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A TUTORIAL GUIDE TO NOISE SHAPING AND OVERSAMPLING IN ADC AND DAC SYSTEMS

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This tutorial paper reviews the principles of oversampling and noise shaping with specific applications to digital audio. Both ADC and DAC systems are considered where in the limit, the 1-bit converter is presented. The paper describes a consolidation of ideas and techniques and attempts to demonstrate the instrumentation advantages to be gained.

(0) Introduction

It is well known that all naturally occurring signals are bounded in terms of bandwidth and signal-to-noise ratio (SNR). In audio, the bound is set by the frequency response of the ear (approximately 20 kHz) and the sensitivity of the ear to the quietest signals which in the limit is close to the noise contribution set by the Brownian motion of air molecules striking the ear drum. Once a signal is so bounded, it can in theory be represented in discrete form by using a combination of Nyquist sampling and amplitude quantisation with the minimum of degradation. In present day digital systems this process translates into a discrete signal with a sampling rate of 44.1 kHz and an amplitude resolution of 16 bit using a uniform quantisation law.

Once a signal is represented in discrete form Information Theory (1) tells us that this format is not unique where there exists the opportunity to explore an exchange between sampling rate and sample resolution without incurring significant corruption of the original message signal. There is an undeniable logic to this proposition where if the sampling rate is increased from the original Nyquist rate f_nHz to a new rate f_sHz (where $f_s > f_n$), then because of the extra samples and therefore lower energy contribution from each sample, the message signal will be more tolerant to sample error enabling coarser quantisation to be used. As a simple example consider a system where $f_s = 3f_n$, that is an oversampling ratio L = 3, where

$$L = \frac{f_s}{f_n} \qquad \ldots \quad 1$$

and where each sample takes the levels 0,1,2,3. A single sample is then formed which is equal to the sum of the past three samples where the distribution of levels is represented in the following tree:

NOISE SHAPING AND OVERSAMPLING

9
888
777777
6666666666
555555555555
444444444444
3333333333
222222
111
0

The resulting samples are shown to have an amplitude range spanning 0 to 9 which is greater than the individual input samples. However, it is also evident that other than levels 0 and 9, there are no unique pulse sequences that will generate the intermediate levels 1 to 8 suggesting further latitude for a more efficient code conversion that can exploit these other pulse combinations. The method by which sequences of oversampled samples are selected to match a message signal or data sequence is through the use of a *noise shaping* algorithm(2).

On first encounter, it may not appear necessary to consider systems exploiting oversampling as conversion between the analogue and digital domains (and vice versa) can be achieved using appropriate analogue filters and direct forms of ADC and DAC performed at the Nyquist rate where theoretically perfect conversion can be achieved. However, the advantage of oversampling and associated noise shaping is realised more from the perspective of instrumentation in association with ADC and DAC performance and the greater ease with which the theoretical limits of digital audio can be approached.

In this paper, the basic principles of noise shaping and oversampling are discussed and their advantages highlighted. The aim is to introduce these concepts although since space prevents a rigorous discussion of all aspects, some reference material is cited for further reading.

(1) The case for oversampling and noise shaping

In this Section, the advantages of oversampling and noise shaping are reviewed with particular reference to the target performance of a digital channel and the instrumentation required to realise that target.

(1.1) Target performance of a digital audio channel

The parameters of a digital channel are bounded by the sampling rate and the (here uniform) amplitude resolution of each sample, where by Nyquist's theorem the allowable message bandwidth must be less than $f_n/2$Hz and the SNR integrated over the band 0 to $f_n/2$Hz is,

$$SNR = 6.02N + 1.76 \text{ dB} \qquad \ldots 2$$

where N is the number of bits per sample (assuming uniform amplitude quantisation).

NOISE SHAPING AND OVERSAMPLING

Although amplitude quantisation is a non-linear process, when used in association with a perfect quantiser and optimal dither, the quantisation distortion can be shown to take on the characteristic of random noise. In other words, the channel behaves as if it is bandlimited and has an additive noise source, whereby low-level signals appear to enter the noise without the sonic character of non-linear distortion. For this ideal to be achieved then the quantiser must offer a uniform transfer characteristic, with equally spaced quanta, any deviation from this ideal will result in low-level distortion even if a dither source is employed. In effect, we require the accuracy of the ADC-DAC combination to exceed by several bits, the resolution of the system, thus for a 16 bit system, the accuracy of the converters should approach 18 to 20 bit if exemplary low-level coding is to be achieved. Alternatively, if such accuracy cannot be achieved overall, then the non-linear residue of the converter system should appear as a benign noise component rather than non-linear distortion. It is in this area of low-level resolution, possibly in the presence of high-level signal components, where many ADC-DAC systems fail. However, it is emphasised that the residual distortions are a function of poor instrumentation rather than any fundamental failing of digital encoding theory.

(1.2) The Oversampling Advantage

Oversampling offers several advantages and can be applied to both ADC and DAC:

(a) By spreading the signal over a greater number of samples, each sample carries less weight, thus the system is desensitised to the effects of sample jitter and sample area error.

(b) Quantisation distortion or, in the case of a DAC, re-quantisation distortion can be shown to be spread over a band 0 to $f_s/2$ Hz, hence if signal bandwidth is $< f_s/2$ Hz then only a fraction of the quantisation distortion now corrupts the message.

(c) By choosing a value of $L > 1$, the designs of anti-aliasing and reconstruction filters are relaxed which enables simpler analogue electronics with less precise frequency response requirements. However, when sampling rate conversion (either upwards or downwards) is performed, a digital low-pass filter is used which augments the analogue filters. In other words, signal filtering can be performed substantially in the digital domain where a far more precise transfer function can be realised.

(1.3) The noise shaping advantage

Once a signal is represented in an oversampled format, the message bandwidth by definition is less than $f_s/2$ Hz, consequently there can be a large fractional signal space in which no useful signal components reside. Also simple reasoning suggests that once a signal is oversampled, that if each sample remains specified to the same amplitude resolution, then there is redundancy. The objective of noise shaping is to reduce the resolution of each sample and to attempt to locate the extra quantisation noise in the redundant frequency space created by oversampling, while inflicting the minimum of corruption to the baseband message signal.

The usual method by which noise shaping is achieved is to enclose the re-quantiser within a recursive loop and to use the distortion shaping characteristic of negative feedback to locate the majority of the additional distortion at high frequency (compared with the message baseband). Such a process can be performed either in the analogue domain (forming an ADC) or in the digital domain (forming a DAC).

It therefore follows that an ADC or DAC can be used with a lower amplitude resolution than a direct encoder providing the converter can operate satisfactorily in the oversampling mode. Effectively a noise shaping procedure enables an exchange to be made between sampling rate and amplitude resolution, hence instead of using an 18 bit converter at 44.1 kHz a 6 bit converter at 11 MHz (for example) could be used. Often it is easier to design a low-resolution converter that operates at high speed than a high resolution converter at lower speed, particularly where linearity with signal level is important and coding of low-level signals is required with a minimum of low-level corruption.

A bonus of noise shaping, particularly where high-order shaping filters are used (see later discussion) is that the coder is self-dithered and exhibits an element of chaotic behaviour with the beneficial characteristic of converting non-linear quantiser errors to noise(3).

In the limit, providing high oversampling and high-order noise shaping are used, the quantiser can be reduced to 1-bit (the so-called delta-sigma modulator (4)). Since there are now only two-levels in the output code, the system can be shown to be tolerant of hardware non-ideality thus exhibiting good linearity with signal level within the bound of quantisation distortion. The audio industry is now turning to 1-bit systems, though it is interesting to observe that 1-bit converters preceded most of the present day converter schemes. Of course, in using high oversampling schemes there is a penalty due to the extra digital processing required for either interpolation (upward conversion of sampling rate) or decimation (downward conversion of sampling rate(5)), though with VLSI technology, this complication can be economically accommodated.

(2) Noise shaped and oversampled ADC

In this Section the basic operation of a high-order ADC is described to illustrate the recursive technique of noise shaping(6). The process has much in common with the distortion reduction properties of negative feedback where in a basic comparison the output stage distortion of an amplifier can be compared with the quantiser of an ADC. This comparison is shown in Fig 1 where A represents an analogue forward path transfer function (common to both structures) B is the amplifier output stage, and Q and T respective amplitude and time quantisation. A primary distinction between the two structures is that the analogue amplifier operates continuously while the quantiser only responds in discrete time, however since A is a continuous process and the sampling rate f_sHz is high compared with the message band, a reasonable comparison can be made.

To facilitate analysis, the distortion introduced by either B or Q and T is represented by the additive component q as shown in Fig 1c. Although additive distortion modelling is an approximation, computer simulation of noise shaping encoders reveals reasonable performance prediction, therefore this method can be used to describe the behaviour of a noise shaping coder. Analysis of the Fig 1c model gives,

$$V_0 = V_{in}\frac{A}{1+A} + \frac{q}{1+A} \quad . \; . \; . \quad 3$$

NOISE SHAPING AND OVERSAMPLING

For the message signal A >> 1 therefore equation 2.1 can be rewritten as,

$$V_0 = V_{in} + D_N q \qquad . \ . \ . \ 4$$

where the system exhibits unity gain together with a noise shaping function D_N,

$$D_N = \frac{1}{1 + A} \qquad . \ . \ . \ 5$$

Equation 5 shows that the noise shaping function D_N is (almost) inversely proportional to the processor gain A, hence the larger A, the greater is the reduction of noise. However, there is also the requirement that the closed loop be stable and exhibit non-divergent oscillation, this is further compounded in the sampled data model of Fig 1b since the half-sample frequency $f_s/2$Hz can be thought of as infinite frequency compared with the continuous-time model of Fig 1a. Hence, for a given message bandwidth f_mHz and oversampling frequency f_sHz there is only a finite bandwidth in which to fit and shape the quantisation noise q, this together with the stability requirement implies A should be maximised in the message band yet fall below unity at $f_s/2$Hz. Functions for A must therefore be selected which offer a high rate of attenuation with increasing frequency yet also meet the stability bound. In choosing A consideration must also be given to any time-dependant behaviour of the quantiser together with the compounded effect of non-linearity. Providing the quantiser response is fast compared with the sample period the linear constraints are not too problematic, however the effects of non-linearity particularly if the quantiser saturates by offering only a finite number of levels (possibly as little as 2 for the DSM) can require a compromise in the transfer function of A.

A typical s-domain description for A consists of N cascaded integrators together with (N-1) transmission zeros to achieve stability ie

$$A = \frac{(1 + ST_1)^{(N-1)}}{(ST_2)^N} \qquad . \ . \ . \ 6$$

whereby,

$$D_N = \frac{(ST_2)^N}{(ST_2)^N + (1 + ST_1)^{N-1}} \qquad . \ . \ . \ 7$$

$$D_N \Big|_{f << f_s/2} \sim (ST_2)^N \qquad . \ . \ . \ 8$$

that is the distortion shaping function appears as N (the order of the noise shaper) cascaded differentiators. The advantage in choosing $N > 1$ is therefore apparent as the greater slope in D_N yields a smaller value in the message band.

This basic example shows how by choosing an appropriate sampling rate, noise shaper order and quantiser amplitude resolution a suppression of quantiser noise in the message band can be achieved possibly meeting the performance of a digital audio channel. The relative simplicity of the loop should be noted and because $f_s >> f_n$, the need for significant input antialiasing filtering is avoided.

Although the noise shaped ADC acts as a means of digitising an analogue signal, the signal format is not compatible with conventional digital audio channels, hence the process of decimation must be implemented. Decimation is effectively a technique of low-pass filtering which then enables sub-sampling without incurring in-message band aliasing distortion. Although decimation can be performed in a single stage, it is more efficient in terms of multiplications, to perform the conversion in at least 2 or 3 stages though the final result is similar. To illustrate the process, an oversampled "front-end" ADC is shown in Fig 2 together with a single-stage decimator, where a frequency domain description also reveals how sub-sampling can be performed without incurring aliasing distortion. It is important to note that in computing the output Nyquist samples (at a rate f_nHz) that each sample is an average taken over successive input samples, thus reducing effects of sample jitter, and that the digital decimation filter mimics the function of an analogue anti-aliasing filter but offering the advantages of digital processing:

* symmetrical, linear-phase FIR filter response
* very high rates of attenuation
* negligible in-band amplitude ripple
* high out-of-band attenuation
* wide signal dynamic range
* repeatability
* no temperature drift or environmental dependence
* small size
* no dependence on analogue component parasitics and non-idealities
* no ground rail or circuit interaction problems

NOISE SHAPING AND OVERSAMPLING

(3) Noise shaped and oversampled DAC

The techniques described in Section 2 can also be used in digital to analogue conversion (7,8). However, in this case the Nyquist samples must first be converted to a higher sampling rate using a process of interpolation, whereby noise shaping can then be used to reduce the sample amplitude resolution, hence removing redundancy, by relocating the re-quantisation noise into the oversampled signal space. The method enables a fast, but low-resolution DAC to be used with only a minimum of analogue electronics. In essence, the process of continuous signal reconstruction is now performed partially in the digital domain, where a digital process (interpolation) initially produces a finer discrete resolution along the time axis thus allowing a simple, low-order analogue filter to complete the conversion to a continuous function.

The system of DAC using oversampling and noise shaping is illustrated in Fig 3 where oversampling is achieved using interpolation. As with decimation, interpolation uses digital low-pass filter techniques together with a technique of zero pulse insertion to increase the sampling rate, where it can again be shown that using a multi-stage structure, less multiplications are required. Effectively oversampling is achieved by inserting zero samples between Nyquist samples so as to achieve a high sampling rate, then a low-pass filter removes the intermediate spectral replications thus achieving sample interpolation.

The noise shaper is similar to the ADC of Fig 1b except A is composed of a cascade of digital integrators (see Fig 3c) together with a series of feedforward paths to achieve loop stability. For the system of Fig 3c, each integrator A_r has the z-domain transfer function

$$A_r(z) = \frac{1}{z-1} \quad . \; . \; . \quad 9$$

which has the corresponding frequency response for $f << f_s/2$ Hz of,

$$A_r(f) = \frac{f_s}{j2\pi f} \quad . \; . \; . \quad 10$$

Analysis of the Fig 3c structure then reveals a distortion shaping function $D_N(f)$, expressed in the frequency domain,

$$\left| D_N(f) \right| = \left[2 \sin\left(\frac{\pi f}{f_s} \right) \right]^N \quad . \; . \; . \quad 11$$

In Fig 4, example plots of $D_N(f)$ are shown for $N = 1$ to 6 which illustrate the improving noise shaping advantage at low frequency as N is increased. However, examination of Fig 4 also reveals that

NOISE SHAPING AND OVERSAMPLING

(i) at $f = f_s/6\text{Hz}$, $\left|D_N(f)\right| = 1$ . . . 12

(ii) at $f = f_s/2\text{Hz}$, $\left|D_N(f)\right| = 2^N$ . . . 13

ie the majority of the re-quantisation noise is located within the frequency space $f_s/6 < f < f_s2$.

Equation 13 is an interesting result which shows an actual gain in the level of distortion surrounding $f_s/2$Hz. If we assume a re-quantisation step of Δ, that is a quantiser distortion range $\pm\Delta/2$, then the level of output signal from the noise shaper will span a range approximately

$$0.5\left|D_N(f_s/2)\right|\Delta \quad \text{ie} \quad 2^{N-1}\Delta$$

Since we are dealing with a negative feedback loop with a substantial non-linearity then performance prediction and verification is best performed using computer simulation. Indeed simulation does reveal signal activity as predicted where the combination of multi-stage signal integration and amplitude quantisation within a recursive loop produces a non-divergent oscillatory component showing chaotic activity. This chaotic behaviour acts as a form of self dither that is particularly effective for $N \geq 3$, where most of the DAC levels within a range $\pm 2^{N-1}\Delta$ participate in the conversion process irrespective of the signal amplitude (at least for signals $< 1.5\Delta$).

Taking account of the noise shaping characteristics shown in Fig 4 then for N = 4 (say) and an oversampling ratio circa 256, together with a 16 level (4 bit) DAC, it can be shown (8) that a performance exceeding the need of the present day digital audio format is achievable where computer simulation also reveals the desirable property of converting DAC distortion to a noise-like residue.

(4) 1-bit converters

So far we have demonstrated the means by which an interchange between sampling rate and sample resolution can be made on the basis of a constant information bandwidth and SNR. In the limit the individual sample resolution can be reduced to a two-level format whereby the DAC becomes a binary logic gate. The basic noise-shaping algorithm that can achieve this reduction is similar to that of Fig 1c and Fig 3c though the amplitude quantiser is reduced to 2 levels. Consequently the now significant distortion produced by 2-level quantisation is once more frequency shaped by the action of the negative feedback loop and concentrated into the redundant frequency space created by oversampling. The effect of this process is to generate a sequence of one-zero pulses that when averaged yield a close approximation to the information signal.

To demonstrate this code generation, Fig 5 illustrates a binary pulse sequence for a sinusoidal input with noise shapers of N = 1 and N = 2, where the method of signal recovery by bandlimitation of the sequence should be evident. The illustration shows that for low-level signals and N = 1, the output code reduces to ..0101.., while for N = 2, there is less regularity. As might be anticipated, for N = 2 the system is capable of more accurate encryption particularly of low-level signals.

On first consideration it may be assumed that by increasing the loop order that greater coding accuracy is achieved. However, the presence of a two-level quantiser represents a gross non-linearity and also, for a given sampling rate, a limit on information transmission. The effect therefore of increasing loop order is an entry into a non-linear regime where poor or in practice no useful coding is achieved. The usual limit of N is normally 2 although, by careful choice of loop filter recent research has demonstrated the opportunity for stable performance with orders as high as 4 (9).

The application of binary DAC systems is a relatively new technique in digital audio application that is being researched by Philips (10). However, the techniques actually have a long history in communication systems and in fact precede many of the present-day methods of DAC. The method finds its roots in a French patent (11) though the seminal paper appeared later in 1952 by de-Jager (12). However in this guise called delta-modulation, the 2-level code represented not the amplitude but the slope of the analogue input signal, consequently signal recovery required both low-pass filtering and signal integration. Although integration reduced substantially high-frequency noise, difficulty can be experienced in representing dc (or low-frequency) signals and the coder exhibits a slope overload phenomena rather akin to slew-rate limiting in audio amplifiers. This later limitation is readily demonstrated by observing the response of an integrator to a regular sequence of 1 (or 0) pulses where a stair-case waveform results whose slope cannot be exceeded within the constraint of a binary sequence and integrator time constant.

It was not until 1962 (4) that Inose, Yasuda and Murakami introduced the delta-sigma modulator (DSM) where the process of integration was transposed from coder output to input, this endowed the digital sequence with a more direct correspondence with the message signal, where signal reconstruction is performed by low-pass filtering of the binary sequence.

The main contribution of Philips in establishing an effective means of DAC is to integrate the functions of interpolation and noise shaping onto a single chip where Fig 6 shows the basic architecture of the 2nd-order system. The IC accepts stereo-PCM data at 44.1 kHz by 16 bit in I^2S format and by using multi-stage interpolation converts this via a N = 2 noise shaper to a 11.2896MHz binary code. A refinement of the DAC is the introduction of a dither signal at 350 kHz the purpose of which is to reduce a phenomena called idle channel noise whereby at low input signals, chaotic bursts of ...0101... type patterns can produce "howls and whistles" rather than benign noise.

The real advantage of the 1-bit DAC is the hardware simplicity of both the converter and the analogue reconstruction filter together with a tolerance to waveform degradation in the output pulse sequence. Although a logic gate may offer a fast response (certainly pulse sequences ~ 12 MHz have few technical problems), the rise and fall times will not be instantaneous producing a pulse area error. However, providing the pulse shapes are consistent from pulse to pulse and both leading and falling edges are processed similarly, then the pulse edges have settled to their final value in less than the pulse period, then waveform degradation can be modelled as an ideal pulse modified by a linear filter. As a consequence, the pulse distortion does not represent a fundamental signal distortion. Of course, if artifacts of pulse rise times propagate over a period greater than a pulse period and non-linearity interact with successive pulses, the filtering is non-linear and some degradation should be anticipated.

Nevertheless, the potential for the 1-bit DAC is considerable where a performance approaching the theoretical ideal can be achieved without the level dependent errors and slew-induced distortion of transresistance amplifiers so common of conventional DACs. These factors in principal allow signal conversion which can demonstrate coding into the noise floor and thus approach more readily and without need for calibration, the theoretical boundary of digital coding theory.

(5) Conclusion

This paper has considered the fundamental principles of oversampling and noise shaping as applied to ADC and DAC in digital audio systems. The means by which sampling rate and sample resolution can be interchanged has been demonstrated where at one extreme full resolution Nyquist samples are employed while at the other a binary sequence of heavily oversampled pulses is used.

The advantages of such processing were highlighted where in summary the following points should be noted:

* exchange of system complexity from analogue to digital processing.
* more precise and repeatable processing from exploitation of DSP such as near-ideal anti-aliasing and reconstruction filtering.
* desensitisation to sample jitter in both ADC and DAC following through to greater accuracy in generating Nyquist samples (44.1 kHz, 48 kHz) for information transmission.
* reduced sensitivity to hardware imperfection
* enhanced coder linearity with better low-level conversion enabling more accurate representation of signals at or near the noise floor, even in the presence of large-signal components
* elimination of over-complex analogue processes such as filters, sample and hold, de-glitch together with improvements resulting from enhanced ground rail structure
* closer approximation to an ideal performance where a digital channel results in only bandlimitation with non-correlated, additive noise.

In the near future one can anticipate wider application of these techniques with hopefully corresponding gains in performance, where the engine for economic system development is the achievements of VLSI which impact not only the professional environment but also the general consumer market.

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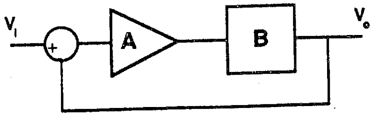
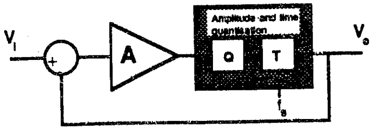
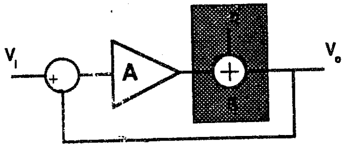


(a) Analogue amplifier

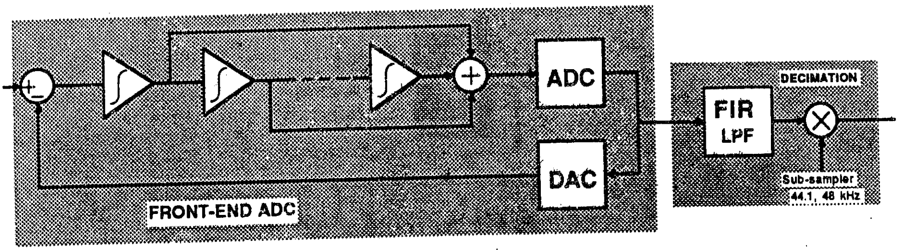


(b) Noise shaping ADC

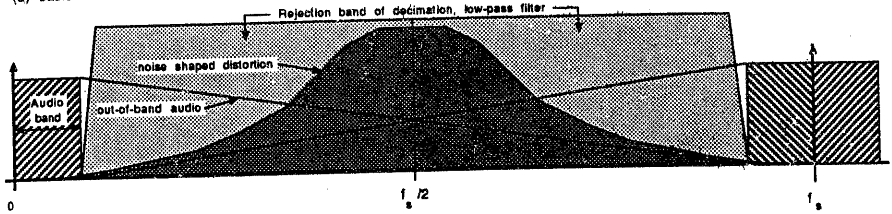


(c) Approximate additive analysis model

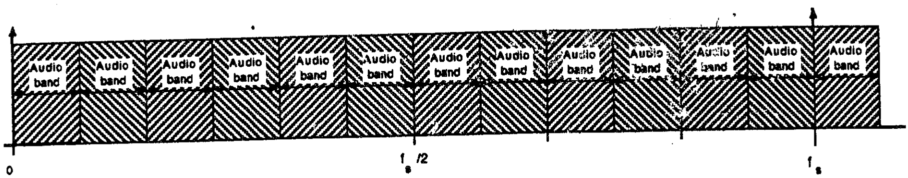
Fig. 1 Recursive noise shaping in both an amplifier and an ADC together with approximate additive distortion modelling.



(a) Basic front-end, oversampled ADC and single-stage dcimation system.



(b) Stectrum of oversampled and noise shaped signal at output of front-end ADC



(c) Decimated signal showing replication of audio spectrum

Fig. 2 Oversampled and noise shaped ADC with single-stage decimation system.

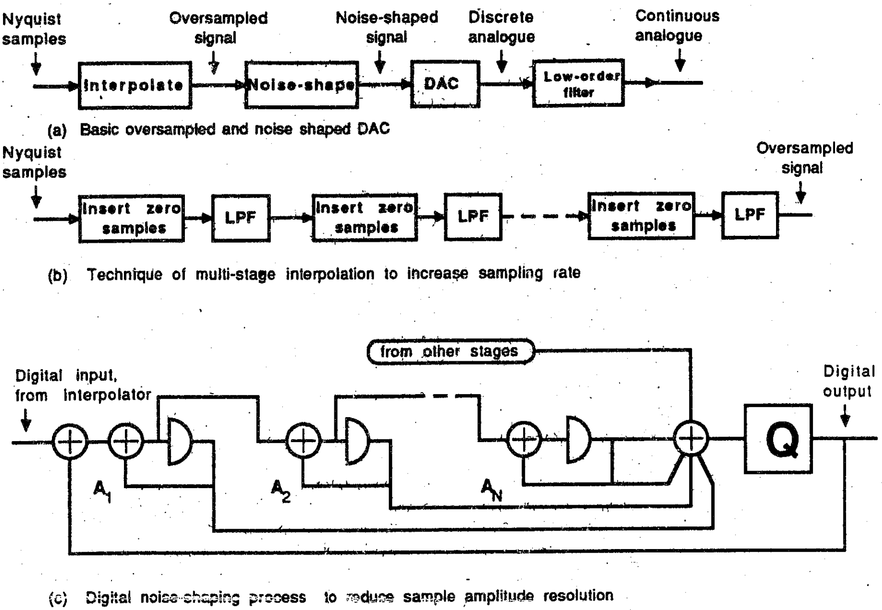


Fig. 3 Oversampled and noise shaped DAC system

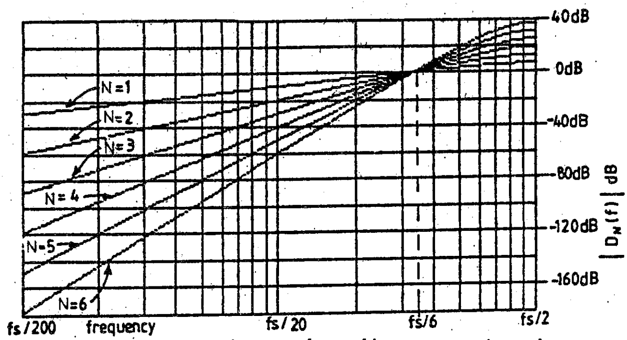
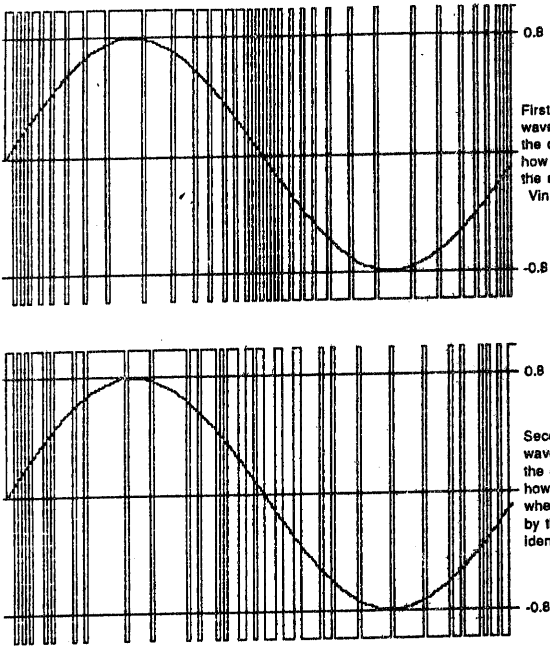


Fig. 4 Family of $|D_R(f)|$ for N=1 to 6 noise shaper



First-order delta-sigma modulator output pulse waveform for a sinusoidal input signal. Note that the output pulses are 100% duration, also observe how low-amplitude inputs are poorly coded where the output sequence follows a ...010101... pattern.
Vin = 0.8SIN(2pift), "ONE" = +1, "ZERO" = -1

Second-order delta-sigma modulator. output pulse waveform for a sinusoidal input signal.Note that the output pulses are 100% duration, also observe how low-amplitude inputs are encoded more accurately where there is evidence of greater pulse activity by the breakdown of ...010101... sequences.Input is identical to first-order case.

Fig. 5 First- and second-order delta-sigma modulation with example time domain output sequences.

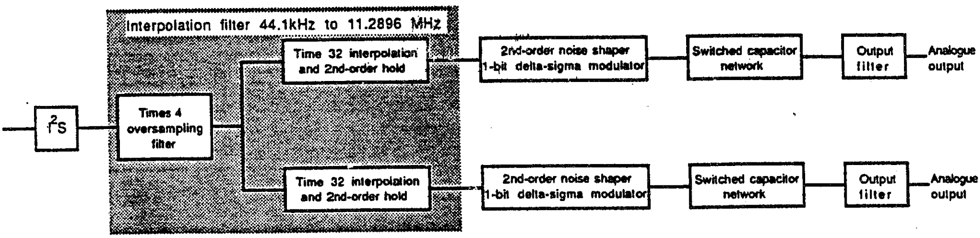


Fig. 6 1-bit DAC based on Philips system using 256 oversampling and 2nd-order delta-sigma modulation.