A 0.13[μm] CMOS 78dB SNDR 87mW 20MHz BW CT [ΔSigma] ADC with VCO-based integrator and quantizer

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In this paper we demonstrate a new technique that eliminates the impact of Kv
11 ENOB in a 20MHz bandwidth.

Voltage-controlled oscillator (VCO) based ADCs have become a topic of great
interest due to the unique and attractive signal-processing properties they
offer in the design of oversampling converters. Assuming a ring-oscillator
structure, the outputs of a VCO toggle between two discrete levels (VDD and
GND) like a CMOS digital gate, enabling simple multi-bit quantization using
D-flip-flops. Since only one VCO output phase transitions at a given sampling
instant while all others saturate to the positive/negative supply, quantization
is robust to flip-flop voltage offsets and metastability, and has guaranteed
monotonicity. Furthermore, the VCO behaves as a CT integrator in that its out-
put phase is proportional to the time integral of the applied control voltage. As
long as it oscillates, the VCO output phase accumulates endlessly, implying
that it is also an integrator with infinite DC gain.

However, the nonlinearity in the VCO’s voltage-to-frequency (Kv) transfer char-
acteristic seriously limits the resolution of VCO-based ADCs (Fig. 9.5.1, top).
In all prior published work, the VCO output frequency is the desired output
variable due to its proportional relationship with the input signal. Therefore,
exercising the full DR of these converters requires the input signal to span the
entire nonlinear Kv transfer characteristic, incurring harmonic distortion and
limiting resolution to less than 8 ENOB [1]. While the circuit in [2] improves
distortion performance by placing a high-gain filter before the VCO and
employing negative feedback, Kv nonlinearity still limits resolution to less than
11 ENOB in a 20MHz bandwidth.

In this paper we demonstrate a new technique that eliminates the impact of Kv
nonlinearity by preserving the integral relationship of the VCO output phase to
the input signal (Fig. 9.5.1, bottom). Leveraging the VCO output phase direct-
ly precludes the need to span the entire nonlinear Kv characteristic since small
perturbations (in the range of 10s of mV) at the tuning node are sufficient to
shift the VCO phase by a substantial amount. Since an open-loop VCO is sens-
sitive to frequency offsets and drift, and easily saturates its phase detector for
large input signals, some form of negative feedback is necessary. Here, a
multibit DAC subtracts the previously quantized phase value from the VCO
input, creating a residue that is integrated during the next clock cycle. This
feedback loop not only allows large signals to drive the VCO without incurring
distortion from Kv, nonlinearity, but also it is a 1st-order CT ΔΣ ADC loop, and
it therefore 1st-order shapes quantization noise.

Although this simple 1st-order architecture is largely free of distortion from Kv,
nonlinearity, a higher-order quantization noise-shaping filter is needed to achieve
>11 ENOB resolution in a 20MHz bandwidth. To that end, a 4th-order
architecture that requires only 3 opamps (a 4th integration is performed by the
VCO) with a 900MHz sampling rate (OSR=22.5) and a 4b quantizer/DAC, is
chosen (Fig. 9.5.2). A feedback loop delay compensation scheme similar to
that in [3] is also adopted here. But since the quantizer can only be
stimulated by the integrated voltage-to-phase signal, the compensation is
implemented using a digital 1st-order difference circuit (to cancel the integra-
tion) and an RZ DAC. The highly digital nature of the VCO integrator/quantizer
is observed in Fig. 9.5.3. The VCO delay element is based on a current-starved
inverter, and enables pseudo-differential control as well as frequency and Kv
tuning to cover process variations. A sense-amplifier flip-flop (popular in dig-
ital memory designs) quantizes the VCO output phase. Phase detection and
1st-order difference computation are achieved using static CMOS XOR gates and
TSPC flip-flops.

Opamp-RC integrators are chosen for their linearity and ability to drive resis-
tive loads. A modified nested-Miller topology [4] is implemented, and achieves
moderate gain (>60dB) and wide unity-gain bandwidth (>3GHz) for fast set-
ing. Since its noise appears directly at the input, the 1st opamp consumes the
most current (15mA), while the 2nd and 3rd opamps consume half as much.
Each integrator includes a fixed capacitance and a 5b binary-weighted capac-
itor bank to enable tuning over the combined RC process variation of ±40%.

Current-steering NRZ and RZ DACs are implemented for their fast switching
speeds. Unit-element mismatch errors in the main NRZ feedback DAC appear
directly at the input of the converter and can seriously degrade SNDR. Errors
from the minor-loop NRZ and RZ DACs are suppressed by the preceding loop
filter gain, but can still limit performance when >11 ENOB resolution is
desired. Therefore, DEM must be performed on all feedback DACs, as indicat-
ed by the shaded regions in the system schematic of Fig. 9.5.2. Fortunately, it
is shown in [5] that when a ring VCO’s quantized output phases are 1st-order
differenced, a thermometer code is generated that is equivalent to a 1st-order
dynamic-weighted-averaging (DWA) sequence. Consequently, if these ther-
mometer bits are tied directly to the RZ DAC unit elements, then the DAC mis-
matches are 1st-order shaped (Fig. 9.5.4, top). DWA need only be explicitly
performed on the undifferenitiated thermometer code that drives the NRZ DACs,
a computation that has a full sample period to complete (Fig. 9.5.4, bottom).

The architecture in Fig. 9.5.3 is fabricated in a 0.13um CMOS process, occu-
pying an active area of 0.45mm² (Fig. 9.5.7) and dissipating 87mW from a
1.5V supply, with the analog and digital supplies drawing roughly 46mA and
12mA, respectively. Simulations indicate that the DWA logic consumes more
than 75% of the digital power due to the use of current-mode logic to meet
timing margins, with the VCO phase quantizer flip-flops and clock genera-
distribution/circuitry comprising the remainder. The SNR and SNDR ver-
sus input amplitude curves (Fig. 9.5.5) are calculated over a 20MHz input
bandwidth. For a 2MHz -2.4dBFS input tone, the ADC achieves a peak SNR of
81.2dB, and a peak SNDR of 78.1dB (12.7 ENOB, FOM ~330 fJ/conversion-
step). A table summarizing the ADC performance is shown in Fig. 9.5.6.

An FFT of the ADC output with a 2MHz input signal at -2.4dBFS is shown in
Fig. 9.5.5. The peak SNR is 3dB lower than anticipated due to a slight increase
in the noise floor in the 5-to-20MHz range, and the peak SNDR is 3dB lower
than the peak SNDR due to the presence of even and odd-order distortion tones
(present even with no input power). Simulations suggest that these errors are
caused by mismatches in the main NRZ DAC unit-current elements and
switching transients (ISI). Varying the DAC switch buffer’s voltage supply pro-
vides evidence of ISI, as perturbing the supply worsens in-band distortion due
to larger clock/data charge injection and slower switching transients. While
results indicate that SNDR is improved with better matching and a lower-ISI
DAC structure, they more importantly demonstrate that the architecture is
robust to Kv, nonlinearity, and that the VCO-based ADC is viable for high-per-
formance (>12 ENOB, 20MHz bandwidth) applications.

References:
Sigma-Delta ADC With a 5-bit, 950-MS/S VCO-Based Quantizer,” IEEE J. Solid-State
with 20-MHz Signal Bandwidth, 80-dB Dynamic Range, and 12-bit ENOB,” IEEE J. Solid-
Figure 9.5.1: Block diagrams of previously reported voltage-to-frequency VCO-based ADCs [1] (top) and the proposed voltage-to-phase VCO-based ADC.

Figure 9.5.2: 4th-order CT $\Delta\Sigma$ VCO-based ADC schematic. The 4th integration is performed by the VCO integrator, and hence only 3 opamps are used.

Figure 9.5.3: 3b example of the VCO integrator and quantizer circuitry, associated phase detector and 1st-order difference logic.

Figure 9.5.4: Implicit DWA from VCO quantization and differentiation (top), and explicit DWA implementation (bottom).

Figure 9.5.5: Measured SNDR/SNR over a 20MHz bandwidth while sweeping input amplitude (left), and a 10,000-point FFT for a 2MHz -2.4dBFS input tone (right).

Figure 9.5.6: Summary table of results.
Figure 9.5.7: Die micrograph of the ADC implemented in 0.13µm CMOS. The total active area is 0.45mm². DAC1 corresponds to the main NRZ feedback DAC, and DAC2 and DAC3 correspond to the minor-loop NRZ and RZ feedback DACs.