On Incremental Sigma-Delta Modulation with Optimal Filtering

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Abstract

The paper presents a quantization-theoretic framework for studying incremental $\Sigma\Delta$ data conversion systems. The framework makes it possible to efficiently compute the quantization intervals and hence the transfer function of the quantizer, and to determine the mean square error (MSE) and maximum error for the optimal and conventional linear filters for first and second order incremental $\Sigma\Delta$ modulators. The results show that the optimal filter can significantly outperform conventional linear filters in terms of both MSE and maximum error. The performance of conventional $\Sigma\Delta$ data converters is then compared to that of incremental $\Sigma\Delta$ with optimal filtering for bandlimited signals. It is shown that incremental $\Sigma\Delta$ can outperform the conventional approach in terms of signal to noise and distortion ratio (SNDR). The framework is also used to provide a simpler and more intuitive derivation of the Zoomer algorithm.

Index Terms

Sigma-Delta ($\Sigma\Delta$), incremental A/D converter, optimal filter, $\Sigma\Delta$ transfer function, time-domain analysis.

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I. Introduction

$\Sigma\Delta$ data converters, also known as scalar predictive quantizers or oversampling Analog-to-Digital converters, are widely used in audio [1], [2] and communication systems [3], [4], [5]. These systems can achieve very large dynamic range without the need for precise matching of circuit components. This is especially attractive for implementation in scaled technologies where transistors are fast but not very accurate. They are also among the most power efficient ADC architectures [4].

Exact analysis of $\Sigma\Delta$ data converters is difficult due to their highly nonlinear structure. As a result, optimal decoding (filtering and decimation) algorithms are generally not known, although there has been much work on decoding schemes [6], [7], [8]. Conventionally, linear models with quantization noise modeled as additive white noise have been used [9]. This noise model, however, was shown to be quite inaccurate for coarse quantization. In [11], Candy derived a more accurate approximation of the noise power spectral density showing that it is not white. Subsequently, Gray [6] derived the exact spectrum and determined the optimal MSE linear filter for constant inputs. Gray's framework was later used in multiple studies (e.g., [12], [13]). This type of exact analysis, however, is too complex for most practical $\Sigma\Delta$ systems and the conventional linear method is still used and then verified and tweaked using time-domain simulations [9], [10].

Recently, $\Sigma\Delta$ data converters have become popular in sensing applications such as imaging [14], [15], [16], [17] and accurate temperature sensing [18]. In such applications, the input signal is very slowly varying with time, and as a result can be viewed as constant over the conversion period. Of particular interest is the special case of *incremental* $\Sigma\Delta$ [19], where the integrator is reset prior to each input signal conversion. In this case, the optimal filter with respect to the mean-squared error criterion denoted by Zoomer was found by Hein et al. [20]. The approach used in [20], however, does not naturally lead to full characterization of the quantizer transfer function or the determination of the MSE or maximum error, especially for second and higher order $\Sigma\Delta$ modulators. Knowledge of the exact transfer function of the $\Sigma\Delta$ converter can enable the system designer to achieve the system requirements with a lower oversampling ratio. Moreover, as noted by Markus et al. [19], data converters with high absolute accuracy are often required in instrumentation and measurement. In [19], bounds on the maximum error of incremental $\Sigma\Delta$ using some conventional linear filters was derived, but there is still a need to derive the maximum error of the optimal filter.

In this paper, we introduce a time-domain, quantization-theoretic framework for studying incremental $\Sigma\Delta$ quantization systems. The framework allows us to determine the quantization intervals and hence the transfer function of the quantizer, and to precisely determine the MSE and maximum error for the optimal filter. For constant inputs, we demonstrate significant improvements in both MSE and maximum error over conventional linear filters [19]. These results imply that an incremental $\Sigma\Delta$ with optimal filtering can achieve the same performance as that of conventional $\Sigma\Delta$ systems at lower oversampling ratio and hence lower power consumption. We show that incremental $\Sigma\Delta$ can outperform the conventional approach in terms of SNDR in addition to not having the idle tones artifacts of conventional $\Sigma\Delta$. Using our framework we are also able to compare the performance of conventional $\Sigma\Delta$ data converter to that of incremental $\Sigma\Delta$ with optimal filtering for bandlimited input signals. We also use our framework to provide a simpler and more intuitive proof of the optimality of the Zoomer algorithm [20].

In the following section, we introduce our framework and use it to study first-order $\Sigma\Delta$ quantization systems. In Section III, we extend our results to incremental second-order $\Sigma\Delta$ quantization systems. In each case, we compare the performance of the optimal filter to that of linear filters. In Section IV, we discuss potential applications of our results to sensor systems and compare the performance of the incremental $\Sigma\Delta$ modulator using optimal filtering to conventional $\Sigma\Delta$ systems using linear filtering for sinusoidal inputs. We also show that incremental $\Sigma\Delta$ does not suffer from the problem of idle tones.

II. Incremental first-order $\Sigma\Delta$

A block diagram of a $\Sigma\Delta$ quantization system with constant input is depicted in Figure 1. The figure also provides the correspondence between the terminology used in the circuit design literature and the quantization-theoretic literature [22]. The $\Sigma\Delta$ modulator itself corresponds to the *encoder* in the quantization-theoretic setting. It outputs a binary sequence of length m that corresponds to the quantization interval that the input x belongs to. The set of such binary sequences is the *index set* in the quantization-theoretic setting. The filter, which corresponds to the *decoder*, provides an estimate $\hat{x}$ of the input signal x. Given an index, the optimal decoder outputs the *centroid*, i.e., the midpoint, of the *quantization interval* corresponding to the index. Figure 2 is an example that illustrates the terminology used to characterize a quantization system. Note that this correspondence applies to all data converters, and while a $\Sigma\Delta$ data converter system performs oversampling internally, the entire system can be viewed as a Nyquist rate data converter. Also note that in practice a data converter system produces an estimate that has finite precision and offset and gain mismatch with the true estimate $\hat{x}$. Finally, note that quantization framework in Figure 1 refers to the $\Sigma\Delta$ data conversion system and not to the coarse quantizer used in the modulator itself.

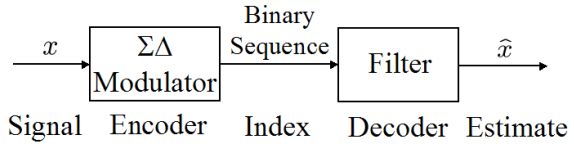

Fig. 1. Block diagram of a $\Sigma\Delta$ quantization system with constant input. The terms under the figure provide the corresponding object names in quantization theory.

In this section we study the discrete time first-order $\Sigma\Delta$ modulator (see Figure 3). For simplicity we focus on the single-bit case. However, our results can be easily extended to multi-bit $\Sigma\Delta$ modulators. Without loss of generality, we assume that the comparator threshold value is 1 and the input signal $x \in [0, 1]$. The input to the modulator is integrated, thus creating a ramp with a slope proportional to the constant input. At time $n = 1, 2, \ldots, m$, the ramp sample $u(n)$ is compared to the threshold value of 1 and the output $b(n) = 1$ if $u(n) > 1$, otherwise $b(n) = 0$. If $b(n) = 1$, a one is subtracted from the ramp value. Thus in effect the modulator is predicting the ramp value by counting the number of ones and subtracting off its prediction via the feedback loop. This operation is illustrated in Figure 4,

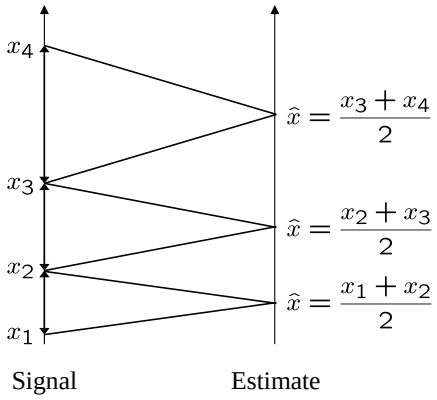

Fig. 2. Illustration of the quantization-theoretic terminology. $x_1 \ldots x_4$ are the transition points, $[x_1, x_2) \ldots [x_3, x_4)$ are the quantization intervals that correspond to different indices generated by the encoder and should be optimally decoded to $(x_1 + x_2)/2 \ldots (x_3 + x_4)/2$.

where the output of the integrator $u(n)$ is plotted versus time for a fixed input value.

The output of the integrator is thus given by

$$u(n) = x - b(n-1) + u(n-1) = nx - \sum_{0}^{n-1} b(i). \quad (1)$$

We shall assume that the integrator is reset to zero at the beginning of conversion, i.e. $u(0) = 0$, which is the case in incremental $\Sigma\Delta$.

In conventional $\Sigma\Delta$ data converters, the binary sequence generated by the modulator is decoded using a linear filter, such as a counter or a triangular filter, to produce an estimate of the input signal [19]. Such linear filters, however, are not optimal and result in significantly higher distortion compared to the optimal filter. The optimal filter in the form of "Zoomer" implementation was derived in [20]. In [21] McIlrath developed a nonlinear iterative filter that achieves significantly better performance compared to linear filters.

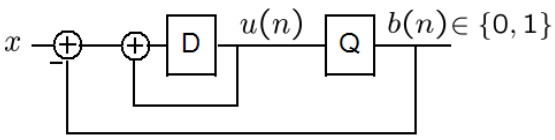

Fig. 3. Block diagram for the first-order $\Sigma\Delta$ modulator. D refers to the delay element and Q refers to the one-bit quantizer.

We are now ready to introduce our framework for studying incremental $\Sigma\Delta$ data converter. Note that to completely characterize a quantization system, one needs

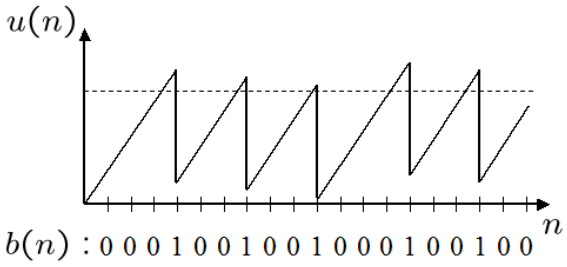

Fig. 4. First-order sample integrator output and the corresponding sequence. The discrete-time waveform is plotted as a continuous waveform for simplicity of illustration in all waveforms throughout the paper.

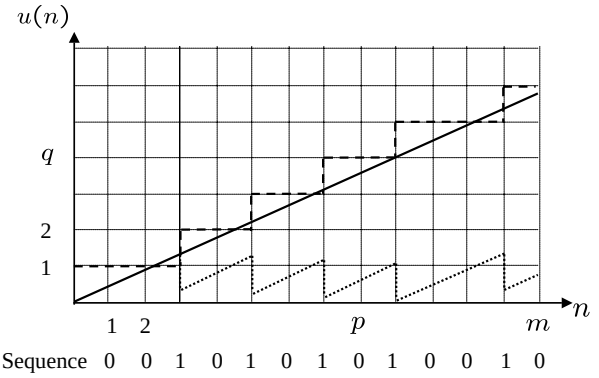

Fig. 5. The equivalent ramp and it's corresponding predictor. The dotted line is the integrator output, the solid line is the equivalent ramp, and the dashed line is the predictor.

to determine the quantization intervals induced by its encoder. Equivalently, one needs to determine the *transition points* between consecutive quantization intervals.

The $\Sigma\Delta$ modulator can be viewed to be performing a sequence of *effective comparisons*, $u(n) \gtrless 1$ which are equivalent to $x \gtrless (1 + \sum_0^{n-1} b(i))/n$. Clearly any encoder that produces bits corresponding to the same effective comparisons is equivalent to the $\Sigma\Delta$ modulator. Each effective comparison defines and upper denoted by UB or a lower bound denoted by LB, on the input. To find the transition points of the first-order $\Sigma\Delta$ modulator, we define the following equivalent encoder. Referring to (1), it is clear that the sequence of comparisons $u(n) \gtrless 1$, $1 \leq n \leq m$, is the same as the sequence of comparisons of the *predictor* $1 + \sum_0^{n-1} b(i)$ to the *equivalent ramp* nx, which is the line segment from the origin of slope x. Therefore, if the modulator is replaced by an encoder that performs the comparisons to the equivalent ramp, its output sequence would be identical to that of the modulator for all input values. Note that this equivalence holds for all inputs $x \in [0, 1]$. Figure 5 depicts this equivalence graphically.

To find the transition points using the above equivalence, consider the square lattice shown in Figure 6

$$L_m = \left\{(p, q) : p, q \in \mathbb{Z}^+, 1 \leq q \leq p \leq m\right\}.$$

We now show that the set of transition points is simply the set of slopes of the equivalent ramps that pass through points of the square lattice, which we denote as

$$S_m = \left\{\frac{q}{p} : (p, q) \in L_m\right\}.$$

The elements of S_m form the Farey sequence [23] of order m.

To show that S_m is the set of transition points, we first show that any element of S_m is a transition point. Consider the equivalent ramp of slope q/p. It is not difficult to see that any equivalent ramp of higher slope, i.e., any signal value $x > q/p$, generates a different sequence from q/p itself. Therefore, any $q/p \in S_m$ is a transition point.

To show that any transition point y belongs to S_m, we need to show that $y = q/p$ for some integers $1 \leq q \leq p \leq m$. Since y is a transition point, the sequence corresponding to y and $y + \delta$, for any $\delta > 0$, must differ in at least one sample time. Let such sample time be $1 \leq p \leq m$, then there must be an integer $q \leq p$ such that the predictor value is q at time p, which implies that $y = q/p \in S_m$.

The transition points can be computed simply by computing the elements of S_m or by parsing the Farey or Stern-Brocot tree [23]. The worst case running time is $\mathcal{O}(m^2)$.

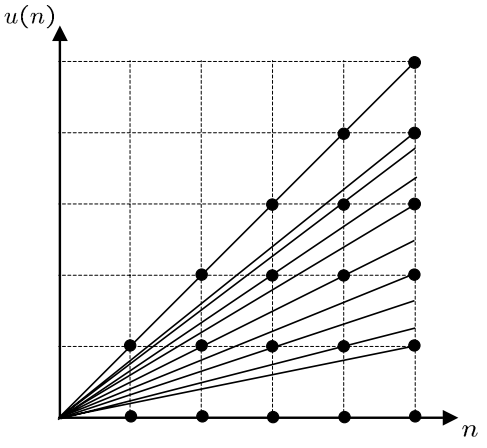

Fig. 6. Lattice for first-order $\Sigma\Delta$.

The optimal filter produces the centroid of the quantization region corresponding to the sequence generated by the modulator. For example for the sequence 00101, which corresponds to $x \in [2/5, 1/2]$, the optimal filter produces the estimate $\hat{x} = 9/20$.

Algorithm 1 Optimal Filtering Algorithm(Zoomer Implementation)

```
begin
SEQ ← Modulator generated sequence
UB ← 1, LB ← 0
Predictor ← 1
for p = 1 : m do
  if SEQ(p) = 1 then
    LB ← max{LB, Predictor/p}
    Predictor ← Predictor + 1
  else
    UB ← min{UB, Predictor/p}
  end if
end for
return (UB + LB)/2
end
```

The equivalent encoder framework provides a simple and intuitive way to prove the optimality of the $\mathcal{O}(m)$ Zoomer algorithm [20] (algorithm 1). To prove the optimality of this implementation, note that a filter is optimal if given any index sequence produced by the modulator, it outputs the centroid of the quantization interval to which the input signal belongs. So, for any input $x \in [0, 1]$, we need to show that the resulting UB and LB are the transition points for the quantization interval of x. We show this by induction. Clearly, this is the case for $m = 1$. Assume that this is also the case up to $m - 1$, i.e., for $m - 1$ samples, UB and LB are the transition points of the quantization interval of x. Since any transition point is a point on the square lattice, $UB - LB \leq 1/(m-1) < 2/m$. Therefore, at the mth sampling time, there can be only one new transition point between UB and LB. Let p be the predictor value at time m. Note that if $p/m \notin (LB, UB)$, then the mth bit does not change the quantization interval for x from that at time $m - 1$ and hence (LB, UB) remains as the quantization interval for x. On the other hand, if $p/m \in (LB, UB)$, then it is a new transition point and the quantization interval specified by (LB, UB) is split into two new quantization intervals with x belonging to one of them. Specifically, if $SEQ(m) = 1$, then $LB \leftarrow p/m$ and the new quantization interval for x becomes $(p/m, UB)$, and if $SEQ(m) = 0$, $UB \leftarrow p/m$ and the new quantization interval for x becomes $(LB, p/m)$. In both cases the end points of the interval are transition points. This proves that algorithm 1 is indeed the optimal filtering algorithm.

The above procedures can be readily used to quantify the MSE of the optimal filter. In order to compare the MSE of the optimal filter to conventional linear filters, we need to find the correspondence of estimates produced by each filter to the quantization intervals. Figure 7 illustrates this correspondence for the counter (i.e., rectangular), triangular and MSE optimal linear [6] filters. The figure depicts the quantization intervals and their centroids, which are produced by the optimal filter, for $m = 4$, and the estimates for each filter and their correspondence to the quantization intervals. The figure clearly illustrates the source of suboptimality of linear filters as an interval is not necessarily mapped into its centroid and may in fact be mapped into an estimate that is outside the interval itself.

The Mean Square Error (MSE) is a measure of average noise power, and implies the overall accuracy of the quantization system. Figure 8(i) compares the MSE for the optimal filter to the MSEs for the counter, triangular and optimal linear [6] filters. In plotting the MSE for each linear filter we corrected for its systematic offset and gain biases using the minimum MSE affine transformation found by computing the coefficients a and b that minimize

$$D = E\big[(x - \hat{x})^2\big] = \sum_{i=1}^{n} (x_i^* - a\hat{x}_i - b)^2 \Delta_i,$$

where $\{x_i^*\}_{i=1}^n$ and $\{\Delta_i\}_{i=1}^n$ are the centroids and lengths of the quantization intervals, respectively, and $\{\hat{x}_i\}_{i=1}^n$ are the estimates corresponding to the quantization intervals provided by the linear filter. Note that the optimal filter does not suffer from any systematic offset or gain biases. Also note that the outputs of all the filters considered here converge to the input value. Their convergence rates, however, differ. As expected, the counter has the worst convergence rate. For small m, the triangular filter is worse than the optimal linear filter, but as m increases, their MSEs become very close. Note that for example, in order to achieve a performance similar to a 10-bit uniform ADC one can use the optimal filter and operate the $\Sigma\Delta$ data converter with half the speed of the case when a conventional filter is used.

Our framework can also be used to completely characterize the transfer function for first-order $\Sigma\Delta$ data conversion system, which can provide further insight into the behavior of various filters. As an example, in Figure 9 we plot the transfer functions for first-order $\Sigma\Delta$ data converter using the optimal filter, the optimal linear filter and the rectangular filter for $m = 7$.

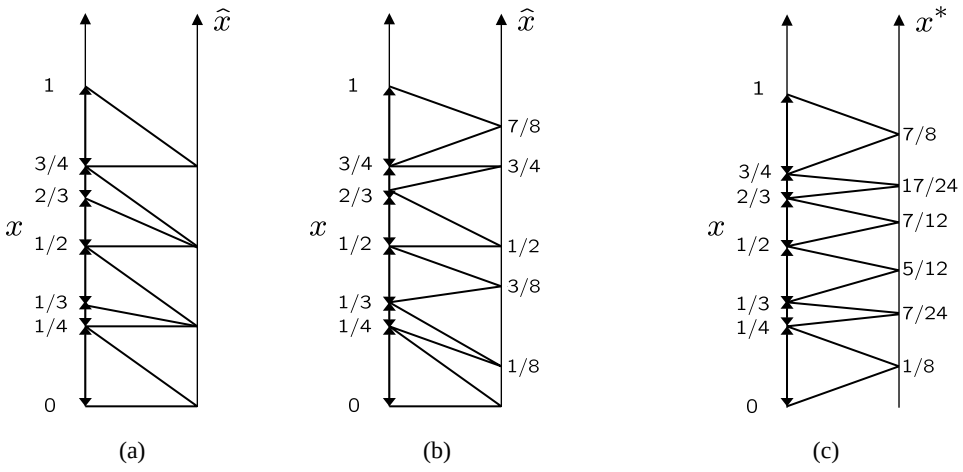

Fig. 7. Quantization intervals and corresponding estimations for $m = 4$ using (a) counter filter (b) triangular filter and (c) optimal filter.

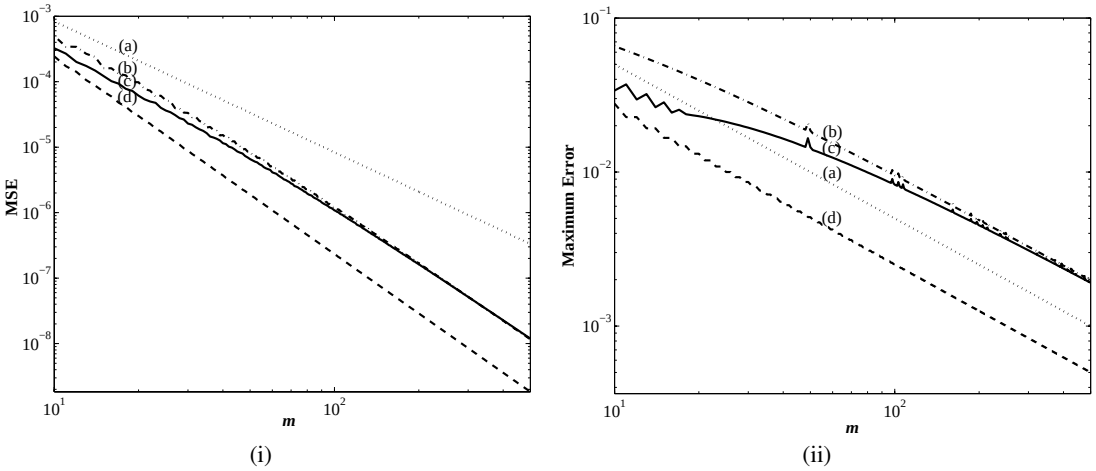

Fig. 8. (i) MSE and (ii) Maximum error behavior as a function of oversampling ratio m, with gain/offset correction using (a) counter filter (b) triangular filter (c) optimal linear filter and (d) optimal filter.

Note that the large distortion at the two ends of the $\Sigma\Delta$ data converter can be intolerable in some applications. For example, in imaging applications [15], it is very important to achieve low distortion at low input signal values (corresponding to low light).

Maximum error is another measure of distortion, which is very important when absolute accuracy is needed. If the application allows the designer to adjust the input range to $x \in [1/m, 1 - 1/m]$, instead of $[0, 1]$, the large distortion at the two ends of the range (see Figure 9) can be avoided, resulting in a factor of two reduction in absolute error with very small decrease in the input range of $2/m$. Figure 8(ii) plots the maximum error in the range of $x \in [1/m, 1 - 1/m]$ for different filters versus the oversampling ratio m. As can be seen from the plots, the optimal filter can achieve significantly lower absolute error than linear filters for the same oversampling ratio. Alternatively, it can achieve the same absolute error at a much lower oversampling ratio. For example, to achieve an absolute error equal to that of a uniform 10-bit ADC, a $\Sigma\Delta$ data converter with optimal filtering requires half the oversampling ratio of a counter filter. It is also interesting to note that the counter outperforms the optimal MSE linear filter in terms of absolute accuracy.

Another consequence of our framework is the ability to characterize the Directed Acyclic Graph (DAG) of the first-order $\Sigma\Delta$ encoder, i.e., the graph representing the sequence of "larger/smaller" comparisons effectively performed by the encoder. This graph may help in

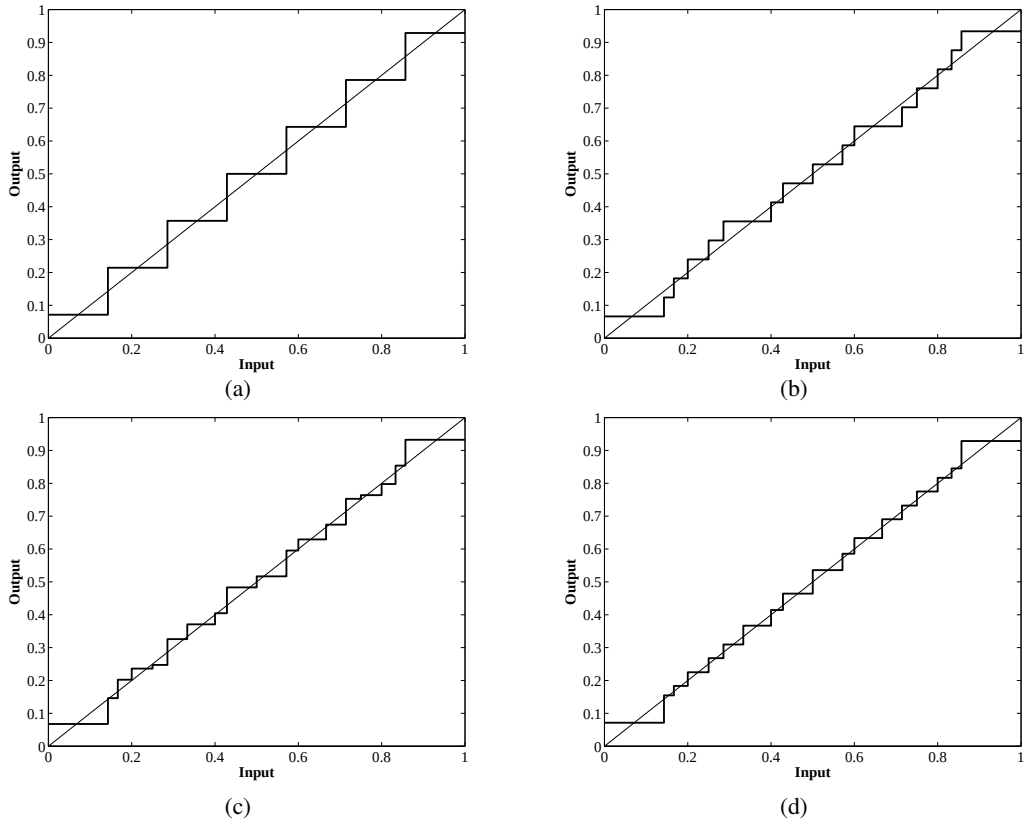

Fig. 9. Transfer functions using (a) counter filter after gain/offset correction (b) triangular filter after gain/offset correction (c) optimal linear filter after gain/offset correction and (d) optimal filter, for $m = 7$.

studying the redundancy in sequences produced by the encoder. It is useful for finding transition points for higher order $\Sigma\Delta$ data converter, as we shall see in the next section. The DAG can be found using $\mathcal{O}(m^2)$ algorithm 2.

Figure 10 plots the DAG for $m = 7$. The horizontal axis corresponds to the sampling times and the vertical axis corresponds to the comparison threshold. The nodes correspond to the effective comparisons while the edges correspond to the outcomes of the comparisons (an upper edge corresponds to a 1, while a lower edge corresponds to 0). For example, the path defined by 011101 corresponds to all $3/4 < x < 4/5$, and provides the following information about the input: $x < 1,\ x > 1/2,\ x > 2/3,\ x > 3/4,\ x < 4/5,\ x > 2/3$. Note that there are two types of nodes, a solid node corresponding to a comparison where the outcome is unknown, and a hollow node corresponding to a comparison with a known outcome given the previous bits. To explain this difference, note that a path leading to a node defines an upper and a lower bound, UB and LB, on the range of input signals that correspond to this path. If the predictor corresponding to the node lies within the (LB, UB) interval, then the comparison has an unknown outcome. On the other hand, if the comparison threshold lies outside the (LB, UB) interval, then the outcome of the comparison is already known. For example, consider hollow node $(4, 0.5)$. The paths leading to the node correspond to the input ranges $(0.5, 0.66)$ and $(0.33, 0.5)$. In either case, the outcome of the comparison is known.

As explained earlier, given m comparisons, the first-order $\Sigma\Delta$ encoder generates $\mathcal{O}(m^2)$ quantization intervals, which is rather small compared to the possible 2^m quantization intervals. Study of the DAG clarifies this difference. The hollow nodes in the DAG do not create transition points and therefore do not add to the number of quantization intervals. However, it is expected that this redundancy in the effective comparisons improves the robustness of the system to errors. Higher order and multi-bit $\Sigma\Delta$ modulators are usually used to increase the number of quantization intervals and therefore achieve improved error. In the next section we extend the framework to second-order incremental $\Sigma\Delta$.

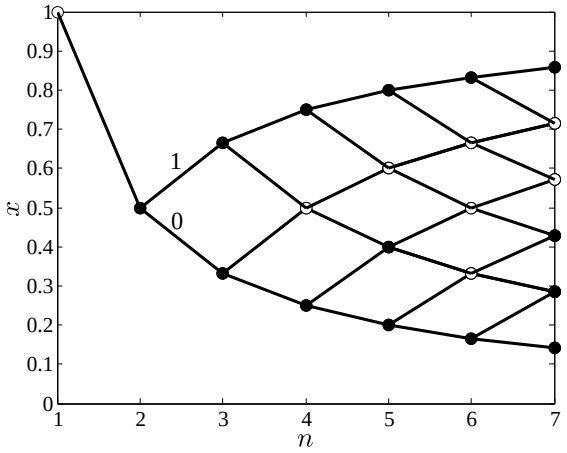

Fig. 10. Directed acyclic graph showing quantization in first-order $\Sigma\Delta$ modulator.

Algorithm 2 Generating the first-order $\Sigma\Delta$ Directed Acyclic Graph

```
begin
SEQ ← [0]
UB ← 1, LB ← 0
q ← 1, s ← 1
mark (q, s) as a hollow node on the graph
q ← 2
CALL DAG(SEQ, LB, UB, q)
end

FUNCTION DAG(SEQ, LB, UB, q)
if q ≤ m then
   Predictor ← Σ_{i=1}^{q-1} SEQ(i) + 1
   s ← Predictor/q
   if LB < s < UB then
      mark (q, s) as a solid node on the graph
      CALL DAG([SEQ; 1], s, UB, q + 1)
      CALL DAG([SEQ; 0], LB, s, q + 1)
   else if s ≥ UB then
      mark (q, s) as a hollow node on the graph
      CALL DAG([SEQ; 0], LB, UB, q + 1)
   else
      mark (q, s) as a hollow node on the graph
      CALL DAG([SEQ; 1], LB, UB, q + 1)
   end if
end if
```

III. Second-order Incremental $\Sigma\Delta$

In this section we study the discrete time second-order $\Sigma\Delta$ quantizer. We focus on the second-order modulator depicted in Figure 11. Our results, however, can be easily extended to a much broader set of architectures. It consists of two integrators followed by a comparator, the output of which is fed back to the input of both integrators. We assume that the comparator threshold is 0 and the input signal $x \in [-1, 1]$. Note that unlike the first-order case the output of the comparator $b(n) = \mathrm{sgn}(u(n)) \in \{-1, 1\}$. We also assume zero initial states, i.e., $u(0) = v(0) = 0$, corresponding to an incremental second-order modulator. Note that the first two bits generated by the modulator are 1 and -1, respectively, independent of the input. To achieve good performance, the modulator gain parameters a_1 and a_2 are chosen so that the output of the second integrator $v(n)$ is as close to 0 as possible for all x and $1 \le n \le m$. In the following discussion we assume a conventional choice of $a_1 = 1/2,\ a_2 = 2$.

The outputs of the two integrators (the states) are given by

$$\begin{aligned} u(n) &= \frac{1}{2}\left(x - b(n)\right) + u(n-1), \\ v(n) &= 2\left(u(n-1) - b(n-1)\right) + v(n-1). \end{aligned} \quad (2)$$

which results in

$$v(n) = xn(n-1)/2 - \sum_{1}^{n-1}(n-i+1)b(i). \quad (3)$$

Similar to the first-order case we can characterize the second-order $\Sigma\Delta$ data converter by determining the quantization intervals induced by the modulator. This can be done by finding the transition points using the following equivalent encoder. Referring to (3) the sequence of comparisons $v(n) \gtrless 0$, $1 \le n \le m$ is the same as the sequence of comparisons of the predictor $\sum_{i=1}^{n-1}(n-i+1)b(i)$ to the equivalent ramp $xn(n-1)/2$, which is quadratic in n. They both represent the same effective comparisons, $x \gtrless 2(\sum_{i=1}^{n-1}(n-i+1)b(i))/(n^2-n)$. Therefore the output of the modulator is identical to the output of an encoder that performs comparisons between the predictor and the equivalent ramp. Figure 12 depicts this equivalence graphically; the dotted line is the integrator waveform, the solid line is the equivalent ramp, and the dashed line is the predictor.

The optimal filtering algorithm for the second-order modulator is essentially any filter that produces the centroid corresponding to the quantization interval. Zoomer implementation is the same as Algorithm 1 with the

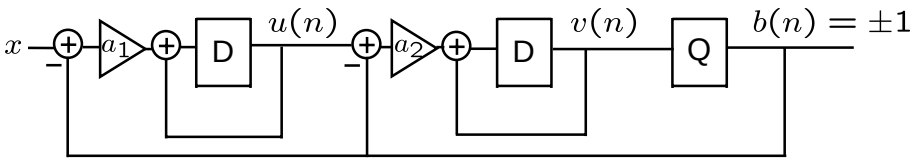

Fig. 11. Block diagram for the second-order $\Sigma\Delta$ modulator. D refers to the delay element and Q refers to the one-bit quantizer.

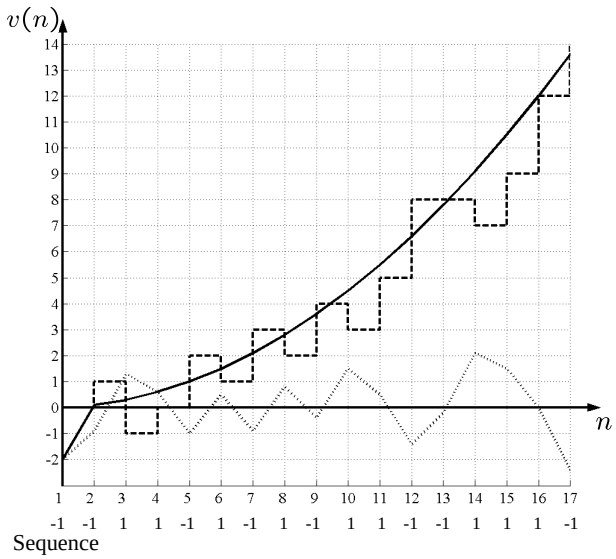

Fig. 12. The equivalent ramp and its corresponding predictor. The dotted line is the integrator output, the solid line is the equivalent ramp, and the dashed line is the predictor.

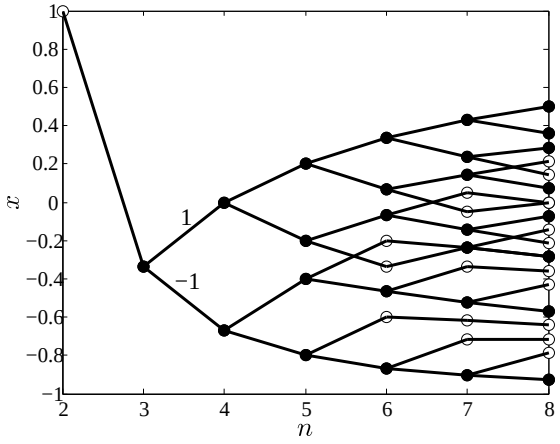

Fig. 13. Directed graph showing quantization in the second-order $\Sigma\Delta$ modulator.

updating of the predictor replaced with that for the second-order predictor.

At any sample time, n, the set of possible values the second-order predictor can assume consists of every other integer in the interval $[-(n-1)(n-2)/2 + 2, (n-1)(n-2)/2]$ on the square lattice as depicted in Figure 12. However, unlike the first-order case, not all such integers correspond to transition points, and therefore, to compute the set of transition points, we find and parse the directed acyclic graph of the second-order $\Sigma\Delta$ modulator.

To find the DAG for the second-order modulator, we use Algorithm 3, which is similar to Algorithm 2. We replace the predictor and the effective comparison for the first-order modulator in Algorithm 2 by that for the second-order, taking into consideration the fact that the second-order modulator produces $+1$s and -1s instead of 0s and 1s. Note that in Algorithm 3 q starts from 2 because of the two delay elements before the quantizer. Figure 13 plots the second-order DAG for $m = 8$. For example, the path defined by $1, -1, -1, -1, 1, -1, -1, -1$ corresponds to all $-8/10 < x < -2/3$, and provides the following information about the input: $x \geq 0$, $x < 1$, $x < -1/3$, $x < -2/3$, $x \geq -4/5$, $x < -3/5$, $x < -13/21$, $x < -9/14$. Again there are two types of nodes, a solid node corresponding to a comparison where the outcome is unknown and a hollow node corresponding to a comparison with a known outcome given the previous bits. The solid nodes correspond to transition points. Figure 14 plots all the transition points and their corresponding equivalent ramps for $m = 7$.

As in the first-order case, the derivation of the transition points can be used to determine the transfer function of the second-order converter. To compare the performance of filters such as the Kaiser or the natural linear filter [19] to that of the optimal filter, in Figure 15 we plot the transfer functions using these different filters. Note that the linear filters considered are again biased and we have corrected for the bias as before. Although the outputs of the linear filters converge to the correct input, they perform poorly compared to the optimal filter. In fact, as can be seen in Figure 15(b), the transfer

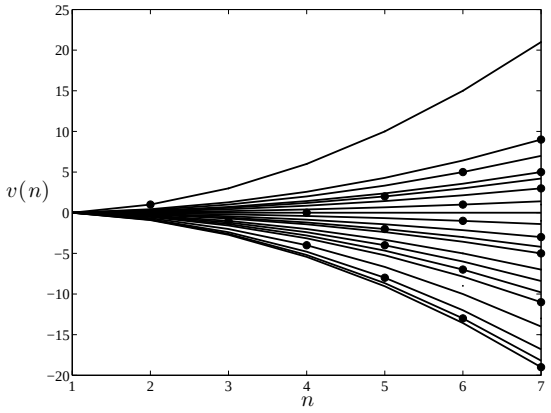

Fig. 14. second-order transition points and their corresponding ramps

Algorithm 3 Generating the second-order $\Sigma\Delta$ Directed Acyclic Graph

```
begin
SEQ ← [1; −1]
UB ← 1, LB ← 0
q ← 2, s ← 1
mark (q, s) as a hollow node on the graph
q ← 3
CALL DAG(SEQ, LB, UB, q)
end

FUNCTION DAG(SEQ, LB, UB, q)
if q ≤ m then
   Predictor ← Σ_{i=1}^{q−1} (q − i + 1)SEQ(i)
   s ← Predictor/q(q − 1)
   if LB < s < UB then
      mark (q, s) as a solid node on the graph
      CALL DAG([SEQ; 1], s, UB, q + 1)
      CALL DAG([SEQ; −1], LB, s, q + 1)
   else if s ≥ UB then
      mark (q, s) as a hollow node on the graph
      CALL DAG([SEQ; −1], LB, UB, q + 1)
   else
      mark (q, s) as a hollow node on the graph
      CALL DAG([SEQ; 1], LB, UB, q + 1)
   end if
end if
```

function of the Kaiser filter is not even monotonic in the input. Also note that there is a significant systematic gain error in the natural filter shown in 15(a). Also in Figure 15 there is a large quantization interval at the high end for all the filters considered. This large interval is due to the fact that the first two bits generated by the modulator are 1 and -1 independent of the input and results in an asymmetric transfer function. Since this is a systematic problem that causes large distortions, in what follows we shall eliminate this large interval.

The transition points can be used to quantify the MSE and maximum error of the optimal filter and to compare it to that of conventional filters. Figure 16 plots the MSE and maximum error of the Kaiser low pass filter with $\beta = 2.5$, the natural filter and the optimal filter as a function of m. Note that the performance gain of the optimal filter over linear filters for second-order $\Sigma\Delta$ converter is more than that of the first-order. For example, in order to achieve performance comparable to a 10-bit ADC, a second-order $\Sigma\Delta$ converter with optimal filtering can be operated at $1/5$th the speed of that using a conventional filter.

IV. Applications

In previous sections we presented a framework for studying incremental $\Sigma\Delta$ data converters and used it to compare the performance of the optimal filter to that of conventional filters. In this section, we use our framework and the results of the previous sections to explore the potential advantages of incremental $\Sigma\Delta$ data converters over conventional $\Sigma\Delta$ for two classes of applications. We first investigate sensor applications where the input varies very slowly and can be considered constant over the conversion period. In such applications, incremental $\Sigma\Delta$ is commonly used to prevent idle tones. We show that the power consumption using the optimal filter can be significantly lower than using conventional linear filters. Next we compare the performance of incremental $\Sigma\Delta$ modulators with optimal filtering to that of conventional $\Sigma\Delta$ systems for bandlimited input signals. We show that for high oversampling ratios, incremental $\Sigma\Delta$ systems achieve higher SNDR.

Our results assume ideal, noise-free circuits. Finding a computationally tractable optimal filter in the presence of noise is an open problem. We briefly discuss how our results may still be useful in practice.

Sensor Applications

Recently, $\Sigma\Delta$ data converters have become popular in sensing applications such as imaging [14], [15], [16], [17] and accurate temperature sensing [18]. In such

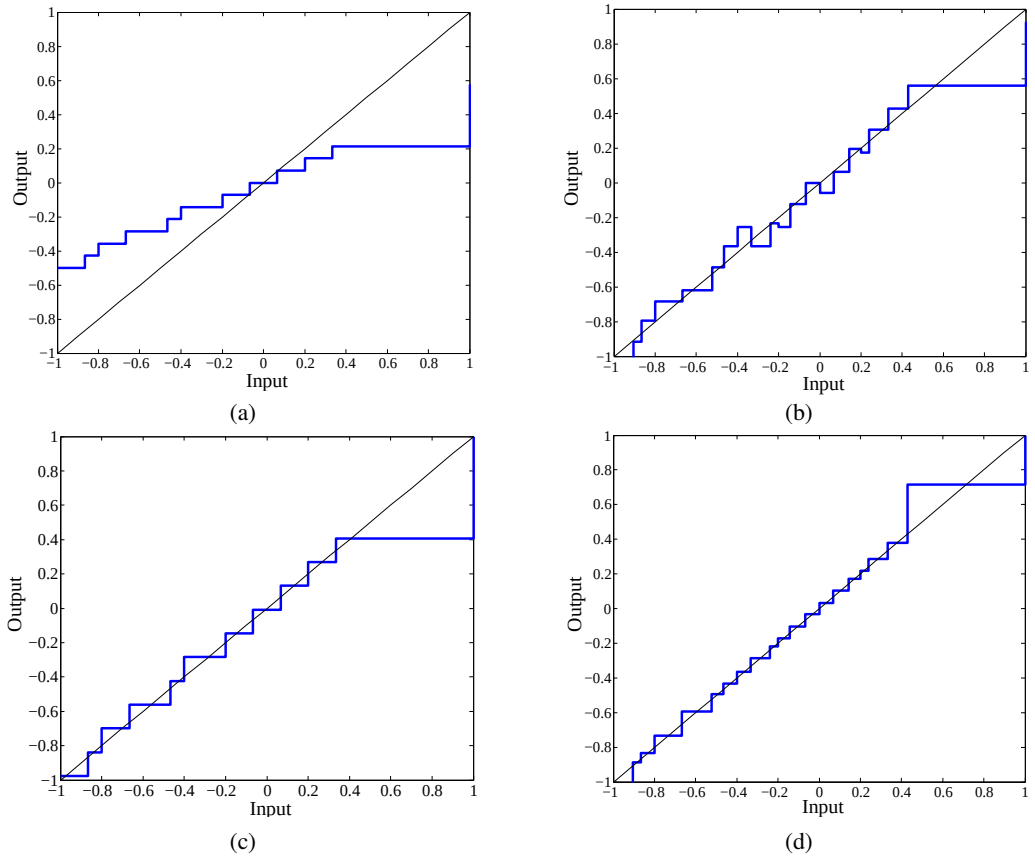

Fig. 15. Transfer functions using (a) natural linear filter without gain/offset correction (b) Kaiser filter with $\beta = 2.5$ with gain/offset correction (c) natural linear filter with gain/offset correction and (d) optimal filter, for $m = 7$.

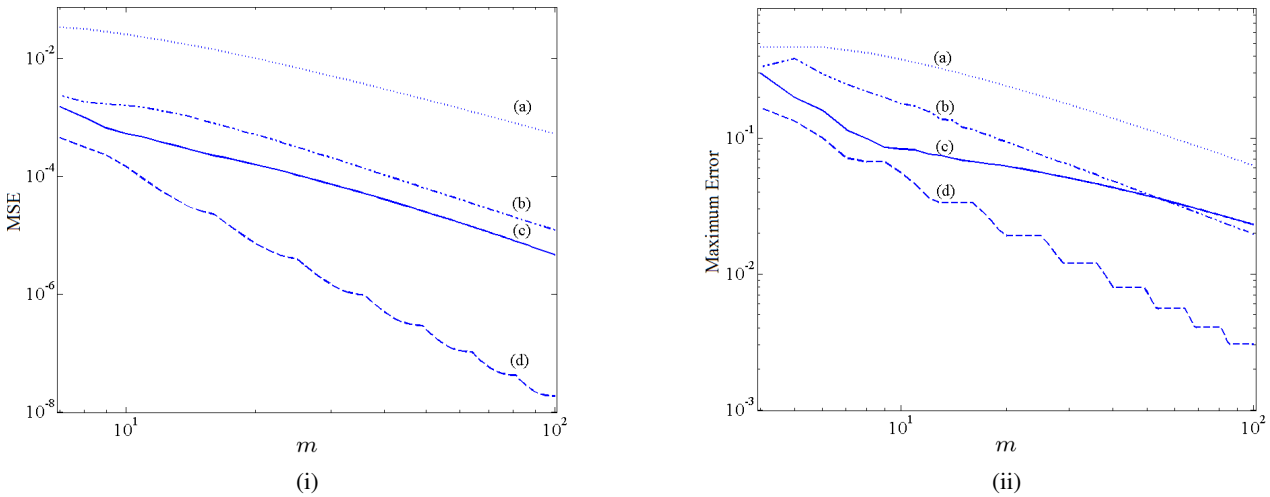

Fig. 16. (i) MSE and (ii) Maximum error behavior as a function of oversampling ratio m, using (a) natural linear filter without gain/offset correction (b) Kaiser filter with $\beta = 2.5$ with gain/offset correction (c) natural linear filter with gain/offset correction and (d) optimal filter.

applications, the input signal is slowly varying with time, and thus can be viewed as constant over the conversion period. One of the main advantages of using incremental $\Sigma\Delta$ in such applications is that they do not suffer from idle tones; thus the measurements are repeatable and a static transfer function exists. This is because in an incremental system the same conversion is performed for all samples and, therefore, all filtered values are the same regardless of the filter used, while in a conventional $\Sigma\Delta$ system, the integrator residues vary with the samples creating periodic tones. Figure 17 compares the Power Spectral Density (PSD) of the output signal of an incremental second order $\Sigma\Delta$ modulator with optimal filter to that of a conventional nonincremental system.

An important limiting factor in the design of many sensor interface circuits is data converter power consumption. We show through an example that the power consumption of an incremental $\Sigma\Delta$ converter can be significantly reduced by using the optimal filter instead of a conventional linear filter. Consider the incremental $\Sigma\Delta$ data converter reported in [18] for use in a temperature sensing system. The converter, which is implemented in 0.5μm CMOS technology, operates at 10 samples/sec with oversampling ratio of 400 and consumes 400μW. As discussed at the end of Section III, the same accuracy can be achieved at $1/5$th the oversampling ratio if we use the optimal filter instead. Since most of the power is dissipated in the modulator, the power consumption of the modulator in this case would be reduced to below 100μW. This reduction, however, is achieved using more complex signal processing, so to make the power consumption comparison more accurate one must take into account the additional power consumed by the filter. To estimate the power consumed by the optimal filter, note that it requires less than 4 multiplications per sample to implement. Assuming that each multiplication consumes $4p$J (see [24]), the total power consumption of the filter is in the order of $10n$W, which is significantly lower than the 100μW consumed by the modulator. Moreover, the power consumed by the optimal filter in this example is not larger than that of a linear filter even though it requires more computations per sample, because the optimal filter operates at $1/5$th the frequency of the linear filter.

The above example assumed ideal circuit implementation. In practice $\Sigma\Delta$ modulators are designed to be thermal noise limited to save capacitor area and mitigate the effect of idle tones [9]. Using the optimal filter in a thermal noise limited design is not possible, as it would fail on too many modulator output sequences. However, since the incremental architecture does not suffer from idle tones, a quantization noise limited design can be used instead. In such a design, sufficiently large modulator capacitor sizes would be used to minimize thermal noise and a modified optimal filter such as the filter presented in [8] would be used. The modulator can then be operated at a lower speed than the thermal noise limited modulator with a linear filter to achieve the same MSE performance at a lower power consumption slightly higher than the discussed example.

Incremental $\Sigma\Delta$ for Bandlimited Signals

We now compare an incremental $\Sigma\Delta$ data converter to a conventional design for bandlimited input signals. Figure 18 depicts the two systems. The input signal in the conventional system is sampled at rate f_1, which is much faster than the Nyquist rate of the input signal and fed into a $\Sigma\Delta$ modulator operating at the same rate. The output sequence is decimated at rate f_2, which is slightly higher than the Nyquist rate of the input signal. Thus, the oversampling ratio is f_1/f_2. In the incremental $\Sigma\Delta$ system, the signal is sampled at rate f_2 and each sample is fed into the $\Sigma\Delta$ modulator, which operates at rate f_1. The modulator is reset after each conversion, i.e., at rate f_2. The output sequence is filtered using the optimal filter.

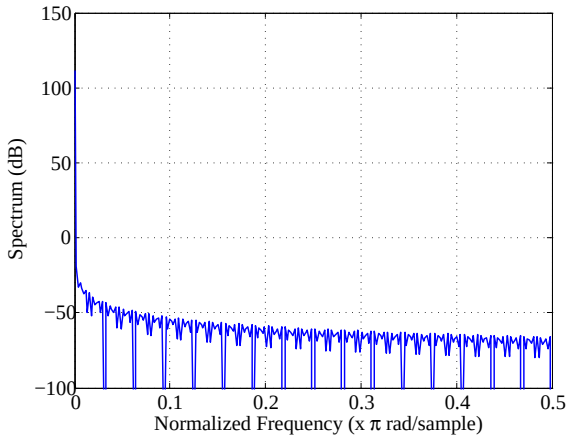

Fig. 19. Spectrum of Nuttall Window.

To compare the two systems, we apply a sinusoidal signal to each system and compute its output PSD and its SNDR. Figure 20 compares the PSDs for a conventional second-order $\Sigma\Delta$ converter with 2500-tap Remez low pass filter with a cut-off frequency of 40KHz and an incremental second-order $\Sigma\Delta$ system that uses the optimal filter. The results are for a sinusoidal input signal with amplitude $a = 0.5$ and frequency $f \approx 6600$Hz,

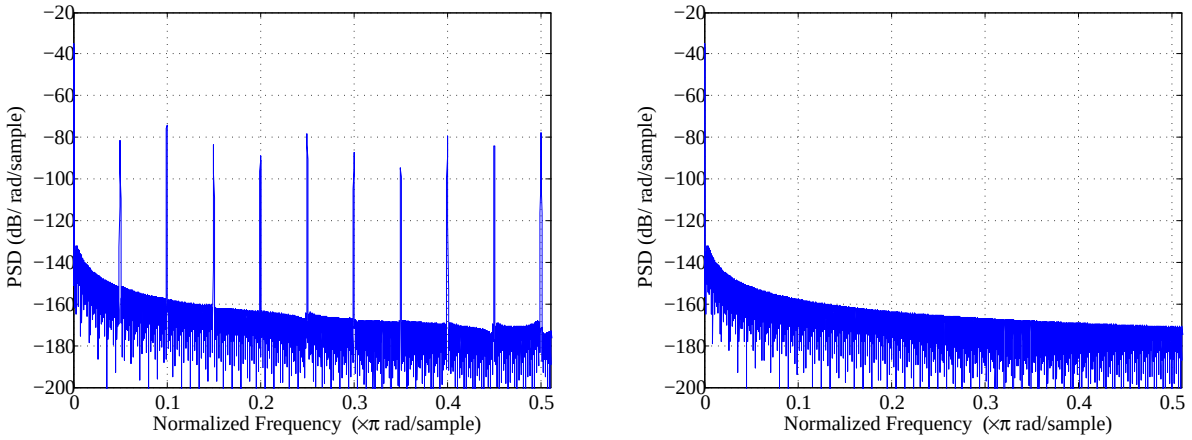

Fig. 17. Power Spectral Densities for (a) a conventional second-order $\Sigma\Delta$ system and (b) an incremental second-order $\Sigma\Delta$ system using the optimal filter and the same modulator. Nuttall Window is applied in both cases. The input is set to a constant value of .001 and ADC is performed at performed at a rate of 200ksamples/sec and the oversampling ratio $m = 100$. A 2500-tap Remez low pass filter with a cut-off frequency of 40KHz is used for the conventional system.

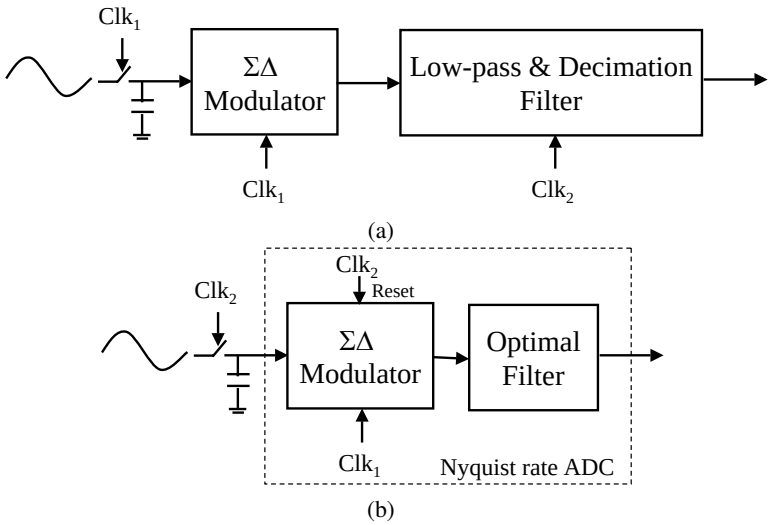

Fig. 18. (a) Conventional scheme (b) Incremental with optimal filter.

modulator frequency sampling rate $f_1 = 20$MHz and Nyquist rate $f_2 = 200$KHz. A Nuttall window [25] with the spectrum shown in Figure 19 is used. The SNDR for the incremental $\Sigma\Delta$ system is 97dB versus 90dB for the conventional system. Another performance measure of interest is the maximum tone power, which is the power of the largest tone created by each system. The maximum tone power of the incremental system is 5dB below that of the conventional system.

Figure 21 shows the PSD of the two systems when a sinusoid with an amplitude smaller than its dc value is applied. Note that the incremental system with optimal filter only has distortion as harmonics of the sinusoid and that the maximum tone power in the conventional system is 10dB higher than the maximum harmonic power in the incremental system.

V. Conclusion

We introduced a quantization-theoretic framework for analyzing incremental $\Sigma\Delta$ quantizers and used it to characterize the quantization intervals and hence transfer functions for first and second-order incremental $\Sigma\Delta$ modulators. We then used the computed quantization intervals and their corresponding sequences to characterize the performance of the optimal filter and conventional linear filters. We also showed that incremental $\Sigma\Delta$ with optimal filtering can outperform conventional $\Sigma\Delta$ systems for bandlimited input signals in terms of SNDR.

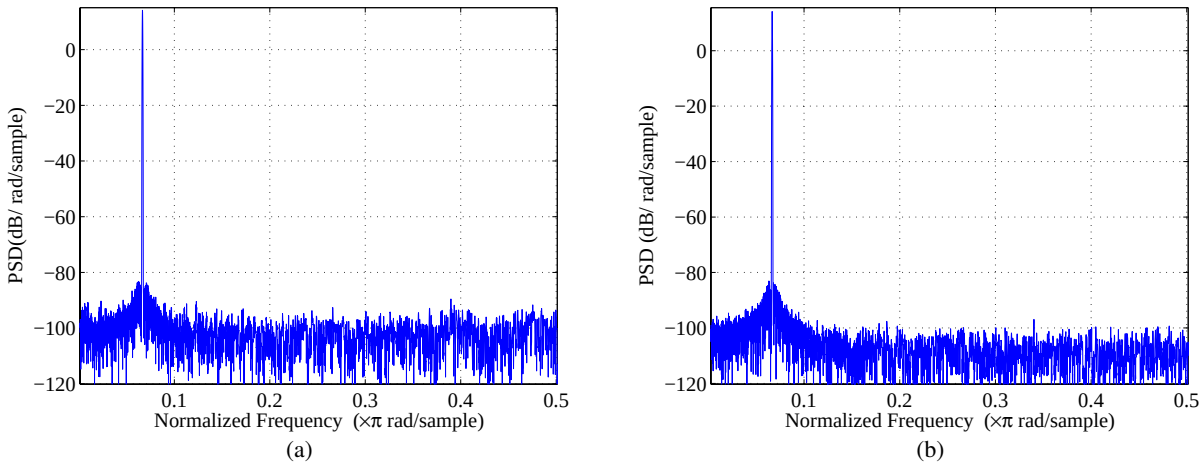

Fig. 20. Power Spectral Densities for (a) a conventional second-order $\Sigma\Delta$ system and (b) an incremental second-order $\Sigma\Delta$ system using the optimal filter and the same modulator. Nuttal Window is applied in both cases. The modulator sampling rate $f_1 = 20\text{MHz}$ and the Nyquist rate $f_2 = 200\text{KHz}$. The input frequency $f \approx 6600\text{Hz}$. A 2500-tap Remez low pass filter with a cut-off frequency of 40KHz is used for the conventional system.

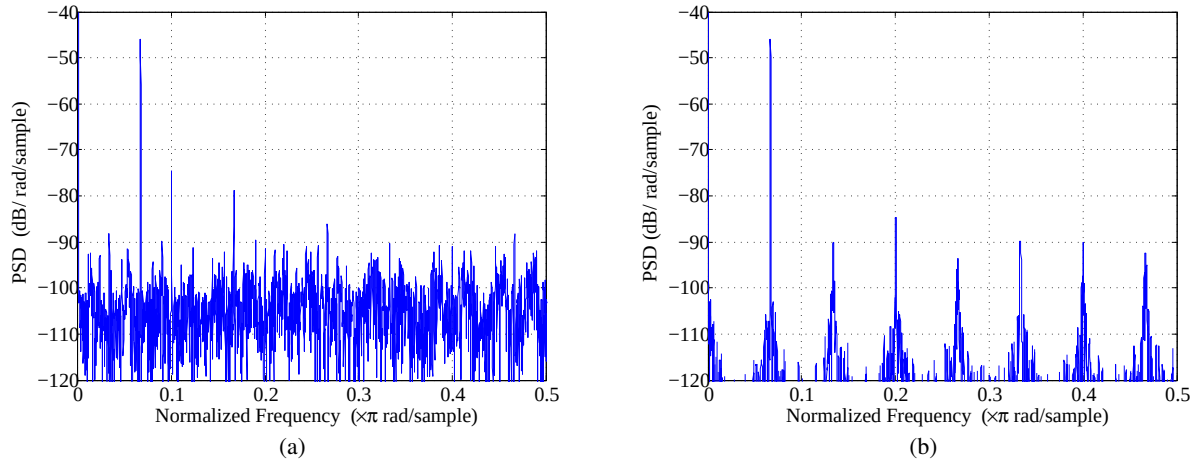

Fig. 21. Power Spectral Densities for (a) a conventional second-order $\Sigma\Delta$ system and (b) an incremental second-order $\Sigma\Delta$ system using the optimal filter and the same modulator. Nuttall Window is applied in both cases. The input is set to a sinusoid with amplitude of 0.0005 and dc value of 0.001. Modulator sampling rate $f_1 = 20\text{MHz}$ and Nyquist rate $f_2 = 200\text{KHz}$. Input frequency $f \approx 6600\text{Hz}$. A 2500-tap Remez low pass filter with a cut-off frequency of 40KHz is used for the conventional system.

Further, we showed that incremental $\Sigma\Delta$ systems do not suffer from idle tones. Even though we considered only single-bit first and second order modulators in this paper, our framework can be readily extended to higher order and multi-bit modulators.

In practice, $\Sigma\Delta$ quantization system designers improve performance by increasing the oversampling ratio and/or using more complex modulators. However, high oversampling ratio results in high system power consumption. Our results show that the oversampling requirements can be relaxed by using optimal filtering and that the gain is more pronounced for higher order incremental $\Sigma\Delta$. The optimal filter requires more complex signal processing than linear filters; however, as discussed the additional power consumed in the digital processing should be negligible compared to the savings in the analog front end, especially in scaled CMOS implementations.

Our analysis assumed ideal circuit components. In practice, nonidealities such as temporal noise, offsets, and nonlinearities of the analog components limit the attainable performance. In conventional $\Sigma\Delta$ systems, temporal noise is averaged out by the filter, and nonlinearities and gain-bandwidth ultimately limit the perfor-

mance [26]. Recently, it was demonstrated that ADCs can achieve a much higher performance by digitally correcting for the nonlinearities [27]. There is a need to develop such digital techniques for correcting nonlinearity, reducing noise, and calibrating for loop-gain variations in $\Sigma\Delta$ modulators. In particular, the factors in the optimal filter need to constantly track the loop gains that vary due to component mismatch and $1/f$ noise.

The time-domain analysis carried out in this paper may prove useful in deriving such filters. Only after developing such techniques and the needed circuits can the potential power savings of the proposed incremental $\Sigma\Delta$ systems be fully assessed.

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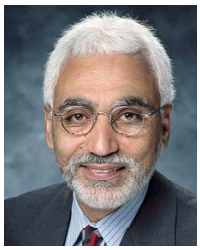

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