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A Wide-Band 280-MHz Four-Path Time-Interleaved Bandpass Sigma-Delta Modulator

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Abstract—This paper describes a 0.35- μm CMOS fourth-order bandpass analog-digital sigma-delta ($\Sigma\Delta$) modulator for wide-band base stations receivers. The modulator, based on a time-interleaved four-path architecture, achieves an equivalent sampling frequency of 280 MHz, although the building blocks operate at only 70 MHz. In measurements, the prototype chip achieves a dynamic range of 72 dB (12 bits of resolution) with a signal bandwidth of 4.375 MHz centered around an intermediate frequency of 70 MHz. The measured spurious-free dynamic range is 69 dB. The $\Sigma\Delta$ modulator dissipates 480 mW from a 3.3-V supply, including voltage reference buffers and output pads with high-driving capabilities, and occupies 20 mm² of silicon area.

Index Terms—Calibration technique, time-interleaved ADC, sigma-delta ($\Sigma\Delta$) modulator.

I. INTRODUCTION

THE TREND in the receiver chains used in modern telecommunication architectures is to move the analog-to-digital converter (ADC) as close as possible to the antenna. This allows performing most of the signal processing in the digital domain, thus gaining in reconfigurability and programmability. A typical superheterodyne architecture requires to directly digitize the intermediate frequency (IF). The resulting specifications for the integrated ADC are challenging, especially in terms of speed, signal-to-noise ratio (SNR), dynamic range (DR), and spurious-free dynamic range (SFDR) [1]. System-level studies show that an optimal receiver uses IF frequencies around several tens of megahertz (typically 60–80 MHz) [2], [3]. With a conventional approach, this leads to the use of a clock frequency for the ADC significantly higher than the IF (for example, twice the IF, i.e., 120–160 MHz), relying on the fast

Fig. 1. Block diagram of a time-interleaved ADC.

response of state-of-the-art CMOS processes, or on the use of bipolar or BiCMOS technologies. However, the use of these technologies increases costs and makes the design of the basic blocks of the converter extremely difficult.

In the literature, solutions based on a double-sampling technique with two nonoverlapping clock phases have been reported [4], [5], which achieve good performance. However, the speed requirements for their building blocks is a critical issue.

The approach used in this paper is the use of time-interleaving. Several ADCs operate in parallel, using different clock phases [6], as shown in Fig. 1. An analog demultiplexer selects sequentially each ADC, which therefore operates at the low speed. The digital multiplexer interleaves the digital output of the ADCs, thus producing the overall A/D conversion result. Any type of ADC can be used, including sigma-delta ($\Sigma\Delta$) modulators [7]. Each of them operates at a clock frequency f_{ck}/N , where f_{ck} is the overall sampling frequency and N is the number of channels (or paths) used. The speed requirements for each converter are therefore relaxed by a factor N . This is achieved at the expense of spurious tones due to offset and gain mismatches among the different ADCs. This tones if necessary can be canceled with suitable calibration techniques [8].

In this kind of application, a bandpass ADC turns out to be more effective than its low-pass counterpart. With a low-pass ADC the requirement to use an IF of several tens of megahertz while the signal band is only few megahertz around the IF would result in a waste of resources. Also, the use of high-speed converters with a low-pass transfer function extended over high Nyquist zone is not effective. By contrast, a bandpass transfer function centred at the desired IF enables converting into the digital domain only the band around the IF and reduces the in-band thermal noise requirements (the thermal noise outside the signal band is rejected in the digital domain).

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Fig. 2. (a) Four-path $\Sigma\Delta$ modulator and (b) its NTF.

This paper describes the design of a four-path fourth-order bandpass $\Sigma\Delta$ modulator whose performance satisfies the requirements of a typical 3G multi-carrier wireless receiver [9], [10]. The signal bandwidth of the modulator is 4.375 MHz centered around an IF of 70 MHz. The overall clock frequency is 280 MHz, while each path works at 70 MHz. The paper is organized as follows. Section II describes the four-path $\Sigma\Delta$ modulator. Section III discusses the switched-capacitor (SC) implementation, while details on the design of building blocks are presented in Section IV. Finally, Section V reports experimental results obtained from an integrated prototype.

II. MULTI-PATH $\Sigma\Delta$ MODULATOR

$\Sigma\Delta$ modulators used in a multi-path architecture obtain a bandpass response by using low-pass $\Sigma\Delta$ modulators in the single paths [Fig. 2(a)]. This feature simplifies the design of the ADC: low-pass $\Sigma\Delta$ modulators are inherently less sensitive to component mismatches than bandpass modulators and, therefore, dynamic range and stability become less critical issues.

It is known that a multiple-path circuit achieves, in the sampled-data domain, a z to z^N transformation. Therefore, if $H_p(z)$ is the transfer function of the single path, the total transfer function becomes

$$H_{TOT} = \frac{V_{out}(z)}{V_{in}(z)} = H_p(z^N). \quad (1)$$

If $H_p(z)$ is a low-pass filter, the response folds at multiples of f_{ck}/N , as shown in Fig. 2(b), thus leading, for $N = 4$, to a bandpass transfer function centered around $f_{ck}/4$.

Moreover, the z to z^N mapping given by the multi-path topology multiplies by N the order of the transfer functions. Thus, an N -path architecture achieves the same signal-to-noise-and-distortion ratio (SNDR) of single-path architectures with lower order loop filter with benefits in terms

TABLE I
FOUR-PATH BANDPASS $\Delta\Sigma$ MODULATOR FOR UMTS BASE
TRANSCEIVER STATIONS

Parameter	Value
Clock frequency	280 MHz
Center frequency	70 MHz
Bandwidth	4.375 MHz
Dynamic range	72 dB
Resolution	12 bits
Spurious-free dynamic range	70 dB
Quantizer	9 levels
Order (of each path)	4

of stability and complexity. Finally, an overall clock frequency which is an integer number of times the IF simplifies the demodulation operation.

Table I summarizes the ADC specifications for a converter used in the base transceiver station for third-generation mobile communication standards. The four paths of the ADC operate with a 70-MHz clock while the overall clock frequency is 280 MHz. The required resolution is 12 bit. With a fourth-order modulator it is necessary to use a nine-level quantizer.

The noise transfer function (NTF) is the first concern of the modulator design. Then, it is necessary to verify if the obtained signal transfer function (STF) is flat in the signal band. As already mentioned the N -path architecture achieves the band-stop NTF using in the single paths a low-pass $\Sigma\Delta$ modulator that features a high-pass NTF response. Therefore, the main focus is on the design of a suitable low-pass $\Sigma\Delta$ modulator.

All the architectures used for $\Sigma\Delta$ modulator implementation have the quantizer in a feedback loop which features a sampled-data transfer function $H(z)$. The STF is $H(z)/[1 + H(z)]$ while the NTF is $1/[1 + H(z)]$. The required noise specifications and the quantization noise level (taking into account the resolution of the quantizer, nine levels in the considered case) lead to a suitable filtering mask for the NTF. Using a nine-level quantizer ensures stability of the modulator and limits idle tones, being the power of the quantization noise limited. For this design we obtain a high-pass NTF mask with a stop-band limit at 2.1875 MHz, in-band attenuation of 85 dB, and ripple of 0.5 dB. The transition band spans from 2.1875 to 10 MHz. Filters that fulfill these specifications have been designed using the elliptical, Chebyshev and Butterworth approximation. As expected, the most effective solution is the elliptical filter. The obtained filter order is four with two couples of complex conjugate poles and two couples of complex conjugate zeros located at 0.95 and 2.1875 MHz, respectively. Assuming the use of a sampling frequency equal to 70 MHz, the NTF in the z -domain is

$$\begin{aligned} \text{NTF}(z) &= \frac{1 - 3.9571z + 5.91445z^2 - 3.9571z^3 + z^4}{1 - 0.949z + 1.0511z^2 - 0.2909z^3 + 0.2544z^4} \end{aligned} \quad (2)$$

leading to the response of Fig. 3.

Fig. 3. NTF obtained with the elliptical filter.

The z -domain loop transfer function $H(z)$ that allows us to obtain NTF(z) results from

$$\begin{aligned} \text{NTF}(z) &= \frac{P_e(z)}{Q_e(z)} = \frac{1}{1 + H(z)} \\ H(z) &= \frac{Q_e(z) - P_e(z)}{P_e(z)} \end{aligned} \quad (3)$$

that requires an architecture with a suitable number of degrees of freedom for the circuit implementation. The use of the polynomials of (2) in (3) leads to seven different coefficients. Conventional architectures, like cascade of resonator in feedback, cascade of resonator in feedforward or cascade of integrators do not provide enough design parameters. It is therefore necessary to add extra elements for increasing the number of degrees of freedom. The nonconventional topology of Fig. 4 uses two extra feedforward branches and allows us to synthesize the required transfer function. The scheme includes two resonators whose structure is shown in the insets of Fig. 4. Behavioral simulations on this architecture have shown that a mismatch as large as 5% of the coefficient values (including the additional feedforward paths) does not degrade the modulator performance.

A critical feature of this topology is that it allows only one clock period delay (z^{-1}) between the output of the third integrator and the output of the digital-analog converter (DAC). The sampled-data circuit implementation will distribute the delay associated to the two phases among the cascaded blocks but the simultaneous settling of a pair of cascaded integrators cannot be avoided. In particular, only half clock period is assigned to the cascade of the third and fourth integrator, which therefore have to settle concurrently, while another half clock period is reserved to the comparators and the DAC.

III. SC IMPLEMENTATION

Fig. 5 shows the SC implementation of the resonators of Fig. 4. The matching of capacitors affects the accuracy of coef-

ficients and, in turn, influences the accuracy of poles and zeros. The required minimum standard deviation of the mismatch $\sigma_{C,\min}$ determines the size of the unity capacitance. As already mentioned, simulations performed with the toolbox described in [11] show a limited sensitivity to coefficient values: the required matching accuracy is as large as 5%. The variance of capacitance mismatch is given by $\sigma_C = A_C / (\sqrt{C/C_{\text{ox}}})$. For the used process $A_C = 0.012$ and $C_{\text{ox}} = 0.86 \text{ fF}/\mu\text{m}^2$; therefore, in order to fulfill the coefficient matching requirements, considering also finite gain and bandwidth of the operational amplifiers, the unity capacitance can be 250 fF. This value is low but the large spread of coefficients would lead to a significant capacitance values. In order to limit the capacitance spread the circuit uses for the most critical elements (the feed-back capacitors in the resonators) T-structures with equivalent capacitance C_T given by

$$C_T = \frac{C_{T1}C_{T3}}{C_{T1} + C_{T2} + C_{T3}}. \quad (4)$$

The transfer function of the resonator of Fig. 5 is

$$H_R(z) = \frac{C_2C_1z^{-1/2}}{C_{f2}C_{f1}(1 - z^{-1})^2 + z^{-1}C_2C_T} \quad (5)$$

giving the pair of poles

$$z_{1,2} = e^{\pm j\vartheta}; \quad \vartheta = \arccos\left(1 - \frac{C_2C_T}{2C_{f1}C_{f2}}\right) \quad (6)$$

where ϑ is the angular position on the unit circle of the z -plane. The resonance frequency of the circuit is $\omega_0 f_s \cdot \vartheta$. For this design the value of the angular positions of the poles of the two resonators are $\vartheta_1 = 0.013571 \text{ rad}$ and $\vartheta_2 = 0.03125 \text{ rad}$.

Fig. 6 shows the schematic of the quantizer which is also used for the addition of the four signals coming from the resonator

Fig. 4. Block diagram of one path of the proposed $\Sigma\Delta$ modulator.

outputs. The capacitors charged during phase Φ_2 are connected in parallel and drive the comparator during phase Φ_1 giving

$$V_x = \frac{V_A 2C + V_B 3C + V_C C + V_D C - V_{\text{ref}} C}{8C + C_p}. \quad (7)$$

The eight comparators of the quantizer use proper reference voltages V_{ref} . C_p is the parasitic capacitance at the input of the comparator.

The passive solution adopted for implementing the adder causes an attenuation of the signal by at least a factor 8 (if C_p is negligible). Moreover, the circuit is sensitive to the comparator offset. The offset mismatches cause nonlinearity and possibly nonmonotonicity. Thus, in order to avoid these problems, we use a simple gain stage with autozero in front of the comparators, as shown in Fig. 7.

IV. BUILDING BLOCKS

The ADC has been fabricated using a 0.35- μm double-poly triple-metal 3.3 V CMOS technology. The sampling input

switch is an nMOS device with “bootstrapped” control. The bootstrap circuit maintains almost constant the V_{gs} in the on phase thus leading to a constant on-resistance [12].

The operational amplifier used is based on a fully differential two-stage topology (Fig. 8). The first stage is an n -channel telescopic cascode and the second stage is an n -channel differential pair with active load [13]. Both the first and second stages use a SC common-mode feedback. Miller capacitors with zero nulling resistor ensure stability. Behavioral simulations of the modulator have shown that at least a dc gain of 80 dB, a gain-bandwidth product of 325 MHz and a slew-rate of 400 V/ μs are required to guarantee the $\Sigma\Delta$ modulator performance (considering also path mismatches). Table II summarizes the simulated performance of the amplifier. The achieved values are larger than the specifications, in order to ensure proper operation also in the process and temperature corners.

Fig. 9 shows the overall clock distribution scheme. A 280-MHz sinusoidal clock is converted to a square waveform and its frequency divided by 4, thus obtaining the 70-MHz square wave main clock. The main clock is also shifted three

Fig. 5. SC implementation of the resonator.

Fig. 6. SC implementation of the quantizer.

times by 90° , producing the clocks for the four modulator paths. They are the basis for the controls required by each modulator path. Simulations of the entire phase generator, including also the bonding wires and package models, gives a clock jitter that goes from 335 fs to 940 fs depending on the considered

Fig. 7. Comparator structure: Gain stage followed by a latch.

phase. The lowest jitter phases are the ones used to drive the bootstrapped input switches.

The comparator (Figs. 6 and 7) consists of a gain stage followed by a latch [14]. During phase Φ_2 the latch is reset while the gain stage is closed in unit gain configuration. The p-channel current source enhance the gain without requiring long p-channel diode-connected elements. Also during phase Φ_2 the eight capacitors of the passive adder sample the offset of the gain stage. During Φ_1 the capacitors sample the outputs A, B, C, D of the resonators (Fig. 4) as well as the reference voltages with the proper weight (Fig. 6) and gain stage settles to the required value. Finally, on the falling edge of Φ_1 the latch is released. This solution strongly reduces the offset as well as the kickback from the latch.

Fig. 8. Schematic diagram of the fully differential operational amplifier.

TABLE II
SIMULATED PERFORMANCE OF OPERATIONAL AMPLIFIER

Parameter	Value
DC gain	95 dB
Gain-bandwidth product	550 MHz
Phase margin	53°
Slew-rate	796 V/ μ s
Settling time (14 bit)	4.1 ns
Current consumption	8.9 mA
Input referred noise (0 - 70 MHz)	13 μ V

The DAC is an SC structure at the input of the integrators. It generates a midread input–output characteristic with levels symmetrical with respect to the analog zero. Fig. 10 shows the structure of the DAC. During Φ_2 , all the capacitors are discharged, while during Φ_1 depending on the modulator outputs some capacitors are connected to the positive or negative reference voltage. The capacitors which are not used are connected to ground. The mismatch among capacitors is corrected a using the noise-shaped dynamic element matching (DEM) technique [15]. By using this technique, it is possible to apply the noise-shaping principle to the errors introduced by

Fig. 9. General block diagram of the clock distribution scheme for the modulator.

capacitor mismatches. In our design, this known technique has been adapted to work with bipolar signals [16], [17].

Noise in the reference voltages, if larger than 5 mV, compromises the overall performance. The problem is mitigated by using low-noise reference buffers.

V. EXPERIMENTAL RESULTS

The micrograph of the fabricated chip is shown in Fig. 11. The area of the complete modulator and an extra single-path for testing purposes is $4 \times 6.8 \text{ mm}^2$. The actual four-path $\Sigma\Delta$ modulator area is 20 mm^2 . The chip uses a CQFP package.

Fig. 10. SC implementation of the DAC.

Fig. 11. Microphotograph of the four-path $\Sigma\Delta$ modulator chip prototype.

The chip has been extensively characterized. The PCB used for the testing employs very low-noise high-linearity input buffer to obtain with a feedback scheme the single-ended to double-ended conversion. The use of an RF transformer is also possible but the solution was not used because it causes higher noise and masks the circuit performance. Moreover, the board includes low-noise circuitry for driving the internal reference buffers.

The output spectrum of the four-paths path $\Sigma\Delta$ modulator is shown in Fig. 12. The clock frequency is 280 MHz and the input signal amplitude is 56 mV peak-to-peak differential at 69.95

TABLE III
PERFORMANCE SUMMARY OF FOUR-PATH $\Sigma\Delta$ MODULATOR

Parameter	Value
Maximum sampling frequency	280 MHz
Maximum signal bandwidth	4.375 MHz
DR @ 4.375 MHz	72 dB (corresponding to 12 bits)
Maximum differential input signal	1 V _{pp}
SFDR	> 69 dB
IMD3 @ 650 mV _{pp}	< -60 dBc
Power supply voltage	3.3 V
Power consumption	480 mW

TABLE IV
COMPARISON OF SOME PREVIOUSLY PUBLISHED WIDE-BAND ADCs

Paper	Bandwidth [MHz]	IF [MHz]	DR [dB]	SFDR [dB]	f_{ck} [MHz]	P [mW]	Technology
This Work	4.375	70	72	69	280	480	0.35 μ m
ISSCC 2002 [18]	3.84	60	48	60	80	38	0.35 μ m
CICC 2003 [19]	2.5	0	82	100	40	495	0.35 μ m
ESSCIRC 2003 [20]	15	0	62	-	300	70	0.13 μ m
ISSCC 2004 [21]	2.5	40	72	86	120	150	0.18 μ m
VLSI 2004 [5]	1.25	60	52	-	240	37	0.35 μ m
ISSCC 2005 [22]	20	0	75	87	80	240	0.18 μ m

MHz. No spurious tones due to offset and gain mismatches among the different paths can be observed, thank to the careful design and to the relatively large noise floor allowed. The SNDR as a function of the input signal amplitude for the modulator is shown in Fig. 13. The DR achieves 72 dB peak (12 bits). The peak SNDR is achieved with an input signal of -13 dB. For larger amplitudes the performance is degraded by quantization noise tones due to the conditional stability of the modulator. The maximum signal amplitude was sacrificed in the design phase to guarantee fast overload recovery (fast resume of normal operation when the input signal becomes momentarily larger than -13 dB). This is an important feature in telecommunication systems and is worth a reduction of the maximum input signal.

Fig. 14 shows the output spectrum for an intermodulation test. The input signal consists of two tones at 69.9/70.1 MHz and amplitude 650-mV peak-to-peak differential each. The intermodulation products are 60 dB below the signal tones, corresponding to an HD₃ of -69 dBc and hence leading to an equivalent SFDR of 69 dB.

The total power consumption is 480 mW with 3.3-V supply voltage, as expected. The features of the modulator are summarized in Table III.

Table IV compares the performance of this design with other published wide-band ADCs. Observe that the previous art either features limited bandwidth or smaller DR when compared to this design. Only power/band ineffective (for the considered application) low-pass solutions achieve performance comparable with the obtained results or better. For example, the solution reported in [20] requires a slew-rate in the operational amplifiers

Fig. 12. Output spectrum of the four-path $\Sigma\Delta$ modulator obtained with clock frequency of 280 MHz, input signal amplitude of 56 mV peak-to-peak differential, and input signal frequency of 69.95 MHz.

Fig. 13. SNDR as a function of the input signal amplitude of the four-paths $\Sigma\Delta$ modulator obtained with clock frequency of 280 MHz and input signal frequency of 69.95 MHz.

15/70 lower than the proposed modulator and uses a technology scaled by about 0.4. Therefore, a rough comparison with homogeneous features would require the multiplication of the power consumption of [20] by a factor of about 12.

VI. CONCLUSION

Bandpass modulators are a promising solution for the conversion of high-speed telecommunication signals. In this paper, we presented a four-path time-interleaved bandpass $\Sigma\Delta$ modulator architecture, which allows us to relax the speed require-

ments of each path, thus making less critical the design of the building blocks, namely operational amplifiers, switches, ADC and DAC. The proposed architecture is demonstrated through the implementation of a four-path bandpass $\Sigma\Delta$ modulator operating with an overall clock frequency of 280 MHz (each path works at 70 MHz). The achieved performance (DR of 72 dB, 12 bits of resolution, with a signal bandwidth of 4.375 MHz centered around an intermediate frequency of 70 MHz, SFDR of 69 dB), satisfies the challenging requirements of 3G multicarrier wireless receivers.

Fig. 14. Output spectrum of the four-paths $\Sigma\Delta$ modulator obtained in the intermodulation test with clock frequency of 280-MHz, input signal amplitude of 650-mV peak-to-peak differential and frequency the two tones of 69.9/70.1 MHz.

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Dr. Malcovati was Guest Editor for the *Journal of Analog Integrated Circuits and Signal Processing* for the special issue on IEEE ICECS 1999. He served as Special Session Chairman for the IEEE ICECS 2001 Conference, as Secretary of the Technical Program Committee for the ESSCIRC 2002 Conference, and as Technical Program Chairman of the IEEE PRIME 2006 Conference. He was and still is member of the Scientific Committees for several International Conferences, including ESSCIRC, DATE, and PRIME. He is associate editor of the *Journal of Circuits, Systems, and Computers*.

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Dr. Maloberti was the recipient of the XII Pedriali Prize in 1992 for his technical and scientific contributions to national industrial production, co-recipient of the 1996 Institute of Electrical Engineers (U.K.) Fleming Premium for the paper "CMOS Triode Transistor Transconductor for high-frequency continuous time filters." He was Vice-President, Region 8, of the IEEE Circuit and Systems Society (CAS-S) from 1995 to 1997 and an Associate Editor of IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS—II: ANALOG DIGITAL SIGNAL PROCESSING, in 1998. He received the 1999 IEEE CAS-S Meritorious Service Award, the 2000 CAS-S Golden Jubilee Medal, and the IEEE Millenium Medal. He was the President of the IEEE Sensor Council (2002–2003) and a member of the BoG of the IEEE CAS-S (2003–2005). He is currently an Associate Editor of the IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS—II: EXPRESS BRIEFS, and member of the Editorial Board of *Analog Integrated Circuits and Signal Processing*.