fig. 1: Overview of the proposed time-domain integrator.

The time-domain integrator is implemented using time-difference detectors (TD), half-delay time-domain adders (TA) and time-domain registers (TR). The TD at the input is used to sample T_{in} and that at the output is used to sample T_{out} and provide feedback to the TA. Now, it can be derived for each clock cycle ‘n’, $T_{out}[n] = bT_{in}[n-1] + aT_{out}[n-1]$, where b and a represents the signal-path gain and feedback-path gain respectively. Then the system transfer function $H(z)$ can be expressed as, $H(z) = bz^{-1}/(1 - az^{-1})$. Following the Forward Euler approximation, it can be written as, $H(s) = b/(1 - a + sT_s)$ where T_s is sampling time period. For an input signal frequency Ω_{in} , the gain and phase of the integrator are $\frac{|b|}{\sqrt{(1-a)^2 + (\Omega_{in}T_s)^2}}$ and $-\tanh \frac{\Omega_{in}T_s}{b(1-a)}$ respectively.

III. CIRCUIT IMPLEMENTATION

A. Half-delay Time-domain Register

The half-delay time-domain register (TR) (Fig. 2(a)) is designed based on a constant-current charging-based time-domain amplifier [8], such that $T_o = T_i z^{-0.5}$ where T_i and T_o are the input and output pulse-time intervals respectively. As shown in Fig. 2(c), T_{CLR} discharges the capacitor C at each rising edge of the CLK . When $CLK=1$, I_1 charges C for the time T_i and thereafter C holds the charge ($\Delta Q_H = I_1 T_i$) deposited. When $CLK=0$, I_1 again starts to charge C . The buffer at the output node of C switches when the voltage across C , V_C crosses V_{TH} , the switching threshold voltage of the buffer. The charge stored during $CLK=0$ till V_C crosses V_{TH} , equals to $\Delta Q_L = I_1(T_{CLK}/2 - T_o)$. Now I_1 is set such that $(I_1/C)(T_{CLK}/2 - T_{CLR}) = V_{TH}$. Then for each clock cycle, $\frac{\Delta Q_H}{C} z^{-0.5} + \frac{\Delta Q_L}{C} = V_{TH}$ and the output T_o can be derived as,

$$T_o = T_i z^{-0.5} + T_{CLR} \quad (1)$$

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Fig. 2: Circuit implementation and timing diagram of the half-delay time-domain register (TR) and time-domain adder (TA)

In case of a pseudo-differential architecture, $\Delta T_o = \Delta T_i z^{-0.5}$. Although ΔT_o is not influenced by the offset (T_{CLR}), it limits the maximum allowable input time difference ($T_{CLK}/2 - T_{CLR}$). For the operation of the TR, $I_{1,min} = CV_{TH}/(T_{CLK}/2 - T_{CLR})$. With $C = 120$ fF, $T_{CLR} = 0.2$ ns and $V_{TH} = V_{DD}/2 = 0.6$ V, the current $I_{1,min}$ equals to $15 \mu A$ for a $T_{CLK} = 10$ ns.

B. Half-delay Time-domain Adder

The half-delay time-domain adder (TA) (Fig. 2(b)) performs an addition of two pulse-time intervals (T_{i1}, T_{i2}) at the input, and produces an T_o with a half clock cycle delay. As shown in Fig. 2(d) when $CLK=1$, C is charged using two matched cascoded current sources (I_1, I_2), enabled by input time intervals T_{i1} and T_{i2} respectively.

For case (i), when T_{i1} overlaps with T_{i2} , the total charge (ΔQ_H) stored in C during sampling can be derived as, $\Delta Q_H = I_1(T_{i1} - T_{ov}) + (I_1 + I_2)T_{ov} + I_2(T_{i2} - T_{ov})$.

For case (ii), without any overlap between T_{i1} and T_{i2} , the charge (ΔQ_H) stored in C can be derived as, $\Delta Q_H = I_1 T_{i1} + I_2 T_{i2}$. As CLK becomes low, I_1 continues to charge C , and the charge stored equals to $\Delta Q_L = I_1(T_{CLK}/2 - T_o)$.

Similar to (1), it can be derived for both the above cases that, $T_o = (T_{i1} + kT_{i2})z^{-0.5} + T_{CLR}$, where $k = I_2/I_1$. For a pseudo-differential architecture with matched current sources ($I_1 = I_2$) the output, $\Delta T_o = (\Delta T_{i1} + \Delta T_{i2})z^{-0.5}$.

C. Time-difference Detector

Fig. 3: Overview of the Time-difference detector (TD)

The time-difference detector (TD) (Fig. 3(a)) comprises 2 D-flipflops with an AND gate driving the RESET in feedback. It produces pulse-time intervals proportional to the time differences between the rising edges of the input signals. The AND-gate delay (τ_d) in feedback reduces the dead-zone and ensures a minimum width (τ_d) pulse at the outputs and thus improves the sensitivity of the circuit towards small ($< 50ps$) input time differences.

Time-domain Differential Integrator: For matched current sources $I_1 = I_2$, the overall system transfer function $H(z)$ can be derived as, $H(z) = z^{-1}/(1 - z^{-1})$ and the gain and the phase of the integrator at the input frequency Ω_{in} equals to $1/(\Omega_{in} T_s)$ and $-\pi/2$ respectively.

IV. CALIBRATION CIRCUIT

Fig. 4: Details of the calibration circuit for bias current

As discussed in section III, the charging current (I_1) for the TRs and TAs need to be adjusted such that $(I_1/C)(T_{CLK}/2 - T_{CLR}) = V_{TH}$ and it varies with changing temperature and different process conditions. In this case, the allowable

range of the input time difference equals to $\pm(T_{CLK}/2 - T_{CLR})$, which is limited by the variation of V_{TH} and I_1 . The calibration circuit operates similar to a DLL and it helps to adjust I_1 to the desired value. As shown in Fig. 4, the circuit comprises a half-delay TR, 16-cell shift register, and a bias circuit with a switchable 4-bit thermometer coded current reference. The off-chip resistor in the current bias generator is used for course tuning to accommodate larger variations across different process conditions.

The TR produces a pulse-time interval ($T_p = C.V_{TH}/I_1$) which is compared with respect to $T_{CLK}/2$ using a D-flipflop. The D-flipflop produces an error output 1, when $T_p + T_{CLR} > T_{CLK}/2$ and vice versa. For a positive error, the shift register shifts right and enables LSBs of the current reference to increase I_1 and reduce T_p to adjust $T_p \rightarrow [T_{CLK}/2 - T_{CLR}]$. It starts with a RST pulse setting the output of the shift register to 8 and thereafter the calibration phase continues for a maximum of 8 clock cycles to adjust I_1 to the desired value. It ends when the output continues to toggle for 1 LSB-bit and hence a maximum $\pm 1/2$ LSB error is expected after calibration.

V. CIRCUIT NON-IDEALITIES

A brief investigation of the impact of the non-idealities is discussed as follows:

1) *Current Mismatch*: For a relative mismatch of ϵ_I between the current sources (I_1, I_2), the feedback-path gain $a = (1 \pm \epsilon_I)$, and then $H(z)$ can be rewritten as, $H(z) = \frac{z^{-1}}{1 - (1 \pm \epsilon_I)z^{-1}}$. Furthermore, the gain and the phase can be estimated as $\frac{1}{\sqrt{\epsilon_I^2 + (\Omega_{in}T_s)^2}}$ and $\pm \tanh \frac{\Omega_{in}T_s}{\epsilon_I}$ respectively. These distortions in the gain and the phase worsen in case of very small input frequencies (i.e, $\Omega_{in}T_s \approx \epsilon_I$).

2) *Non-linear Distortion*: Assuming the PMOSs in the current sources operate in saturation during the charging phase, the cascoded current source can be modeled as a current of $I_{1,sat}$ in parallel with r_o [9]. Then the charging current $I_c(t)$ can be derived as, $I_c(t) = I_{c,sat} \exp(-t/\tau_o)$, where $I_{c,sat} = (VDD/r_o + I_{1,sat})$ and $\tau_o = r_o C$. Now using Taylor series expansion for small inputs ($|T_i[n]/\tau_o| < 1$), the output of the single-ended TR can be approximated from (1) as,

$$T_o[n + 0.5] \approx \left(T_i[n] - \frac{T_i^2[n]}{\tau_o} + \dots \right) \frac{I_{c,sat}}{I_{1,sat}} + T_{CLR} \quad (2)$$

Although, τ_o is large in cascoded configuration, it reduces effectively due to the parasitics in high frequency (100 MHz) applications, resulting in a non-uniform time-domain gain.

3) *Charge Leakage*: The charge injection from the gated PMOS switches, clock feed-through from activating input pulse-time signals generate voltage fluctuations across C . In hold-state, finite impedance (R_{off}) of the discharging NMOS lead to leakage of charge stored in C . This introduces a voltage error (Δv_e) causing a change in the threshold crossing time [9]. The resulting timing error (Δt_e) can be estimated as, $\Delta t_e = \Delta v_e / \text{slewrate} = C \Delta v_e / I_{1,sat}$.

4) *Circuit Noise*: The noise contributions of the cascoded PMOSs (M_{2P}, M_{4P}) are negligible. The thermal noise from the PMOSs (M_{1P}, M_{3P}) and the charging capacitors integrates over C as a wiener process and appear as brown voltage noise at the input of the comparator. As described in [9], [10] the resulting timing error due to thermal noise deposition can be estimated as,

$$\overline{\Delta t_{th}^2} = \frac{kTC}{I_{1,sat}^2} + \frac{2kT\gamma g_m t_{ch}}{I_{1,sat}^2} \quad (3)$$

where $t_{ch} = T_{CLK}/2 - T_{CLR}$ is the total charging time window. Similarly, the timing jitter due to flicker noise can be determined by considering the measurement time $[0, t_m]$ which limits the noise frequencies by the bandwidth $[f_m = 1/t_m, \infty]$. The resulting timing error [10] can be derived as,

$$\overline{\Delta t_{fl}^2} = \frac{K_f g_m^2 t_{ch}^2}{I_{1,sat}^2 C_{ox} W L} \left(\frac{3}{2} - Ci(2\pi f_m t_{ch}) \right) \quad (4)$$

where $Ci(x) = -\int_x^\infty \frac{\cos x}{x} dx$ is cosine integral function.

Total Timing Error: The total timing error due to charge leakage and circuit noise can be estimated as, $\sigma_t^2 = \overline{\Delta t_e^2} + \overline{\Delta t_{th}^2} + \overline{\Delta t_{fl}^2}$, where $\overline{\Delta t_e^2}$ due to charge-leakage is dominant. Consequently the SNR in dB can be expressed as, $SNR = 10 \log_{10} (T_{in,pp}^2 / 8\sigma_t^2)$.

VI. SIMULATION RESULTS

The following results are obtained by analysing post-layout simulations with a supply voltage of 1.2 V and using a veriloga testbench for generating the input time differences. Fig. 5 shows the layouts of the proposed time-domain integrator and the calibration circuit. A symmetric layout of the integrator has been implemented to minimize potential differential offset arising due to mismatch conditions in the differential path.

Fig. 5: The layouts of the proposed integrator and the calibration circuit.

The TD is simulated with respect to different input combinations (T_{i1}, T_{i2}) within the range of ± 4 ns (at TT, 27°C), and as shown in Fig. 6(a), linear transfer characteristics can be observed for each cases. Fig. 6(b) shows the values of the calibrated current (I_1) with estimated error-bars. The calibration circuit adjusts I_1 around 15 μA with an error of $\pm 0.7 \mu A$, which has been verified at 3 different corners (FF, TT, SS) across temperature range of -40°C to 80°C.

The transient behavior of the proposed integrator is studied in Fig. 7(a). It can be observed that the integrator produces triangular output time differences when a square wave time differences ($T_{in,pp} = \pm 1$ ns, $f_{in} = 7.69$ MHz) is applied

Fig. 6: (a) Transfer characteristics of the TA and (b) response of the calibration circuit at different corners (FF,TT,SS).

at the input. In case of a sinusoidally varying input time difference ($T_{in,pp} = \pm 1.5$ ns, $f_{in} = 6.2$ MHz), we can also observe a phase shift of 90° between T_{in} and T_{out} . As indicated in the frequency spectrums (Fig. 7(b)), the integrator produces a gain of 6.14 dB at the input signal frequency ($f_{in} = 6.2$ MHz). As evident from the Fig. 7(b), the non-idealities present in the circuit create harmonics (HD2 = -42 dB, HD3 = -30 dB) at the output. As compared to [7], the degradation in harmonic distortions can be attributed to larger signal bandwidth. However, these can easily be reduced if the time-domain integrator is used as filter in a $\Delta\Sigma$ modulator [8].

Fig. 7: (a) Output of the proposed integrator with a square wave input time differences (top), and a sinusoidal input time differences (bottom), (b) Input (T_{in}) and output (T_{out}) spectrums of the time-domain integrator.

The Nyquist criteria requires a tradeoff between the time-signal amplitude and the system's bandwidth. Larger bandwidth implies smaller sampling window ($T_{CLK}/2$) resulting

in smaller achievable time-signals amplitudes. There, a larger power is spent to reduce the jitter induced time-domain noise and improve the SNR performance. A performance comparison of the proposed time-domain integrator with respect to others reported in literature is illustrated in Table. I.

TABLE I: Performance comparison with other time-domain integrators

References	Tech (nm)	fs (MHz)	Power (μW)	HD2[@ f_{in}] (dB)
ISCAS'16 [6]*	130	25	189.2	-42[@317 kHz]
Elec. Letters'18 [7]*	65	50	152.7	-50.9[@317 kHz]
MWCAS'17 [11]*	130	25	101.6	-22[@415 kHz]
This work [†]	65	100	740	-42[@6.2 MHz]

*schematic simulation results, [†]post-layout simulation results

VII. CONCLUSION

A pseudo-differential time-domain integrator is proposed which performs accumulation of time information using charge-based time-domain register and adders. The differential architecture helps to process bipolar time information and is particularly suited for applications as loop filters in noise shaping $\Delta\Sigma$ TDCs [3], [4] and $\Delta\Sigma$ modulators for sensing applications. The architecture can also easily be extended to realize higher order IIR and FIR filters for time-domain signal processing [1].

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