

**On the Design of Stable, High Performance Sigma Delta
Modulators**

by

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The following individuals certify that they have read, and recommend to the Faculty of Graduate and Postdoctoral Studies for acceptance, the thesis entitled:

On the Design of Stable, High Performance Sigma Delta Modulators

submitted by **Brett Christopher Hannigan** in partial fulfillment of the requirements for the degree of **Master of Applied Science in Biomedical Engineering**.

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Abstract

This document provides brief instructions for using the `ubcdiss` class to write a **UBC!**-conformant dissertation in \LaTeX . This document is itself written using the `ubcdiss` class and is intended to serve as an example of writing a dissertation in \LaTeX . This document has embedded **URL!**s (**URL!**s) and is intended to be viewed using a computer-based **PDF!** (**PDF!**) reader.

Note: Abstracts should generally try to avoid using acronyms.

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Lay Summary

The goal of this work was a method to better design analog-to-digital converters with special interest to recording weak bio-signals, such as those from electroencephalography and electrocardiography.

The sigma delta architecture of analog-to-digital converters is known for having high resolution for signals of this class while requiring fewer expensive analog circuit components. However, as its performance is increased, it tends to become unstable, a point at which the digitized signal no longer accurately represents the original.

To this end, a theory and set of software tools were developed that use mathematical optimization and control theory to design sigma delta circuits with varying degrees of performance and stability. It is even possible to generate a design that is guaranteed to be stable. The method is generalizable to any kind of signal, medical or otherwise. These developments were used to analyze and synthesize designs and will hopefully inspire future high-resolution analog-to-digital converters.

Preface

At **UBC!**, a preface may be required. Be sure to check the **GPS!** guidelines as they may have specific content to be included.

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List of Symbols

K Variable quantizer gain (linearized model).

P_Q In-band quantization noise power.

$S(\lambda)$ Sensitivity function.

$T(\lambda)$ Complementary sensitivity function.

Δ_Q Quantization step size.

λ Placeholder for the continuous-time Laplace variable s or discrete-time z -transformation variable z .

d Quantization noise source (linearized model).

e Feedback error signal.

n Filter order.

r Analog reference input signal.

u Quantizer input signal.

y Digital bitstream output signal.

Glossary

A/D analog-to-digital.

AAF antialiasing filter.

BIBO bounded-input bounded-output.

CLANS closed-loop analysis of noise shaper.

CT continuous-time.

D/A digital-to-analog.

DRF digital reconstruction filter.

DSP digital signal processing.

DT discrete-time.

ECG electrocardiography.

EEG electroencephalography.

FIR finite impulse response.

GKYP generalized Kalman-Yakubovič-Popov.

IIR infinite impulse response.

LF loop filter.

LFT linear fractional transformation.

LMI linear matrix inequality.

NTF noise transfer function, equivalent to the sensitivity function.

OSR oversampling ratio.

PPG photoplethysmography.

S/H sample-and-hold.

SQNR signal-to-quantization-noise ratio.

STF signal transfer function, equivalent to the complementary sensitivity function.

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Chapter 1

Introduction

The conversion of signals between analog and digital domains is an often encountered problem in signal processing. For an analog signal to be represented digitally, it must undergo the processes of sampling and quantization (Figure 1.1). The former is the conversion from continuous-time (CT) to discrete-time (DT) and can be done without loss of information by the Nyquist-Shannon sampling theorem, given a sufficiently high sample rate. The latter is the mapping from an infinite set of possible values to a finite number of quantization levels. Unlike sampling, the process of quantization is non-injective and thus irreversible. The design of signal conversion circuits that minimize the error introduced by quantization is a major problem in mixed signal electronics.

Sigma delta modulation is a widely used technique for analog-to-digital (A/D) and digital-to-analog (D/A) conversion of signals that provides high resolution through the techniques of oversampling and noise shaping. Oversampling trades throughput for resolution, thus the sigma delta modulator generally lies between integrating converters, which are specialized for near-DC signals, and high-speed architectures, such as successive approximation and flash. The sigma delta quantization scheme is especially applicable to signals with low to moderate frequency content. Signals with these properties include most biosignals such as those recorded electrically (electroencephalography (EEG), electrocardiography (ECG)) or through other means using transducers (photoplethysmography (PPG)), as well as audio signals.

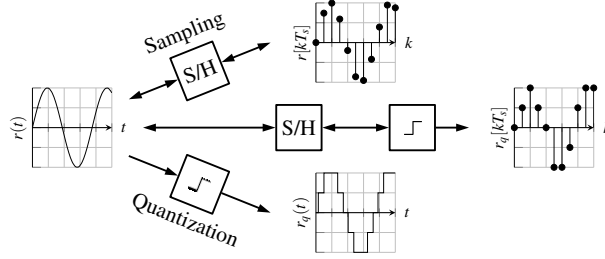


Figure 1.1: A continuous-time, continuous-value signal $r(t)$ is sampled to produce a discrete-time, continuous-value signal $r[kT_s]$. $r(t)$ independently undergoes quantization to yield a continuous-time, discrete-value signal $r_q(t)$. When both processes are applied in sequence, a discrete-time, discrete-value signal $r_q[kT_s]$ is the result.

1.1 Oversampling and Noise Shaping

Oversampling is simply the process where the analog signal is sampled at a rate higher than what the sampling theorem would dictate for perfect reconstruction, expressed as the oversampling ratio (OSR) relative to the Nyquist frequency. It may seem that this does not have a direct benefit *per se*, but it allows a less demanding analog antialiasing filter (AAF) to be used, saving circuit area. It also permits the quantization error to be spread across a larger bandwidth to increase resolution. Assuming quantization error can be modelled by white noise, oversampling reduces the in-band quantization noise power P_Q by a factor directly proportional to OSR [2] as seen in Equation 1.1, where Δ_Q is the difference between quantization levels. These two advantages — reducing analog circuit complexity and increasing resolution — are common goals in sigma delta modulator design.

$$P_Q = \frac{\Delta^2}{12 \cdot \text{OSR}} \quad (1.1)$$

It may appear that oversampling alone quickly becomes impractical because one must approach very high sampling frequencies to increase the signal-to-quantization-noise ratio (SQNR) substantially. However, this assumes that the quantization noise is evenly distributed across the spectrum. Noise shaping is the use of a filter operating on the oversampled signal to push quantization noise out of the signal band where it can be removed by digital reconstruction filter (DRF). This

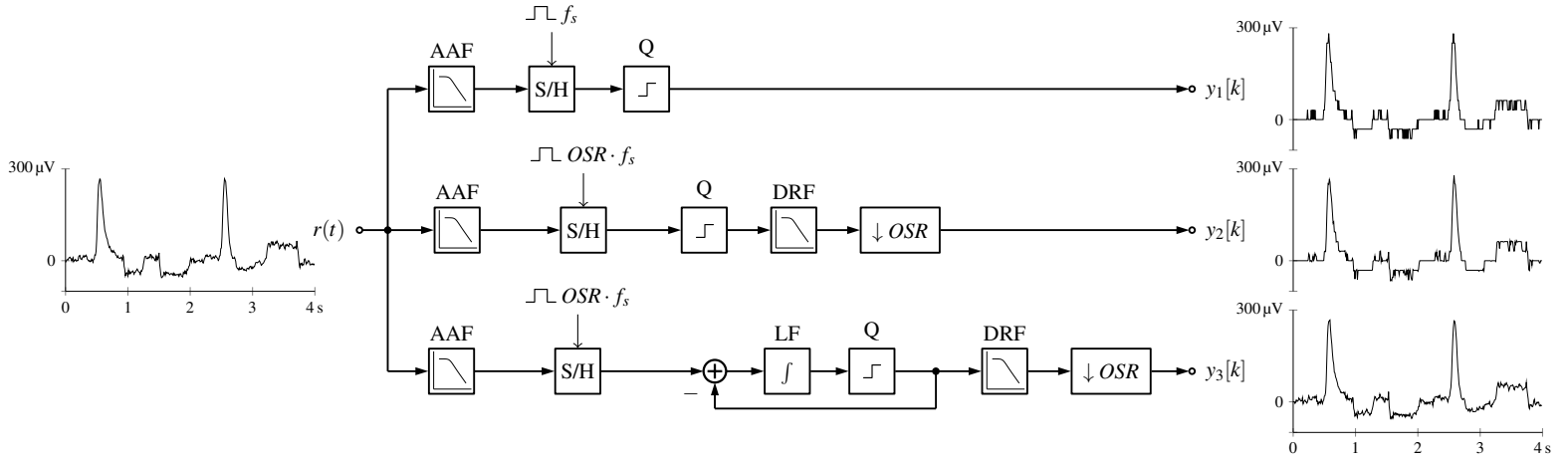


Figure 1.2: A comparison between naïve quantization (top), 10 times over-sampled quantization (middle), and first order sigma delta modulation (bottom). The graphs on the right show the increasing quality of an EEG signal [1] sampled to a final rate of 100 Hz and quantized by Q with 5 bits by each scheme.

behaviour is implemented by wrapping the filter and quantizer in a feedback loop. With the same white noise assumption, the tradeoff between in-band shaped quantization noise and OSR is improved for ideal loop filters when order n is increased as shown in Equation 1.2 [2]. The effect of oversampling and noise shaping is demonstrated in Figure 1.2.

$$P_Q = \frac{\Delta^2 \pi^{2n}}{12(2n+1) \cdot OSR^{2n+1}} \quad (1.2)$$

1.2 Basic Structure

The basic block diagram of a sigma delta modulator and nomenclature that will be used herein is now introduced. For brevity, the scope is limited to sigma delta A/D converters but the concepts are easily transferrable to the D/A case. Modulators can be one of two main classes, CT or DT referring to the nature of the loop filter (LF).

1.2.1 Discrete-Time Modulator

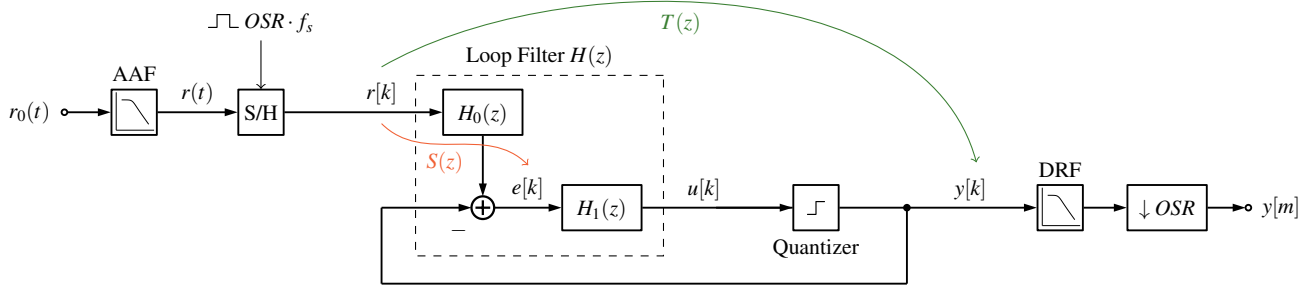


Figure 1.3: The basic block diagram of a DT sigma delta A/D converter.

Consider the DT modulator block diagram shown in Figure 1.3. The analog front-end includes the AAF and sample-and-hold block. This subsystem conditions the input signal $r_0(t)$ and samples it outside the loop to produce DT signal $r[k]$. In the modulator loop, the 2-input 1-output LF operates on $r[k]$ and the feedback signal, producing intermediate signal $u[k]$ with shaped noise. Then, $u[k]$ undergoes quantization producing discrete-value output $y[k]$. The quantizer output is fed back to the LF and also passed along. The final subsystem filters the signal from the shaped noise in the digital domain with a downsampling DRF to yield the final digital output $y[m]$.

From a control systems perspective, there are a couple of transfer functions that will be used to analyze and synthesize loop filters. The sensitivity function $S(\lambda)$, where $\lambda = z$, is known as the noise transfer function (NTF) of the modulator because it shows how the quantization error is filtered in the linearized model. The complementary sensitivity function $T(\lambda)$ is known as the signal transfer function (STF) of the modulator and shows how the signal is transformed by the modulator loop.

1.2.2 Continuous-Time Modulator

For the CT class of modulators, consider the structure of Figure 1.4. Both types are similar except the LF operates directly on analog input $r(t)$ in the CT domain and sampling is done inside the loop. The AAF is no longer necessary in most

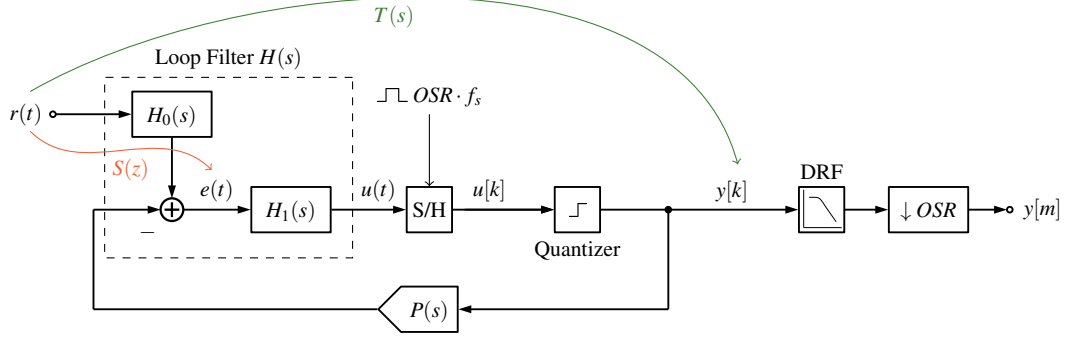


Figure 1.4: The basic block diagram of a CT sigma delta A/D converter.

cases as the LF precedes the sampling block and implicitly attenuates components of the signal that would result in aliasing. Finally, signal $y[k]$ must undergo D/A conversion during feedback, modelled with the pulse transfer function $P(s)$.

The NTF and STF of a CT sigma delta modulator are more difficult to define because they are transfer functions involving both CT and DT signals. The DT equivalence principle states that there is a DT modulator model that exactly describes the CT design at the sampling instants, because the modulator is overall a sampled data system [3, Sec. 3.2]. Thus, DT transfer functions can be derived for this purpose. However, these equivalent transfer functions may be difficult to manipulate due to their dependence on $P(s)$. For the purposes of this analysis, we omit the sampling block during design and use the simplification that $S(\lambda)$ and $T(\lambda)$ are CT ($\lambda = s$) transfer functions mapping $t(t) \rightarrow e(t)$ and $r(t) \rightarrow y(t)$, respectively.

1.3 Loop Filter

Together, quantization and noise shaping permit a coarser quantizer element to be used. A common design pattern is to use a high (> 2) order LF paired with a 1-bit quantizer, which is advantageous from a circuit design perspective because a quantizer with just two levels is inherently linear. In addition, low order sigma delta loops often suffer from spurious tones [4, Sec. 2.6.1]. Unfortunately, as LF order is increased, the tendency of the loop to become unstable does as well. While first

and second order designs are provably stable for DC inputs [5], high order filters require careful design to avoid instability. Ensuring stability while maintaining performance is a difficult task due to the presence of the highly nonlinear quantizer. The nonlinearity makes analysis complicated, a stable linear model does not imply a stable modulator while an unstable model can even result in a stable modulator known as the chaotic type [6].

The design of the noise shaping loop filter is the focus of this thesis. Modelling the loop filter as a 2-input 1-output system as shown in Section 1.2 allows the NTF to be determined by $H_1(\lambda)$ alone while the STF can be modified independently with filter $H_0(\lambda)$, without loss of generality:

$$S(\lambda) = \frac{1}{1 - H_1(\lambda)} \quad (1.3)$$

$$T(\lambda) = \frac{H_0(\lambda)}{1 - H_1(\lambda)}. \quad (1.4)$$

A desirable NTF is one that results in a stable linear model, rejects noise in the signal band as much as possible, and has low gain in the out-of-band region to promote stability. The STF is less important as $H_0(\lambda)$ can be interpreted as a pre-filter to modify the STF, but we prefer unity gain in the signal band.

For a first order modulator, a pure integrator can be used as the loop filter $H_0(\lambda)$. For higher orders, it is common to choose a prototype NTF from a family of filters. For example, the popular Delta Sigma Toolbox for MATLAB [4, Appx. B] uses a Chebyshev type II filter for this purpose. The choice of filter greatly affects the stability of the loop, so the traditional design procedure involves extensive simulation under varying input conditions to ensure instability is unlikely during normal operation. Once unstable, the filter states must be reset in order to restore operation. Various schemes to detect the onset of instability [7] and avoid it with gain scaling [8], internal linear feedback [9], and automatic resetting schemes [10].

1.4 Related Works

Optimization techniques have been used to design NTFs with more degrees of freedom than those made with a single filter prototype. A simple example is that from

[4, Sec. 4.3], where the zeros of the prototype NTF are optimized by approximating the integral of the NTF in the pass-band, then minimizing it analytically by equating its derivative to zero. The procedure results in an optimal spreading of zeros across the signal bandwidth for the given NTF poles. One of the first optimization-based approaches to NTF design was the closed-loop analysis of noise shaper (CLANS) methodology that minimizes P_Q under the white quantization noise assumption [11]. This is done using nonlinear optimization to find stable NTF pole locations that minimize the accumulation of quantization error subject to some stability and realizability constraints.

Using the principles from \mathcal{H}_∞ control and its associated linear matrix inequality (LMI) methods, one can define the quantizer as a very simple feedthrough plant and introduce weighting filters on the feedback error signal e , loop filter output u , and quantizer output y to design the loop filter as a controller for various performance and stability constraints [12]. However, the system is bound to the order of the plant augmented with weighting filters and relies on the designer to choose the weights. Choosing weighting filters that are ideal is almost as difficult a task as just choosing the prototype NTF directly. Despite this, if a known AAF or DRF is specified in advance, it may be used as a sort of weighting filter and an optimal LF can be designed around it [13]. Applications for this method could be optimizing the STF to a psychoacoustic model or making use of existing filters in the signal path.

More recently, the generalized Kalman-Yakubovič-Popov (GKYP) lemma has been applied to sigma delta modulator design. The lemma provides a link between a finite frequency domain inequality, such as specifications on the NTF gain, and a linear matrix inequality condition, which can be solved using efficient interior point methods. Using this lemma, the techniques of \mathcal{H}_∞ control can be applied to a transfer function but restricted to a frequency band. This eliminates the need for weighting filters that specify a select band of interest. Unfortunately, the problem becomes non-convex and hard to solve if both poles and zeros are to be optimized simultaneously as is the case with an infinite impulse response (IIR) filter. As a workaround, the NTF poles may be fixed to a prototype design and just the zeros optimized [14], similar to what was described above. Alternatively, a finite impulse response (FIR) NTF form may be assumed [15, 16] then possibly converted to IIR

form using approximate methods such as least-squares or Yule-Walker [17]. Aside from the large delay introduced, the FIR form is not the optimal choice according to [18]. Iterative methods have shown promise in providing a workaround to the non-convexity associated with direct IIR design. A survey of some of these methods is presented in [19] while Table 1.1 summarizes the major contributions of each and differences between them.

Table 1.1: A comparison of some recent work on sigma delta modulator design as a control optimization problem.

Reference	Optimized norms	NTF Type	Performance goal	Stability criteria (see Chapter 3)
Oberoio (2004) [12]	$\mathcal{H}_\infty, \mathcal{H}_2, \ell_1$	IIR	Weighting filters	Uses heuristic bounds on $\mathcal{H}_\infty, \mathcal{H}_2$ norms
Osqui & Megretski (2007) [14]	\mathcal{H}_∞	IIR ¹	GKYP lemma	Not reported
Nagahara & Yamamoto (2012) [15]	\mathcal{H}_∞	FIR	GKYP lemma	ℓ_1 criterion mentioned, but Lee criterion used in design
Li, Yu, & Gao (2014) [20]	\mathcal{H}_∞	IIR	GKYP lemma	Lee criterion
Tariq & Ohno (2016) [16]	$\mathcal{H}_\infty, \mathcal{H}_2, \ell_1$	FIR	Weighting filters	ℓ_1 criterion mentioned but Lee criterion used in design

¹ Only the zeros of the IIR filter are optimized.

1.5 Organization of this Thesis

Having established some background on the workings and nomenclature of a sigma delta modulator, Chapter 2 expands on this to show modifications to the general sigma delta model based on control theory that will permit it to be used in an optimization framework. In Chapter 3, various stability criteria are introduced, ranging

from heuristics to sufficient conditions and their impact on performance. Following the discussion of the role of optimization in loop filter design, Chapter 4 bridges the model and stability criteria chapters by introducing a semidefinite programming framework that supports the aforementioned criteria. The design process is discussed in Chapter 5, with emphasis on simulation results as well as an empirical study of the tradeoff between performance and stability when designing to different criteria. Finally, Chapter 6 concludes the thesis with some discussion about the merits and shortcomings of this method of sigma delta modulator design and possible directions for future work.

Chapter 2

Modelling the Sigma Delta Modulator

In order to apply an optimization framework to the design of the LF, the system from Figures 1.3 and 1.4 must be placed in a form that allows tractable application of the desired performance and stability targets. This includes omission of blocks that have minimal or no effect on the loop as well as linearization of the quantizer. The AAF (when present) can be considered as a pre-filter operating on the input signal. The filter $H_0(\lambda)$ serves as an additional degree of freedom for the STF can be set to unity for the purposes of the model. These two filters are not required in stability analysis, because the NTF depends only on $H_1(\lambda)$ as seen in Equation 1.4. After noise rejection performance has been optimized, $H_0(\lambda)$ can be tuned as necessary to ensure that the combined gain of the AAF and LF is close to unity in the signal band. In a similar way, the digital signal processing (DSP) in the output path serves only to filter out the signal and decimate to the original sampling frequency which may be dealt with separately without impacting loop stability.

2.1 Linearization of the Quantizer Element

Next, the nonlinear nature of the quantizer is dealt with. As mentioned before, a common linearization approach is to replace the quantizer with an additive noise source d . Furthermore, the linear model can incorporate a variable gain K . The

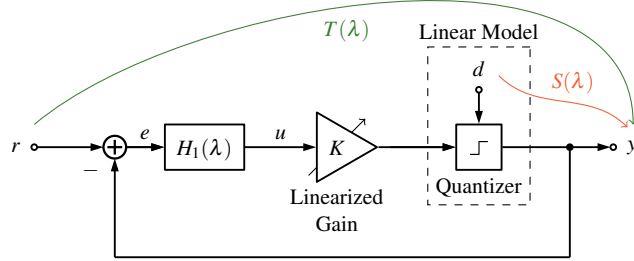


Figure 2.1: The linearized sigma delta loop block diagram with omission of extraneous filters and the quantizer replaced by a variable gain and additive quantization noise signal.

inclusion of K has uses in linearization, stability, and performance that will be expanded upon in Chapter 3. After these simplifications, the block diagram in Figure 2.1 is obtained, which is applicable to DT or CT designs. In the DT case, the loop is operating entirely in the oversampled domain and the sample-and-hold (S/H) block is not shown. In the CT case, the S/H block in the loop is neglected so that $S(\lambda)$ and $T(\lambda)$ are CT transfer functions¹.

2.2 Well-Posedness and Internal Stability

The meaningful application of feedback to reduce an uncertainty (in this case, error introduced by the nonlinear quantizer) requires that the system be well-posed in order for a solution to exist. Figure 2.1 can undergo block diagram manipulation bringing it into the standard feedback form shown in Figure 2.2 with signals r , e , d , and y .

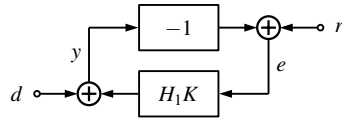


Figure 2.2: The linearized model converted into standard feedback form.

¹Note that regarding Figure 2.1 and Figures 1.3/1.4, the NTF $S(\lambda)$ is the same transfer function whether interpreted from $d \rightarrow y$ or $r \rightarrow e$.

The equations describing this loop are:

$$\begin{bmatrix} r \\ d \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ -H_1 K & 1 \end{bmatrix} \begin{bmatrix} e \\ y \end{bmatrix}. \quad (2.1)$$

A feedback system is considered well-posed if the inverse of the transfer matrix in Equation 2.1 exists and each of its elements are proper. Equation 2.2 shows that this is the case if both $S(\lambda)$ and $T(\lambda)$ are proper transfer functions.

$$\begin{bmatrix} e \\ y \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ -H_1 K & 1 \end{bmatrix}^{-1} = \begin{bmatrix} \frac{1}{1+H_1 K} & \frac{-1}{1+H_1 K} \\ \frac{H_1 K}{1+H_1 K} & \frac{1}{1+H_1 K} \end{bmatrix} \begin{bmatrix} r \\ d \end{bmatrix} = \begin{bmatrix} S & -S \\ T & S \end{bmatrix} \begin{bmatrix} r \\ d \end{bmatrix}. \quad (2.2)$$

The principle of internal stability is stricter than bounded-input bounded-output (BIBO) stability because it guarantees that the internal states of the system remain bounded. The system in Equation 2.2 is internally stable if each element of the transfer matrix belongs to the set \mathcal{RH}_∞ , that is the set of stable real rational proper transfer functions.

2.2.1 Constraints on the Noise Transfer Function

A sufficient condition for $S(\lambda)$ and $T(\lambda)$ to be proper is that transfer function $H_1(\lambda)$ is a strictly proper real rational transfer function. Internal stability of the system follows if $S(\lambda)$ and $T(\lambda)$ are stable. This leads to the following constraints on the NTF:

1. $S(\lambda)$ is stable, and ,
2. The following equivalent conditions hold:
 - (a) $S(\infty) = 1$,
 - (b) If $S(\lambda)$ is in state-space form, the feedthrough matrix $D = 1$, and,
 - (c) The first element of the impulse response of $S(\lambda)$ is one.

Most prior work in the area performs optimization directly on the NTF of the system. This is effective because it is a relatively accurate model of the noise

shaping performance. In addition, constraint 2 enforces causality on the feedback loop ensuring the system is physically realizable.

2.3 Modelling Uncertain Quantizer Gain

Having established conditions to ensure the closed-loop system is realizable and internally stable, there remains a nonlinear gain block K . K can be understood as a time-varying gain dependent on the quantizer input. For example, a 1-bit quantizer ($\Delta_Q = 2$) with output $\{-1, 1\}$ would have instantaneous gain $K(t) = \frac{1}{u(t)}$. As the value of u at each sample time is not known in advance, K may be modelled as a multiplicative uncertainty. The upper linear fractional transformation (LFT) allows K to be separated into a constant gain matrix $M_{2 \times 2}$ and a normalized, \mathcal{H}_∞ norm-bounded uncertain block Δ by Expression 2.3.

$$K \leftrightarrow \mathcal{F}_U\{M, \Delta\} \quad \|\Delta\|_\infty \leq 1 \quad (2.3)$$

The model from Figure 2.2 is shown in Figure 2.3 with the quantizer and variable gain replaced by this LFT interconnection. In Chapter 4, it is of interest to ensure the robustness of the system to Δ , which may be achieved using this form.

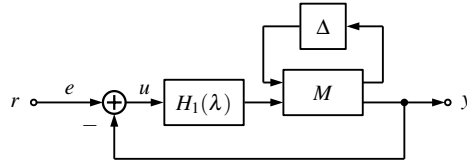


Figure 2.3: The linearized block diagram with the quantizer replaced by a multiplicative uncertainty extracted via LFT.

2.4 Derivation of Augmented System

2.4.1 Extraction of Performance and Stability Channels

Finally, the model is abstracted into an augmented form where all desired input and output channels are present and all unnecessary ones hidden. Let the LFT input to M and output from M be w and z , respectively. These channels are required to

Table 2.1: Input and output channels of interest for the augmented system.

Output Input	z	e	u	y
r	Not used	NTF performance channel	Constraint on quantizer input signal	STF constraints for CT design
w	Quantizer gain robustness channel	Not used	Not used	Not used

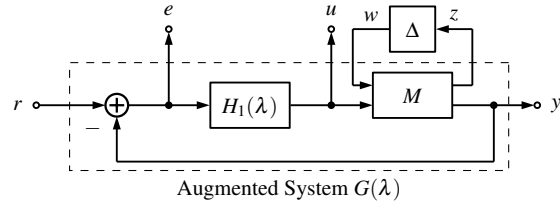


Figure 2.4: The augmented plant is derived by setting $H_0(\lambda) = 1$, taking the LFT of the uncertain gain, extracting the signals of interest, and writing the closed-loop equations.

be accessed in addition to r , e , u , and y for the purposes listed in Table 2.1. The augmented system $G(\lambda)$ is shown as the dashed block in Figure 2.4.

2.4.2 Derivation of State-Space Model

Now that the desired input and output signals are captured by the model, it is a simple exercise to write the system in state-space form. To begin, let filter $H_1(\lambda)$ be the transfer function of order n in variable $\lambda = z$ in the DT case (or $\lambda = s$ in the CT case). The numerator and denominator coefficients are shown in Equation 2.4 which has the equivalent state-space representation of Equation 2.5.

$$H_1(\lambda) = \frac{b_{n-1}\lambda^{n-1} + b_{n-2}\lambda^{n-2} + \dots + b_1\lambda + b_0}{\lambda^n + a_{n-1}\lambda^{n-1} + a_{n-2}\lambda^{n-2} + \dots + a_1\lambda + a_0} \quad (2.4)$$

$$= C_H(\lambda I - A_H)^{-1} B_H \quad (2.5)$$

Naturally, $H_1(\lambda)$ is a strictly proper transfer function and state-space feedthrough matrix $D_H = 0$ due to the constraints proposed in Section 2.2. The constant gain matrix M may be split into its constituent parts as shown in Equation 2.6.

$$M = \begin{bmatrix} m_{11} & m_{12} \\ m_{21} & m_{22} \end{bmatrix} \quad (2.6)$$

With some algebra, the augmented system $G(\lambda)$ from Figure 2.4 may be written in state-space form with notation from Equations 2.5 and 2.6 by introducing state vector x . We use the notation $G_{qp}(\lambda)$ to indicate the transfer function of $G(\lambda)$ from some input channel p to some output channel q and name the closed-loop state-space matrix blocks with cursive symbols as shown in Equation 2.8.

$$G : \begin{bmatrix} \dot{x} \\ z \\ e \\ u \\ y \end{bmatrix} = \left[\begin{array}{c|ccc} A_H - m_{22}B_H C_H & -m_{21}B_H & B_H & \\ \hline m_{12}C_H & m_{11} & 0 & \\ -m_{22}C_H & -m_{21} & 1 & \\ C_H & 0 & 0 & \\ m_{22}C_H & m_{21} & 0 & \end{array} \right] \begin{bmatrix} x \\ w \\ r \end{bmatrix} \quad (2.7)$$

$$= \left[\begin{array}{c|cc} \mathcal{A} & \mathcal{B}_w & \mathcal{B}_r \\ \hline \mathcal{C}_z & \mathcal{D}_{zw} & \mathcal{D}_{zr} \\ \mathcal{C}_e & \mathcal{D}_{ew} & \mathcal{D}_{er} \\ \mathcal{C}_u & \mathcal{D}_{uw} & \mathcal{D}_{ur} \\ \mathcal{C}_y & \mathcal{D}_{yw} & \mathcal{D}_{yr} \end{array} \right] \begin{bmatrix} x \\ w \\ r \end{bmatrix} \quad (2.8)$$

With the channels of interest exposed and the system in a state-space form, one can express design goals as constraints on these channels. In Chapter 3, various stability measures and performance goals are discussed from which those that are ideal from an optimization perspective are selected. In Chapter 4, the framework is introduced to allow the targets from the previous chapter to be applied to the augmented system in a way that allows the optimization problem to be efficiently solved.

Chapter 3

Stability Criteria and Performance Goals

Due to the nonlinear effects of the quantizer, the stability of the sigma delta feedback loop is difficult to prove. An excellent exploration into the mechanisms of instability may be found in [6]. There are no known necessary conditions for stability of sigma delta modulators but there are several heuristic and sufficient conditions with various degrees of conservativeness. In Section 3.1.1, some theory from relay feedback control is introduced to establish formal methods for ensuring stability. There are some shortcomings of these methods when applied to practical sigma delta modulator design, therefore Sections 3.1.2 to 3.1.5 describe some stability criteria that may be less robust but allow a greater performance-stability tradeoff. For completeness, additional stability methods of interest that are not compatible with this optimization framework are presented in Section 3.2. Finally, the performance goal is discussed in Section 3.3.

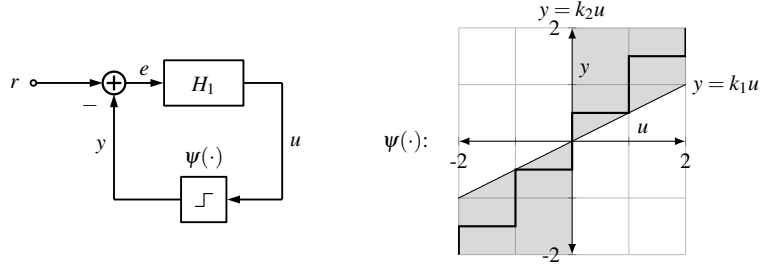


Figure 3.1: A Lur'e system (left) with the example nonlinear transfer curve of an infinite quantizer ($\Delta = 1$) shown with a shaded sector bounded region (right).

3.1 Stability Criteria Used by this Optimization Framework

3.1.1 Ideas from Nonlinear Control

An early theoretical treatment of nonlinear control is the circle criterion, which provides a graphical frequency domain method for evaluating the stability of a CT Lur'e system. A Lur'e system is a simplified negative feedback loop consisting of a linear plant $L(s)$ with a nonlinear element $\psi(\cdot)$ in the feedback path such as the one shown in Figure 3.1. The transfer curve of the nonlinear element may be time-varying and even non-monotonic but is bounded by a sector condition, a set of two lines passing through the origin with slopes k_1, k_2 that bound the curve in each direction.

Theorem 3.1.1 (Circle criterion [21, Sec. 7.1.1]). *Given the Lur'e feedback system in Figure 3.1 where the denominator of $H_1(s)$ is Hurwitz and $\psi(t, \cdot)$ is a memory-less function sector bounded by $[k_1, k_2]$, the closed-loop system $L(s)$ is globally asymptotically stable if one of the following cases is true:*

1. Case $\psi \in [k_1, \infty)$: The inequality in Equation 3.1 is satisfied.

$$\Re \left\{ \frac{H_1(s)}{1 + k_1 H_1(s)} \right\} > 0 \quad (3.1)$$

2. Case $\psi \in [k_1, k_2]$, $k_2 - k_1 > 0$: The inequality in Equation 3.2 satisfied.

$$\Re \left\{ \frac{1 + k_2 H_1(s)}{1 + k_1 H_1(s)} \right\} > 0 \quad (3.2)$$

The graphical interpretation for the circle criterion¹ is that the Nyquist plot of $H_1(s)$ does not enter the disk passing through the points $-\frac{1}{k_1} + j0$ and $-\frac{1}{k_2} + j0$ if $0 < k_1 < k_2$. When $0 = k_1 < k_2$, the Nyquist plot must lie to the right of the vertical line $\Re\{s\} = -\frac{1}{k_2}$. If a single-bit quantizer is used as is the scope of this thesis, the sector bounds include the entire first and third quadrants. Case 1 from Theorem 3.1.1 then applies and the Nyquist plot of $H_1(s)$ must lie entirely in the right half-plane. The optimization framework presented in Chapter 4 may be used with the circle criterion although with a single-bit quantizer, the method is too restrictive for practical use.

The Popov criterion in Theorem 3.1.2 is a slightly less conservative approach that restricts the problem to time-invariant nonlinearities.

Theorem 3.1.2 (Popov criterion [21, Sec. 7.1.2]). *Given the Lur'e feedback system in Figure 3.1 where the denominator of $H_1(s)$ is Hurwitz and $\psi(\cdot)$ is a time-invariant memoryless function sector bounded by $[0, k_2]$, the closed-loop system $L(s)$ is globally asymptotically stable if there exists a scalar $\gamma \geq 0$ such that the following inequality is satisfied:*

$$\frac{1}{k_2} + \Re\{H_1(j\omega)\} - \gamma\omega\Im\{H_1(j\omega)\} > 0 \quad \forall \omega \in [0, \infty). \quad (3.3)$$

The graphical interpretation for this criterion is that the Popov plot of $\omega\Im\{H_1(j\omega)\}$ versus $\Re\{H_1(j\omega)\}$ remains to the right of a line passing through point $-\frac{1}{k_2} + j0$ with slope $\frac{1}{\gamma}$.

The DT version is the Tsytkin criterion, which has cases valid for time-varying and time-invariant nonlinearities. The analog to the circle criterion is shown in Theorem 3.1.3 and the analog to the Popv criterion is shown in Theroem 3.1.4.

¹The circle criterion has different graphical interpretations for the cases where $k_1 < 0$ and where $H_1(s)$ has zeros in the open right half-plane, but these are omitted because they are not valid quantizer transfer curves or because nonminimum phase $H_1(s)$ are not considered here (see [21, Sec. 7.1.1] for more details).

Theorem 3.1.3 (Tskypin criterion for time-varying nonlinearities [22, Sec. 4.6]). *Given the Lur'e feedback system in Figure 3.1 where the denominator of $H_1(z)$ is Schur and $\psi(t, \cdot)$ is a memoryless function sector bounded by $[0, k_2]$, the closed-loop system $L(z)$ is globally asymptotically stable if the following inequality is satisfied:*

$$\frac{1}{k_2} + \Re\{H_1(z)\} \geq 0 \quad \forall |z| = 1. \quad (3.4)$$

Theorem 3.1.4 (Tskypin criterion for time-invariant nonlinearities [22, Sec. 4.7]). *Given the Lur'e feedback system in Figure 3.1 where the denominator of $H_1(z)$ is Schur and $\psi(\cdot)$ is a time-invariant memoryless function sector bounded by $[0, k_2]$, the closed-loop system $L(z)$ is globally asymptotically stable if there exists a scalar $\gamma \geq 0$ such that the following inequality is satisfied:*

$$\frac{1}{k_2} + \Re\{(1 + \gamma(1 - z^{-1}))H_1(z)\} \geq 0 \quad \forall |z| = 1. \quad (3.5)$$

The Jury-Lee criteria are less strict cases of the Tsyppkin criteria requiring that the nonlinearity be slope bounded and monotonic. However, this is not applicable to quantizer feedback, where the slope may go to infinity. The above techniques from nonlinear control are sufficient conditions and are related to important results from passivity theorem.

3.1.2 \mathcal{H}_∞ Stability Criterion

The \mathcal{H}_∞ stability criterion, commonly known as Lee's rule, is a heuristic predictor of stability which states that a modulator is likely to be stable if the NTF out-of-band gain, or $\|S(\lambda)\|_\infty$, does not exceed a benchmark value. The rule was initially based on the empirical study of a fourth-order DT sigma delta modulator with single-bit quantization [23]. The criterion is not necessary nor sufficient for stability and must be verified with extensive simulations. Despite this, the rationale for its use as a suggestion of stability comes from the Bode sensitivity integral shown in Equation 3.6 for Schur stable $H_1(z)$ [24, Thm. 1].

$$\frac{1}{2\pi} \int_{-\pi}^{\pi} \log |\det(S(\omega))| d\omega = 0 \quad (3.6)$$

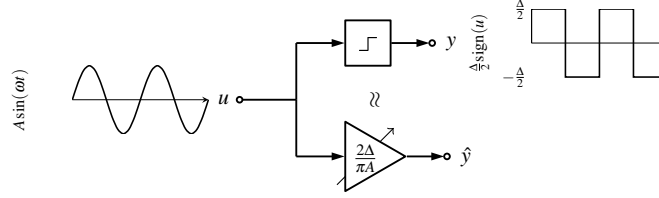


Figure 3.2: An ideal 1-bit quantizer (above) and its describing function approximation (below).

The integral enforces that the total area under the curve of the NTF log-magnitude versus frequency is equal to zero. Applied to the sigma delta linear model, if the sensitivity of the closed-loop system to the quantization error is suppressed in the signal band, it must be compensated for by an equal area of amplified sensitivity outside the signal band. Because the quantization error is nonlinear and signal dependent, the higher the gain of the sensitivity function, the greater chance there is for a limit cycle at that frequency to destabilize the loop. Thus, the Lee's rule is a indicator of the performance-stability tradeoff. In practice, $\|S(z)\|_\infty \leq 2$ is often used, but this has been found to be conservative for low-order and inadequate for high-order designs [25]. However, the criterion is used extensively as a starting point for practical design due to its inclusion in popular software tools. It is easy to formulate as part of an optimization problem as it may be applied with an \mathcal{H}_∞ constraint on the $r \rightarrow e$ channel of Equation 2.7.

3.1.3 Describing Function Approximation and Root Locus Stability

Two closely related stability methods are the describing function approximation and the root locus approach. These both rely on the variable gain K introduced in Section 2.1 but use different interpretations of it to stabilize the sigma delta modulator.

The describing function method [26] is an approximate technique of linearization often applied to steady-state electrical circuits or nonidealities in mechanical systems. As shown in Figure 3.2, a zero-mean sinusoidal input to the quantizer is assumed: $u(t) = A \sin(\omega t)$. The Fourier series of the output is truncated at the first odd coefficient because the quantizer transfer curve is also an odd function. For a

single-bit quantizer, the first coefficients are:

$$a_1 = \frac{1}{\pi} \int_0^{2\pi} \psi(t) \cos(\omega t) d(\omega t) \quad (3.7)$$

$$b_1 = \frac{1}{\pi} \int_0^{2\pi} \psi(t) \sin(\omega t) d(\omega t), \quad (3.8)$$

Where $\psi(t)$ is the nonlinear quantizer function. Integral 3.7 evaluates to zero while the period of Integral 3.8 may be split into two parts:

$$b_1 = \frac{1}{\pi} \left(\int_0^{\pi} \psi(t) \sin(\omega t) d(\omega t) + \int_{\pi}^{2\pi} \psi(t) \sin(\omega t) d(\omega t) \right).$$

In the interval $\omega t \in (0, \pi)$, the single-bit quantizer outputs $\frac{\Delta}{2}$ whereas in the interval $\omega t \in (\pi, 2\pi)$, the single-bit quantizer outputs $-\frac{\Delta}{2}$. By symmetry, this integral is equal to Equation 3.9.

$$b_1 = \frac{\Delta}{\pi} \int_0^{\pi} \sin(\omega t) d(\omega t) = \frac{2\Delta}{\pi} \quad (3.9)$$

Using the Fourier series approximation $\hat{y}(t) \approx y(t)$, the describing function is derived as follows:

$$\begin{aligned} N(A) &= \frac{\hat{y}(t)}{u(t)} \\ &= \frac{b_1 \sin(\omega t)}{A \sin(\omega t)} \\ &= \frac{2\Delta}{\pi A}. \end{aligned}$$

Thus, the describing function of the single-bit quantizer is a variable gain $N(A)$ dependent on the quantizer input amplitude. As expected, when the input amplitude approaches zero, the gain approaches infinity and when it approaches the quantizer limit, the gain approaches one. In fact, instability in sigma delta modulators is often associated with low frequency, large amplitude limit cycles where the quantizer gain is low. The describing function method is a good approximation to large signal stability but often fails to predict small limit cycles because the higher

harmonics of the output are neglected. The describing function has been applied to the design of sigma delta modulators [27] and extended to be dependent on phase in addition to gain to design a sixth-order modulator [28].

In the open loop the describing function method approximated the quantizer as a variable gain. The root locus can determine stability by showing the position of the closed-loop poles as a function of this gain. One method to design stable sigma delta modulators is to position the poles and zeros of the loop filter such that the root locus remains in the stable region of the complex plane when sweeping through valid quantizer gain values [29–31]. Recall the LFT used to model the varying gain in Section 2.3. Equation 3.10 defines M when the gain is within a given range $K \in [k_l, k_h]$ with nominal value k_0 .

$$M = \begin{bmatrix} \frac{k_h - 2k_0 + k_l}{k_h - k_l} & \frac{-2(k_0 - k_h)(k_0 - k_l)}{k_h - k_l} \\ 1 & k_0 \end{bmatrix} \quad (3.10)$$

The root locus stability criterion may be used in robust control fashion by choosing a range for K , e.g. $[k_l, k_h] = [1/||u||_\infty, \infty]$, then constraining the \mathcal{H}_∞ norm to unity for the $z \rightarrow w$ channel. This ensures that the linearized model is stable for the selected gain values.

3.1.4 \mathcal{H}_2 Stability Criterion

Previously, the quantizer was replaced by an additive noise source and some performance estimations were presented assuming that the quantization noise was uncorrelated with input had a white spectrum. The white noise model is only a close approximation if the following hold [32, Ch. 6]:

1. The quantizer is not overloaded,
2. There are a large number of quantization levels with small Δ_Q , and
3. The probability density function (PDF) of input samples is smooth.

In reality, especially with single-bit quantization, the approximation does not hold. The \mathcal{H}_2 stability criterion, sometimes called the power gain rule, uses a statistical look at the quantizer input [6]. The output $y[k]$ of a single-bit quantizer may

be considered the superposition of three signals: a DC component μ_y , AC component amplified by the quantizer gain $K(u[k] - \mu_u)$, and the quantization noise $d[k]$. With these additional degrees of freedom, one can enforce that $d[k]$ is white and uncorrelated with the quantizer input $u[k]$ by setting K to that in Equation 3.11.

$$K = \frac{\text{cov}\{u[k], y[k]\}}{\sigma_u^2} \quad (3.11)$$

The gain K is now entirely dependent on μ_y and the distribution of u .

3.1.5 ℓ_1 Stability Criterion

3.2 Stability Concepts Not Used by this Optimization Framework

3.2.1 Methods Ensuring Bounded States

A different way of ensuring stability of a modulator system is the positive invariant set ap

3.2.2 Diagonal Modulators

3.3 Performance Goals

Chapter 4

Optimization of Loop Filter Design

Chapter 5

Design Examples

Chapter 6

Conclusions

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Appendix A

Supporting Materials

This would be any supporting material not central to the dissertation. For example:

- additional details of methodology and/or data;
- diagrams of specialized equipment developed.;
- copies of questionnaires and survey instruments.