

An FPGA Based Passive K-Delta-1-Sigma Modulator

Angsuman Roy, Matthew Meza, Joey Yurgelon, and R. Jacob Baker
 Department of Electrical and Computer Engineering, University of Nevada, Las Vegas
 angsumanroy@gmail.com

Abstract—An FPGA based 2nd-order passive KD1S sigma-delta modulator was designed, simulated and tested. The design is implemented on an Altera Cyclone IV EP4CE115 FPGA. All active components such as digital logic, clock circuitry, and registers are located internally on the FPGA chip with only passive RC lumped analog components located off chip. The circuit uses eight logic elements and two PLL blocks on the FPGA to create an eight path KD1S sigma-delta modulator. The design performance was quantified at effective sampling rates of 80 MHz and 450 MHz. The implementation achieved a peak SNR of 58 dB and an ENOB of 9.3 bits at a 450 MHz effective sampling rate. The key benefit of this approach is the absence of active analog components, very low power, and high-speed sampling.

I. INTRODUCTION

Sigma-delta ($\Sigma\text{-}\Delta$) analog to digital converters (ADC) are popular for high resolution, relatively low frequency data conversion. The primary reasons for their popularity are the heavy reliance on digital components, low cost, and ease of implementation in nano-CMOS processes. As a result, sigma-delta ADCs have dominated the fields of industrial sensing, audio, and narrow bandwidth radio frequency applications. Despite their ubiquity, two main drawbacks have limited the use of sigma-delta ADCs in wide bandwidth applications. These drawbacks are the requirement for a high oversampling ratio and the high gain-bandwidth product (GBW) required from analog integrators [1]. The high oversampling ratio (OSR) is needed to get high resolution from the ADC. This high OSR requires a correspondingly high GBW from the analog integrators for proper feedback loop operation. Furthermore, there is much difficulty in designing an appropriate active analog integrator in a nano-CMOS process. Maximum clock rates are also a limiting factor. In order to increase the adoption of $\Sigma\text{-}\Delta$ ADCs in wide-bandwidth applications, it is imperative that the constraints of analog integrators and fast clock rates are eliminated. In this work, we propose a KD1S $\Sigma\text{-}\Delta$ modulator topology to relax clock rate requirements as well as a passive integrator to eliminate the difficulty in designing an active integrator. We also propose implementing this $\Sigma\text{-}\Delta$ ADC on a field-programmable-gate-array (FPGA) chip which lends itself to be readily adapted to a wide variety of applications.

II. PASSIVE KD1S SIGMA-DELTA TOPOLOGY

Conventional $\Sigma\text{-}\Delta$ ADCs usually have one quantizing element such as a comparator or D flip-flop (D-FF). This

quantizing element is clocked at a significantly higher frequency than the bandwidth of the desired input signal with the quotient of the two being defined as the OSR. Generally, the OSR is a value ranging from 64-1024. For a wide input signal bandwidth in the MHz range, this would require a clock frequency of hundreds of megahertz to a few gigahertz. As clock frequencies increase, the design of clock generators becomes more difficult with the result being lower stability and higher jitter. One solution for maintaining a high OSR with lower clock frequencies is the KD1S $\Sigma\text{-}\Delta$ ADC. This topology has been both theoretically and experimentally verified [4].

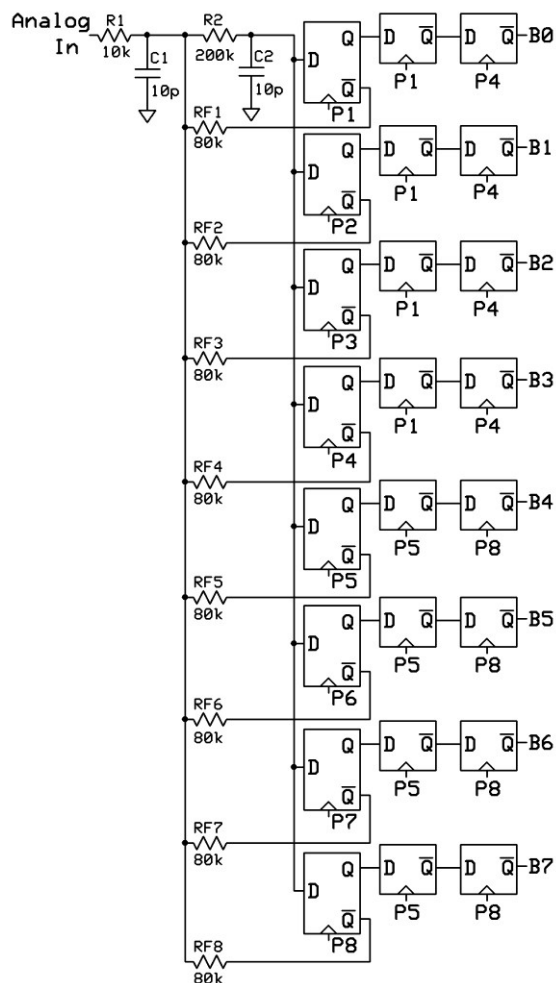


Figure 1 – Passive KD1S Σ - Δ Modulator

The KD1S Σ - Δ ADC gets its name from the fact that there are K -delta feedback paths while there is only one integrator, or 1 -sigma. For K number of feedback paths there needs to be K quantizers. This can be clearly seen in Fig. 1, a conceptual schematic of the proposed KD1S modulator. On the surface, this topology may appear similar to a parallel or time-interleaved Σ - Δ modulator, such as in [2] and [3] but there are key differences and the fundamental operation is different. There are eight clocks, one for each D-FF. The clocks are phase shifted such that the edges are equally spaced and sequential as illustrated in Fig. 2. Although these clocks are at the same frequency, the effective sampling rate $f_{s,new}$, is equal to the clock frequency f_s , multiplied by the number of K paths, which is eight in this case, as given in equations (1) and (2). The reason for this is due to the fact that each D-FF is edge triggered; the sampling rate is only dependent on the number of rising edges in a time interval and not the actual clock frequency. A theoretical noise transfer function for a KD1S modulator is illustrated in Fig. 3. The noise shaping occurs up to the nyquist frequency $Kf_s/2$. Therefore, the noise shaping of a KD1S modulator is identical to a conventional Σ - Δ modulator clocked at a rate K times faster. The proof of this can be found in [5].

$$(1) \quad \frac{T_s}{K_{path}} = \frac{T_s}{8}$$

$$(2) \quad f_{s,new} = K_{path} * f_s$$

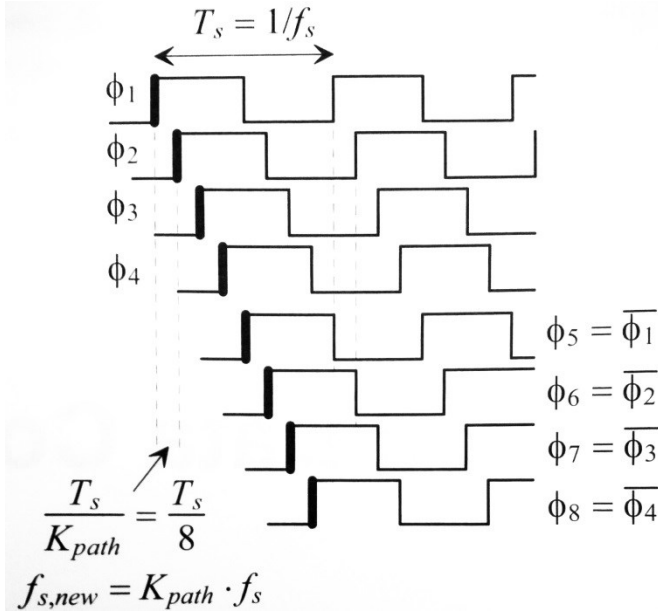


Figure 2 – Phase Shifted Clocks [1]

The proposed KD1S Σ - Δ modulator does not use a conventional active integrator. Instead it uses a passive

network of resistors and capacitors to perform its integration function. There are a few passive topologies to choose from, but it appears that a modified second order topology found in [6] has the best performance. The filtering in the proposed design in Fig. 1 is done by R_1 , C_1 , R_2 and C_2 . The noise shaping is 2nd order and the transfer functions can be adapted from [5] to find that the bandwidth of this network is about 400 kHz. The feedback resistors in Fig.1, R_{F1} - R_{F8} are eight times the value of R_1 . This is needed to ensure that the gain of the modulator is unity. The parallel combination of the feedback resistors is equal to the value of R_1 .

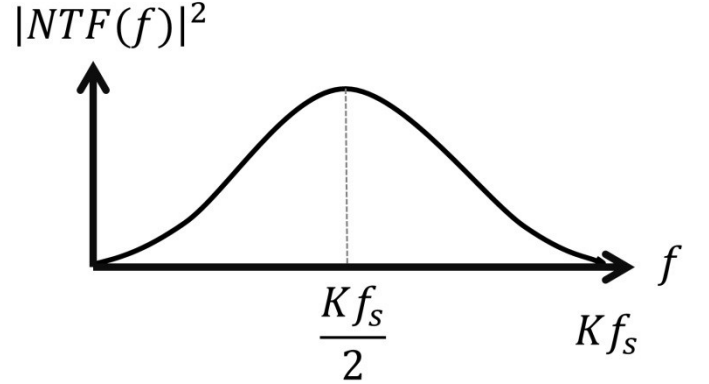


Figure 3 – NTF of a KD1S Modulator

III. FPGA IMPLEMENTATION

The proposed KD1S Σ - Δ modulator is well suited for implementation on an FPGA. An almost fully digital ADC can be created with just an FPGA chip and the required passive off-chip RC components. No active analog circuitry is required so the implementation of the ADC with the FPGA is very simple. This is advantageous in systems which are mainly digital but require an ADC. An external ADC chip can be avoided and existing FPGA hardware can be leveraged to implement a medium resolution, wide bandwidth ADC which could be suitable for many applications.

In this work, we implemented the proposed Σ - Δ modulator on an Altera Cyclone IV EP4CE115 FPGA as found on the Terasic DE2-115 evaluation board. This FPGA has 114,480 logic elements, four phase-locked loops (PLL), and 528 user I/O pins [7]. The PLLs on the FPGA make it easy to implement the eight phase shifted clocks required by the proposed modulator. Each PLL has five outputs which can have their phase individually controlled. A total of two PLLs are required to get eight phase shifted clocks, each shifted sequentially by 22.5 degrees. These PLL outputs are labeled P1-P8 in Fig. 1. The PLL outputs are not available for probing off-chip, but a simulation shows proper operation in Fig. 4. The rest of the design consisting of the D-FFs was done with the schematic capture tool in Quartus. The first column of D-FFs in Fig. 1 are used as the sampling elements. The next two columns of D-FFs form a register that is used to relock the data for further digital signal processing (DSP). In this work, DSP filtering was not implemented on the modulator output within the FPGA and only the performance of the Σ - Δ

modulator was examined. The outputs from the registers, B0-B7 are connected to general purpose input/output (GPIO) pins which allow the FPGA to drive off chip loads. The entire modulator uses two PLLs and eight logic elements which is only 0.007% of the total number of logic elements.

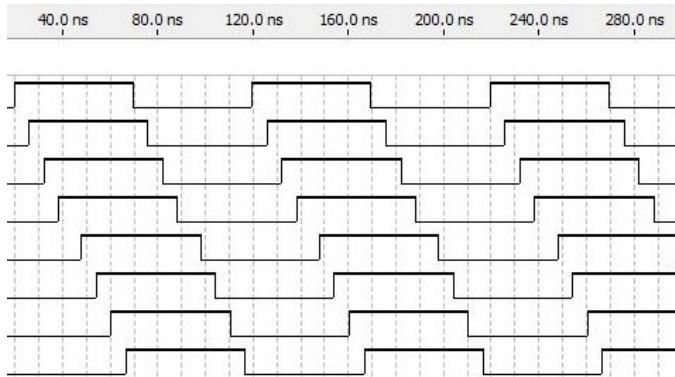


Figure 4 – PLL Simulation in Quartus

IV. TEST RESULTS

The proposed FPGA based KD1S $\Sigma\text{-}\Delta$ modulator's performance was extensively verified using an Advantest R3267 spectrum analyzer. The bandwidth of this spectrum analyzer is 100 Hz to 8 GHz. Two test configurations were used to make measurements. The first configuration, shown in Fig. 5A was used for wide band measurements to show noise shaping up to the sampling frequency of Kf_s . This configuration is a 10X attenuator to provide lighter loading on the FPGA's GPIO pins. The second configuration shown in Fig. 5B was used to make high resolution measurements to quantify signal-to-noise ratio (SNR) as a function of input frequency. This configuration uses a high impedance input buffer with a gain of two. This buffer is used to drive a 7th order low-pass filter that removes the quantization noise [8]. The eight 47 ohm resistors are the series termination resistors inherent to the DE2-115 board. In all cases proper 50 ohm coaxial cables were used and all parts of the system were properly terminated.

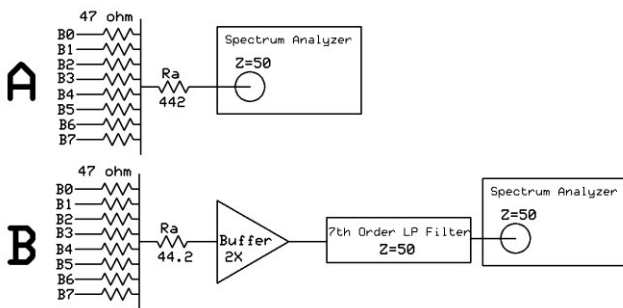


Figure 5 – Test Configurations

Initially, the proposed modulator was clocked at a frequency of 10 MHz to verify proper operation. A clock frequency of 10 MHz with eight K paths results in an effective sampling frequency of 80 MHz. Using the test configuration in Fig. 5A, the noise shaping was verified. One would expect

the spectrum analyzer to display an image similar to Fig. 3. However, since the digital outputs were summed together into one single analog output the noise transfer function (NTF) displays the comb filtering characteristic which is given by equation (3). The end result of this is that there are nulls in the NTF at every integer multiple of the clock frequency. This is verified in the screen display of the spectrum analyzer in Fig. 6. An input of 100 kHz was applied to the modulator and can be seen as the spike on the far left. The noise displayed around the input signal is excessive and does not accurately reflect the performance of the modulator. This is due to the need to set a large resolution bandwidth (RBW) on the spectrum analyzer due to the displayed 80 MHz span.

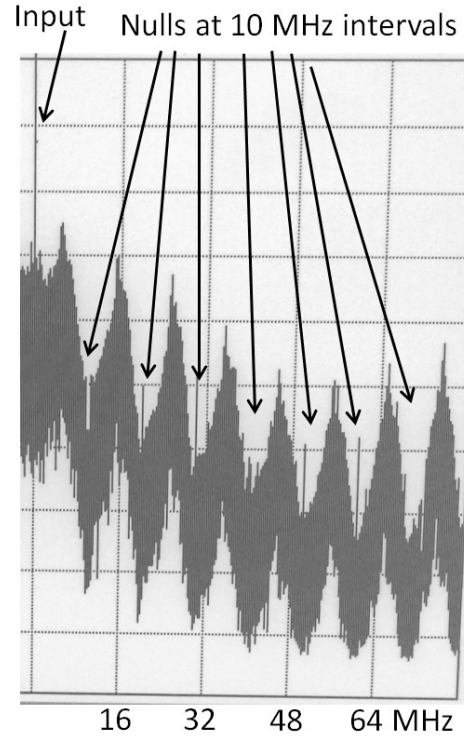


Figure 6 – Comb Filtered Modulator Noise

$$(3) H(z) = \frac{1-z^{-Kpath}}{1-z^{-1}} = \frac{1-z^{-8}}{1-z^{-1}}$$

In order to properly quantify the SNR of the proposed modulator, the test set-up in Fig. 5B was used. The input signal applied to the modulator is from an HP654A test oscillator with a measured total harmonic distortion (THD) of less than 0.1%. The input signals amplitude is 1.5 V peak with a 1.65 V offset. Since the KD1S modulator is configured to interface with 3.3 V low-voltage CMOS (LVCMOS) logic levels, this input amplitude is close to a full scale input. This allows meaningful SNR measurements to 60 dB. The buffer used has a bandwidth of 200 MHz. The 7th order LP filter was chosen such that the bandwidth was restricted to around six times the input signal frequency. For example, a 600 kHz 7th order LP filter was used for a 100 kHz input frequency. The SNR was computed using the power function on the spectrum

analyzer. This measurement was repeated for a wide range of frequencies.

The results of SNR testing are shown in Fig. 7. The proposed modulator was set up for an 80 MHz effective sampling rate and the input signal was varied from 10 kHz to 200 kHz. For each input frequency, the SNR was measured with the 7th order LP filter in place. The highest SNR achieved was 55 dB for an input frequency of 10 kHz corresponding to an effective number of bits (ENOB) of 8.8 bits. As the input frequency is increased, the SNR falls to a fairly constant range between 45 dB and 43 dB corresponding to an ENOB of between 7.3 and 7 bits. After verifying that the proposed modulator does indeed work, the test was repeated for a higher effective sampling rate.

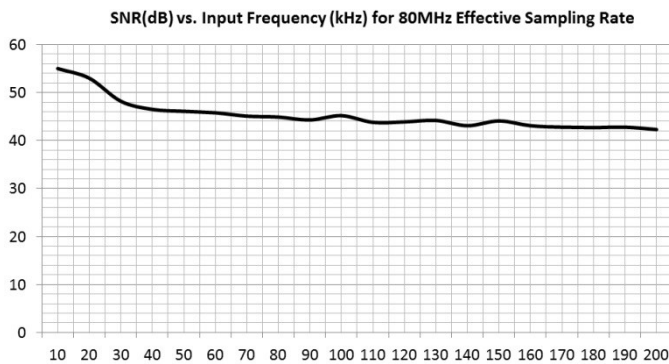


Figure 7 – Measured SNR with 80 MHz Sampling Rate

An upper limit on the effective sampling rate was chosen at 450 MHz. This is achieved with a 50 MHz clock frequency. This limit was chosen due to the fact through-hole mounted passive components were used. The use of surface mount components can push this limit higher. The range of input frequencies was increased to 1 MHz. The results are shown in Fig. 8. At an input frequency of 10 kHz the SNR increases to 58 dB, a 3 dB improvement over the previous sampling rate corresponding to an ENOB of 9.3 bits. Within the same frequency range up to 200 kHz, the SNR stays above 50 dB which is a significant improvement over the 80 MHz sampling rate. As the input frequency rises to 1 MHz, the SNR drops steadily to a minimum of 42dB.

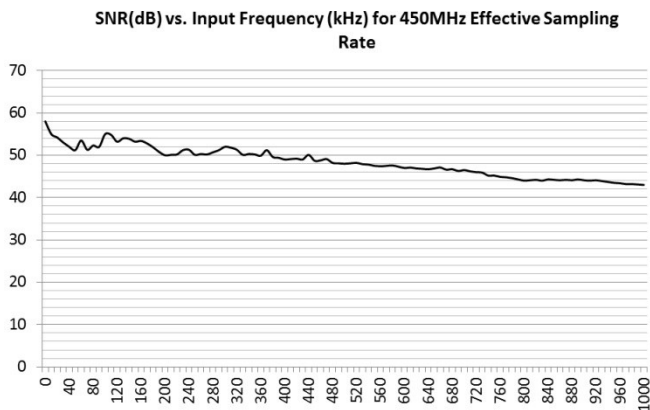


Figure 8 – Measured SNR with 450 MHz Sampling Rate

V. CONCLUSION

Through the use of passive RC components and KDS principles, a medium resolution, wide-bandwidth ADC was implemented using a small number of logic elements on an FPGA. The logic element budget for the proposed Σ - Δ modulator is very low at only 0.007% of the total number of available logic elements. With modern FPGAs, and small SMD passive components, effective sampling rates into the GHz range should be possible. Furthermore, the sophisticated PLLs available on FPGAs could allow for a wide range of sampling frequencies and even dynamic adjustment of sampling rates. For many applications, an ADC built around the proposed Σ - Δ modulator could prove to be the lowest cost ADC option for FPGA based systems.

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