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An 8-GS/s 200-MHz Bandwidth 68-mW $\Delta\Sigma$ DAC in 65-nm CMOS

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Abstract—This paper presents an 8-GS/s, 12-bit input $\Delta\Sigma$ DAC with 200-MHz bandwidth in 65-nm CMOS. The high sampling rate is achieved by a two-channel interleaved MASH 1-1 digital $\Delta\Sigma$ modulator with 3-bit output, resulting in a highly digital DAC with only seven current cells. The two-channel interleaving allows the use of a single clock for both the logic and the final multiplexing. This requires each channel to operate at half the sampling rate, which is enabled by a high-speed pipelined MASH structure with robust static logic. Measurement results show that the DAC achieves 200-MHz bandwidth, 26-dB SNDR and -57-dBc IMD3, with a power consumption of 68-mW at 1-V digital and 1.2-V analog supplies. This architecture shows potential for use in transmitter baseband for wideband wireless communication.

Index Terms—Digital $\Delta\Sigma$ modulator (DDSM), digital-to-analog converter (DAC), time-interleaving, oversampling, MASH.

I. INTRODUCTION

OVERSAMPLED $\Delta\Sigma$ DACs have been reported in all-digital [1], “almost” digital [2] and RFDAC [3], [4] transmitters for UMTS, WLAN and WiMAX standards. Their use has been driven by the goal of a software defined radio wherein the bulk of the processing is moved into the digital domain in order to reduce the analog complexity and enable easy reconfiguration. The digital-to-analog conversion is pushed closer to the antenna and hence operates at a higher frequency compared to traditional transmitters. $\Delta\Sigma$ DACs thus become attractive in digital-centric transmitters because they greatly reduce the number of analog cells required compared to Nyquist DACs by using digital processing. Secondly, the oversampling relaxes the order of the reconstruction filter after the DAC. The example in Fig. 1 shows the difference between a traditional and a digital baseband transmitter that uses a $\Delta\Sigma$ DAC. It can be seen that the $\Delta\Sigma$ DAC, with much lesser number of current cells than a Nyquist DAC, relaxes the analog matching requirements. It extends the digital-analog boundary closer to the up-conversion mixer and operates at a frequency that depends on the radio standard, signal bandwidth and the required oversampling ratio (OSR).

High-speed high-data-rate communication requires increasingly larger bandwidths, such as the recent wideband standards for UWB [5] and 60-GHz [6] radios that have large RF

Fig. 1. Example of a Nyquist DAC in a traditional transmitter (top) and a $\Delta\Sigma$ DAC in a digital baseband transmitter (bottom). Only the I path is shown.

Fig. 2. Conceptual diagram of a M-channel interleaved DDSM $\Delta\Sigma$ DAC with sampling rate F_s . Each modulator operates at F_s/M and the MUX at F_s .

bandwidths of 528 MHz and 1.7 GHz per channel respectively. However, the bandwidth of $\Delta\Sigma$ DACs reported so far has been limited to 100 MHz (i.e. 200-MHz RF bandwidth when used in I & Q paths of transmitters) [3]. Aiming to improve the bandwidth of $\Delta\Sigma$ DACs for wideband communication, this paper presents a $\Delta\Sigma$ DAC with 200-MHz bandwidth in 65-nm CMOS.

OFDM modulation is commonly used for transmission at high data rates. If an OFDM signal is used with a modulation scheme like QPSK, 16-QAM or 64-QAM, then a SNR in the range of 25-35 dB is required from the DAC [3], [7]. Arriving at such an SNR for bandwidths over 200 MHz requires DAC sampling rates exceeding 8 GS/s. The speed of the Digital $\Delta\Sigma$ Modulator (DDSM) in the DAC then becomes the limiting factor for achieving this sampling rate when a conventional implementation is used.

Time-interleaved DDSM-based DACs are, therefore, required in order to relax the input data rate and the speed of the logic in the modulator. Fig. 2 shows this concept for a M-channel interleaved DDSM DAC. The digital processing contains M modulator sections operating at a relaxed $1/M^{\text{th}}$ of the desired sampling rate. The multiplexing of the channels just before the DAC is still at the full sampling rate [8].

Eight-channel interleaved DDSMs that operate at 2.5 GHz

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Fig. 3. (a) Eight-channel traditional multiplexing [9]. (b) Four-channel multiplexing with phase rotators and dual current cells [12]. (c) Two-channel multiplexing in this work.

have been previously described in [9] and [10]. However, the digital modulator in [9] does not address the DAC design, and in [10], the time-interleaving of the $\Delta\Sigma$ DAC is demonstrated by simulations. Furthermore, in [9] and [10], interleaving has been used with the aim of relaxing the logic speed, while its impact on the final full-speed multiplexing has not been considered due to the moderate speed of 2.5 GHz. However, for the >8 GHz speed targeted in this work, a high degree of interleaving complicates the routing and more importantly, it increases the timing complexity of the multiplexing which then requires accurate multi-phase clock generation. Therefore, a two-channel interleaving strategy is utilized in this work which enables a simplified multiplexer by allowing the use of a single clock for the logic as well as the multiplexing, while pushing the speed of the logic in each channel to 4 GHz.

This paper is organized as follows. Section II motivates the architecture choice. Section III describes the circuit design of the proposed $\Delta\Sigma$ DAC. The test chip and measurement results are presented in Section IV followed by the conclusion in Section V.

II. $\Delta\Sigma$ DAC ARCHITECTURE

A MASH architecture based $\Delta\Sigma$ DAC is suited for high-speed implementation because it is inherently stable, has a short integrator critical path and easily allows pipelining as compared to other typical modulator architectures [9]. Matlab simulations show that a MASH 1-1 DDSM DAC with seven thermometer-coded current cells (3-bit output) can provide a SNDR of 45-dB at 200 MHz bandwidth and 8-GS/s (OSR = 20) with a current mismatch (σ) of 3.6%. A second order modulator presents only a moderate filtering complexity for suppressing the generated out-of-band quantization noise [2], [3]. For the chosen 1-1 MASH modulator, simulations show that a second order low pass filter can satisfy the spectral mask for the UWB standard in [5]. With these relaxed analog constraints, this becomes mainly a digital problem of designing a DDSM for such a high speed. For a conventional MASH 1-1 DDSM that uses static CMOS logic, the 8 GHz speed is found to be outside the capability of a standard 65 nm technology node at 1 V even when extensive pipelining with 1-bit integrators is utilized [4]. Hence, a time-interleaved MASH

Fig. 4. Proposed two-channel interleaved second order MASH $\Delta\Sigma$ DAC with a $2F_s$ sampling rate. Critical path is enclosed by a dashed rectangle.

DDSM is necessary to relax the critical path and the clock rate.

In the 2.5 GHz DDSM described in [9], eight-channel interleaving was chosen with the aim of relaxing the timing of the logic while minimizing the area. The multiplexer (MUX) was not a bottleneck at this speed and was easily achieved using a standard parallel-to-serial conversion. Fig. 3(a) shows this commonly used scheme where the eight individual channels working at an $F_s/8$ speed are retimed to F_s . It can be noted that the clock divider and channel select logic still operate at the full sampling rate, F_s . At 8 GHz, this multiplexing scheme becomes a bottleneck as meeting the clock divider and the channel select logic paths becomes a challenge with full-swing CMOS logic. In order to use this scheme at 8 GHz, a reduced-swing logic is required instead, e.g. CML. However, the reduced swing in the DAC switch drivers would then affect the DAC linearity [11].

Another multiplexing style (Fig. 3(b)), described in [12] for four channels, uses CML and analog phase rotators for setting the clock phases. In this MUX, the final latches are moved into the current cell itself and two current cells per bit are used. This scheme requires a modified current cell and a complicated routing with three times the number of wires going into the cell. The number of current cells also doubles, resulting in an overall increase of analog complexity.

The two MUX choices show that the increased relaxation in the logic timing resulting from an increased degree of interleaving comes at a cost of complicated MUX and current cell circuits. Hence, two-channel interleaving (Fig. 3(c)) is chosen in this work by giving a higher priority to the simplification of the MUX and the current cells. This two-channel approach has the following benefits. Firstly, the logic and the MUX now work with a single half-rate clock and secondly, it allows multiplexing with rail-to-rail swing and use of traditional current cells. However, this requires the logic in each channel to operate at 4 GHz, which is still a challenging task. This speed has been achieved by a careful integrator design that is tailored for the interleaved structure and is described in the following section.

Fig. 5. Design of first order interleaved MASH DDSM. Each 2-bit integrator slice uses four 1-bit carry select adders and transmission gate FFs. Two critical paths exist here.

III. $\Delta\Sigma$ DAC CIRCUIT DESIGN

A. DDSM design

Fig. 4 shows the block diagram of the proposed 12-bit $\Delta\Sigma$ DAC that uses the two-channel interleaved MASH 1-1 architecture and seven current cells. Inputs $x_0[11:0]$ and $x_1[11:0]$, form the two half-rate channels and are the divided even and odd streams of the original data-rate. The 10-bit back-to-back integrators/adders, i.e. the feedback path in each of the first order MASH DDSMs form the critical path in the design (shown with a dashed rectangle). The carry generated from the integrators is added to the remaining two MSB input bits, $x_0[11:10]$ and $x_1[11:10]$. Being in a forward path, this addition and also the end processing are not critical paths. Therefore, the main design problem becomes that of two 10-bit back-to-back adders that must operate at over 4 GHz (250 ps cycle time).

The aim of this work has also been to enable the high sampling rate by utilizing only robust static logic so as to avoid the lower noise margins that are associated with faster logic styles such as pre-charged domino, ratioed or pass transistor logic. Pipelining is essential for meeting the speed requirement. Furthermore, in order to maximize the speed of the pipeline and the number of bits that can be accommodated in each pipeline, an important observation about the two adders can be made. In the first adder (Integrator 0), both the carry chain and the sum delays are in the critical path but in the second adder (Integrator 1), only the carry chain is in the critical path. Hence, using the same adder cells for both the adders does not result in the fastest speed. Using this observation, all the non-critical nodes of both the adders are slowed down by reducing their drive strength. This helps to reduce the capacitance on the critical path and speed it up.

Fig. 5 shows the implementation of a first order two-channel interleaved MASH stage that is instantiated twice in the design. Only static CMOS logic and Transmission Gate Flip-flops (TGFF) are used. Each MASH consists of

TABLE I
PATH 1 AND 2 DELAY BREAKDOWNS. THE DELAYS ARE POST-LAYOUT SIMULATED AT TYPICAL CORNER, 1 V, 75° C AND 4.5 GHz.

Path 1			Path 2		
Circuit	Delay Type	Delay (ps)	Circuit	Delay Type	Delay (ps)
$y_1[0]$ FF	clk $\rightarrow$ q	40	$y_1[0]$ FF	clk $\rightarrow$ q	40
Adder A1	b $\rightarrow$ co	60	Adder A1	b $\rightarrow$ sum	63
Adder A2	ci $\rightarrow$ sum	27	Adder A3	a $\rightarrow$ co	44
Adder A4	a $\rightarrow$ sum	41	Adder A4	ci $\rightarrow$ sum	19
Reset NOR	in $\rightarrow$ out	27	Reset NOR	in $\rightarrow$ out	27
$y_1[1]$ FF	T_{setup}	24	$y_1[1]$ FF	T_{setup}	24
Total Path 1 Delay (ps)		219	Total Path 2 Delay (ps)		217

Fig. 6. Pseudo-differential clock driver used for local distribution. One clock driver is used per 18 FFs.

five pipeline stages of 2-bit integrators and one 3-bit forward pipeline stage. Each 2-bit integrator pipeline uses four 1-bit carry select adders. Adders A1-A2, together form Integrator 0 while A3-A4 form Integrator 1. Based on the observations about the adders mentioned previously, all the four carry select adders are sized differently. There are two critical paths in each pipeline stage, from $y_1[0]$ FF $\rightarrow$ A1 $\rightarrow$ A2 $\rightarrow$ A4 $\rightarrow$ NOR $\rightarrow$ $y_1[1]$ FF (Path 1) and $y_1[0]$ FF $\rightarrow$ A1 $\rightarrow$ A3 $\rightarrow$ A4 $\rightarrow$ NOR $\rightarrow$ $y_1[1]$ FF (Path 2). Table I shows the post-layout delay contributions from both the paths at the typical corner, 1 V supply and 75°C. The table shows that this optimized 2-bit structure supports a frequency greater than 4 GHz and the delays in the two paths are nearly equalized.

Adder A1 is the slowest in the path as it generates both the *sum* and *co* (carry) complementary outputs with equal delay. It also internally generates the complementary inputs, which slows it by one additional inverter delay. In comparison, A2 has a faster *ci* $\rightarrow$ *sum* delay as its *ci* $\rightarrow$ *co* delay is non-critical. On the same lines, A3 has a faster *a* $\rightarrow$ *co* delay as its *a* $\rightarrow$ *sum* delay is non-critical. A4 equalizes both the *co* and *sum* delays similar to A1, but is faster due to its lower fan-out. The NOR gate at the end of the adder path is an additional overhead required to synchronously reset the integrator at start-up.

B. Clock Distribution

In order to keep the TGFFs compact in size by avoiding a clock inverter inside the flip-flops, both the clock phases are provided from outside the FF. While the global clock distribution is single-ended, the pseudo-differential clock driver of Fig. 6 distributes the clocks with a 30 ps slope locally to the FFs. Each such clock driver is used for every 18 FFs (fan-out of 3) in a $70\mu\text{m} \times 45\mu\text{m}$ area. This driver also minimizes the clk-clk overlap as both the clock phases have an equal load.

Fig. 7. Final MUX and DAC current cell with the timing diagram. Switch driver circuit is the same as the local clock driver of Fig. 6.

C. Final Multiplexer and Current Cell Design

Fig. 7 shows the static 2:1 MUX per cell. The data from the second channel is shifted by half a cycle prior to the MUX. The MUX is single-ended and the complementary DAC switch signals are generated with a 15 ps slope after the MUX by a center-crossing pseudo-differential switch driver. The switch driver uses the same circuit as the local clock driver of Fig. 6. A high-crossing switch driver [11] that is modified for this MUX structure is required in the DAC, but such a driver is a challenge at 8 GS/s due to the high capacitance and contention at its cross-coupled nodes. Hence, the center-crossing driver is chosen to meet the speed and fast slope. The DAC is sensitive to a mismatch in driver rise and fall delays and requires a careful driver design. However, due to its pseudo-differential nature, a 3 ps mismatch is the smallest that could be achieved in this design and this is found to degrade the SNDR by 4 dB in post-layout simulation. The clock to the MUX bypasses the main clock distribution and uses a minimized buffering of four stages. The MUX is sensitive to the clock duty cycle and simulations show an SNDR reduction of 4 dB for every 1% variation from the desired 50% duty cycle.

The DAC current cell, also shown in Fig. 7 is designed for 0.3 $V_{pp-diff}$ swing with a 100 Ω passive load and 1.2 V supply. The current source dimensions for the 3.6% mismatch requirement are derived using foundry matching parameters. The current cell has M4-M5 transistors as the cascode pair on top of the switching M2-M3 pair that is biased in the linear region and M1 as the current source. The seven current cells are laid out in a single column with dummies on either side to simplify the route matching between the switch driver and the switches.

IV. MEASUREMENT RESULTS

The proposed $\Delta\Sigma$ DAC is implemented in a standard 65 nm CMOS process and uses only general purpose low- V_T (DDSM & DAC) and standard- V_T devices (DAC). Fig. 8 shows the chip photograph. The active area of the DAC is 0.13 mm². The test chip is directly wire-bonded to an FR4 PCB.

Fig. 9 shows the measurement setup used. A 12-bit input of 8192 sample length and frequency F_{in} is sent in to the chip

Fig. 8. Chip photograph of the implemented $\Delta\Sigma$ DAC.

Fig. 9. Measurement setup for the $\Delta\Sigma$ DAC with the expected spectrum at output of every block. An upsampling filter is not used to simplify testing. Upsampling images in the output are out of the band of interest.

at a rate F_{bb} using a Tektronix AWG5012C pattern generator. Internal to the chip, this data is upsampled to F_s rate by a zero-order hold operation. In addition, the two DDSM channel inputs are also shorted together. The zero-hold operation and the shorted inputs, together results in upsampling of the input data by $2F_s/F_{bb}$ since the DAC effectively works at a $2F_s$ sampling rate. A FIR low-pass filter is not implemented to remove the upsampling images at the DDSM input. With this simplified setup, these unfiltered images appear at the DAC output also, but they lie outside the band of interest. In a real application, however, the upsampling images are filtered out prior to the DDSM. By keeping $F_{in} \leq F_{bb}/4$, the unfiltered images do not result in intermodulation products in the band of interest (DC to F_{in}). The duty cycle of clock F_s is tuned off-chip before the measurements.

Fig. 10 shows the measured single-ended output spectrum of a 200-MHz full-scale single tone input at 8 GS/s operation of the $\Delta\Sigma$ DAC. With $F_{bb} = 800$ MHz and $F_s = 4$ GHz, the main 200 MHz tone and the 9 expected images are seen along with the noise shaping, thus demonstrating the desired DAC operation. The DAC achieves 26-dB SNDR for the 200 MHz single tone input consuming 68 mW (40 mW-clocking, 23 mW-logic/FF, 5 mW-analog) from 1-V digital and 1.2-V analog supplies. The measured IMD3 is -57 dBc (Fig. 11) for two -6 dBFS tones near 200 MHz placed 2 MHz apart.

Fig. 12 shows the simulated DAC output spectrum which has a SNDR of 42 dB. The measured SNDR is found to be lower than the simulated value primarily due to a 10 dB loss resulting from a test setup limitation of synchronizing the F_s and F_{bb} clocks. The two clock sources have been locked to a common 50 MHz reference, but they are still not truly synchronous. This results in an upsampling uncertainty in the

Fig. 10. Measured single-ended spectrum showing 8 GS/s operation with $F_s=4$ GHz, $F_{bb}=800$ MHz and input frequency, $F_{in}=200$ MHz. The noise shaping and the 9 out of band images can be seen.

Fig. 11. Measured -57-dBc IMD3 with two -6 dBFS tones near 200 MHz spaced 2 MHz apart.

zero-hold operation and restricts the SNDR. The higher noise floor up to 800 MHz in the measured spectrum results from this synchronization problem. Table II shows the comparison of this work with previous $\Delta\Sigma$ DACs having >2.5 GHz sampling rate.

V. CONCLUSION

This paper has presented a 8-GS/s 200-MHz bandwidth $\Delta\Sigma$ DAC with 26-dB SNDR, -57-dB IMD3 and 68-mW power consumption in 65 nm CMOS. The high sampling rate was achieved by a two-channel interleaved 4-GHz MASH 1-1 modulator structure. This allowed a single clock solution, thus simplifying the timing complexity of the final full-rate multiplexing. Using only seven current cells with relaxed matching requirements, this work demonstrates the potential of this predominantly digital DAC for use in baseband of transmitters for wideband wireless communication, e.g. UWB and 60-GHz radios.

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REFERENCES

- [1] A. Frappe, A. Flament, B. Stefanelli, A. Kaiser and A. Cathelin, "An All-Digital RF Signal Generator Using High-Speed $\Delta\Sigma$ Modulators", *IEEE J. Solid State Circuits*, vol. 44, pp. 2722-2732, Oct. 2009.

Fig. 12. Output spectrum obtained from simulation.

TABLE II
COMPARISON WITH $\Delta\Sigma$ DACs HAVING >2.5 -GS/S SAMPLING RATE.

Paper	[2]	[4]	[1]	[3]	This Work
DAC Type	$\Delta\Sigma$ DAC	$\Delta\Sigma$ RFDAC ¹	$\Delta\Sigma$ DAC	$\Delta\Sigma$ RFDAC	$\Delta\Sigma$ DAC
Modulator Type	Error Feedback	MASH	Error Feedback	MASH	Interleaved MASH
Technology	90nm	65nm	90nm	0.13 μ m	65nm
Input Bits	10	5	13	12	12
Output Bits	3	3	1	3	3
Order	2	3	3	2	2
Sampling Rate (GS/s)	3.6	5.4	4	2.6	8
Bandwidth (MHz)	10	10	50	100	200
SNDR (dB)	70	-	53	30	26
IMD3 (dBc)	-70	-	-	-51	-57
Area (mm ²)	0.06*	-	<0.15	<0.11*	0.13
Power (mW)	16	>50*	54*	40*	68
Swing (V _{pp-diff})	0.3*	-	1.3	0.35	0.3

* Estimated. ¹ [4] is a hybrid $\Delta\Sigma$ and Nyquist RFDAC.

- [2] P. Seddighrad, A. Ravi, M. Sajadieh, H. Lakdawala and K. Soumyanath, "A 3.6GHz 16mW $\Delta\Sigma$ DAC for a 802.11n / 802.16e transmitter with 30dB digital power control in 90nm CMOS", *Proc. IEEE European Solid-State Circuits Conf.*, 2008, pp. 202-205.
- [3] A. Jerng and C. Sodini, "A Wideband $\Delta\Sigma$ Digital-RF Modulator for High Data Rate Transmitters", *IEEE J. Solid State Circuits*, vol. 42, pp. 1710-1722, Aug. 2007.
- [4] A. Pozsgay, T. Zounes, R. Hossain, M. Boulemlakher, V. Knopnik and S. Grange, "A Fully Digital 65nm CMOS Transmitter for the 2.4-to-2.7GHz WiFi/WiMAX Bands using 5.4GHz $\Delta\Sigma$ RF DACs", *IEEE. Int. Solid-State Circ. Conf. Dig. Tech. Papers*, 2008, pp. 360-619.
- [5] *High Rate Ultra Wideband PHY and MAC Standard*, ECMA-368, Dec. 2008 [Online]. Available: www.ecma-international.org.
- [6] *WirelessHD Specification Version 1.1 Overview*, May. 2010 [Online]. Available: www.wirelesshd.org.
- [7] Q. Gu, *RF System Design of Transceivers for Wireless Communication*, Springer, 2005.
- [8] S. Balasubramanian *et al.*, "Systematic Analysis of Interleaved Digital-to-Analog Converters", *IEEE. Trans. Circuits Systems-II:Express Briefs*, vol. 58, pp. 882-886, Dec. 2011.
- [9] P. Madoglio *et al.*, "A 2.5-GHz, 6.9-mW, 45-nm-LP CMOS, $\Delta\Sigma$ Modulator Based on Standard Cell Design With Time-Interleaving", *IEEE J. Solid State Circuits*, vol. 45, pp. 1410-1420, July 2010.
- [10] J. Pham and A. C. Carusone, "A Time-Interleaved $\Delta\Sigma$ -DAC Architecture Clocks at the Nyquist Rate", *IEEE. Trans. Circuits Systems-II:Express Briefs*, vol. 55, pp. 858-862, Sept. 2008.
- [11] C. Lin *et al.*, "A 12 bit 2.9 GS/s DAC With IM3 <-60 dBc Beyond 1 GHz in 65 nm CMOS", *IEEE J. Solid State Circuits*, vol. 44, no. 12, pp. 3285 -3293, Dec. 2009.
- [12] J. Savoj, A. Abbasfar, A. Amirkhany, M. Jeeradit, and B. Garlepp, "A 12-GS/s Phase-Calibrated CMOS Digital-to-Analog Converter for Backplane Communications", *IEEE J. Solid State Circuits*, vol. 43, no. 5, pp. 1207 -1216, May 2008.