ARCHIVES OF ACOUSTICS
18, 2, 323–335 (1993)

OVERSAMPLING ANALOG TO DIGITAL CONVERSION IN ACOUSTIC MEASUREMENTS

P. KLECZKOWSKI

Institute of Mechanics and Vibroacoustics
Academy of Mining and Metallurgy
(30-059 Kraków, al. Mickiewicza 30)

The new method of analog to digital conversion known as "oversampling" or "delta-sigma", is inherently more suitable for acoustics measurements than the traditional method was. The theoretical background of this new method is presented. Both methods of conversion are compared with reference to their use in acoustics. The tests of parameters important in processing of acoustics signals performed on monolithic "delta-sigma" converters are reported. A digital procedure for evaluating the THD+N parameter is proposed.

1. Introduction

Contemporary means of digital processing of acoustic signals are very powerful. This power comes from developments in hardware (monolithic digital signal processors) and from developments in the theory of signal processing. Wide access to theoretically unlimited precision of digital processing evokes the need for inexpensive analog to digital conversion systems introducing low error.

The recent developments in microelectronic technology made possible a large scale production of integrated circuits performing the task of analog to digital (*A*/*D*) and digital to analog (*D*/*A*) conversion according to the so-called "delta-sigma" conversion technique, known also as "sigma-delta" or "one-bit" conversion. This technique is the most popular version of a larger defined method of "oversampling" conversion. The actual method is not new. It stems from the patent of C. Cutler (filed in 1954), describing a multibit implementation in vacuum-tube technology [6]. The first presentation of the one-bit version of this method comes from 1962 [7]. The authors called their method "delta-sigma modulation".

The delta-sigma conversion technique applies very well to monolithic implementation in CMOS technology, which is substantial for low cost mass production. However, such production has become practicable only recently.

Oversampling *A*/*D* converters supersede the systems with "classical" *A*/*D* converters in most of the applications requiring high resolution of 16 bits or more. Achieving the parameters of monolithic delta-sigma conversion systems with „classical" method is possible but much more expensive.

2. The classical data acquisition system

This system results from the function performed by a "classical" A/D converter. Any circuit, converting an isolated analog sample into a digital code representing its value will be meant under this term. Such converters are in fact quantizers. A full A/D interface must usually include also a low-pass antialiasing filter and a sample-and-hold (S/H) circuit (Fig. 1) [3, 5].

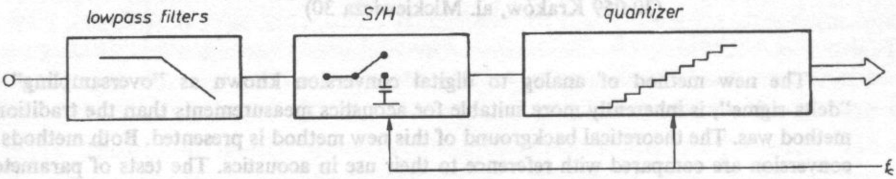

Fig. 1. "Classical" data acquisition system.

The systems performing all three tasks of Fig. 1 will be called data acquisition systems.

The realization of a lowpass filter of Fig. 1 is not easy. It must have sharp cut-off slope (as required to reduce sampling frequency) and very flat passband characteristics. It should not introduce phase and harmonic distortion and generate noise. It is very difficult to obtain a filter with such characteristics in monolithic technology.

The S/H circuit can be a source of nonlinear distortion or sampling jitter.

In classical quantizers it is not easy to exceed the resolution of 16 bits, because of the 2^N factor (where N is the number of bits), magnifying technological difficulties (and hence costs). This factor is present in all three basic methods of traditional A/D quantization. In successive approximation method it is the precision of components to the order of one part in 2^N, in flash conversion it is the number of elements and in integrating quantizers it is the speed of conversion that must increase with the 2^N factor.

Oversampling method eliminates the filter and 2^N factor from the implementation technology of the quantizer. It also makes the sample and hold circuit easy to realize.

3. Oversampling data acquisition systems

The term "oversampling" denotes sampling an analog signal at frequency deliberately above the rate resulting from the sampling theorem for that signal. The ratio of actual sampling frequency to the frequency given by the sampling theorem will be denoted by D.

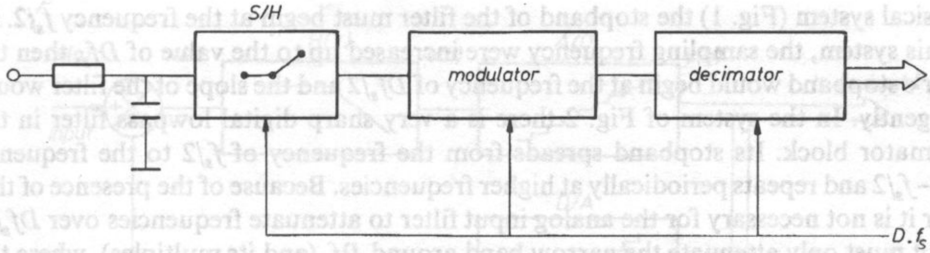

Fig. 2. Oversampling data acquisition system.

In oversampling method of A/D conversion the three separate tasks shown in Fig. 1 are performed in a different way. They can be seen to be redisposed to other three blocks, two of them specific to this method: a S/H circuit, a modulator and a decimator (Fig. 2).

In the bibliography the S/H circuit is usually not separated (because of its trivial implementation at relatively high sampling frequencies). It is included as a part of a modulator instead. In Fig. 2 it was shown, because of the method of analysis of the system used.

In such a system the task of antialiasing filtering is performed by an RC network at the input and the decimator, while the process of quantization should be assigned to both modulator and decimator. The modulator will be discussed below. The decimator is a low-pass digital filter which yields on its output the data stream at a rate reduced D times, that is at a rate close to that resulting from the sampling theorem. An important feature of the digital filter in the decimator is that it is a finite impulse response filter, hence it has linear phase characteristics despite very sharp slope in the transition band.

4. Oversampling

Oversampling is used in a system from Fig. 2 to achieve two different goals at the same time: to facilitate low-pass filtering and to increase the resolution of quantization.

Fig. 3 shows the spectrum of the signal of interest (up to f_B) together with some energy over the band of interest, at the input of any signal acquisition system. In the

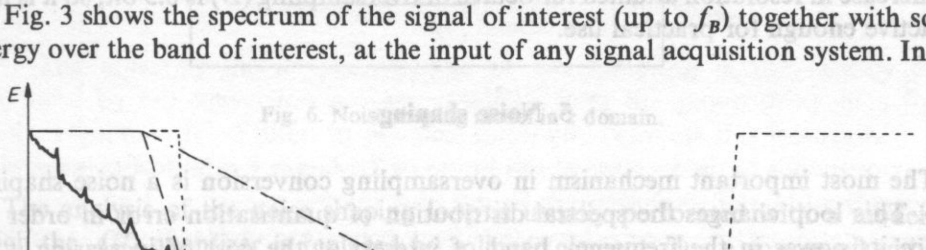

Fig. 3. Antialiasing in data acquisition system.

classical system (Fig. 1) the stopband of the filter must begin at the frequency $f_s/2$. If, in this system, the sampling frequency were increased up to the value of Df_s, then the filter's stopband would begin at the frequency of $Df_s/2$ and the slope of the filter would fall gently. In the system of Fig. 2 there is a very sharp digital lowpass filter in the decimator block. Its stopband spreads from the frequency of $f_s/2$ to the frequency $Df_s - f_s/2$ and repeats periodically at higher frequencies. Because of the presence of this filter it is not necessary for the analog input filter to attenuate frequencies over $Df_s/2$, and it must only attenuate the narrow band around Df_s (and its multiples), where the digital filter has its passband. In practice this task is performed well enough by a simple RC two-termination network.

The other task of oversampling is to increase the resolution of the quantization process. If the classical quantizer has on its input a signal of moderate amplitude (i.e. well above the value of a least significant bit and below the saturation limit) and of a band not too narrow, then the quantization error introduced resembles additive white noise. The spectrum of signal on the output of the quantizer can be, under such condition shown as in Fig. 4.

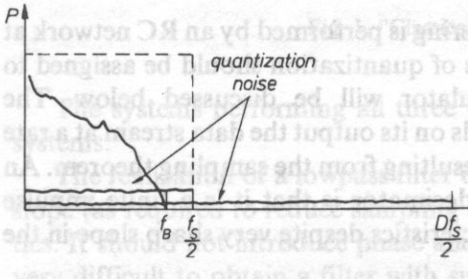

Fig. 4. Effect of oversampling on increasing the resolution of conversion

The power of the quantization noise corresponds to the resolution of the quantizer, but is independent of the sampling frequency so that at higher sampling rates this power will be distributed over a wider range of frequencies. Hence, if all spectral components over $f_s/2$ were filtered out, the power of quantization noise in the band of interest up to $f_s/2$ would be reduced by a factor of D. It can be shown [6] that the increase in resolution attained for octave of oversampling (D) is 0.5 bit, so it is not attractive enough for practical use.

5. Noise shaping

The most important mechanism in oversampling conversion is a noise shaping loop. This loop changes the spectral distribution of quantisation error in order to reduce its power in the frequency band of interest at the cost of increasing it at higher frequencies. The increased high-frequency noise is then filtered out by a very sharp digital filter. In fact, this is the main task of the digital filter in oversampling systems.

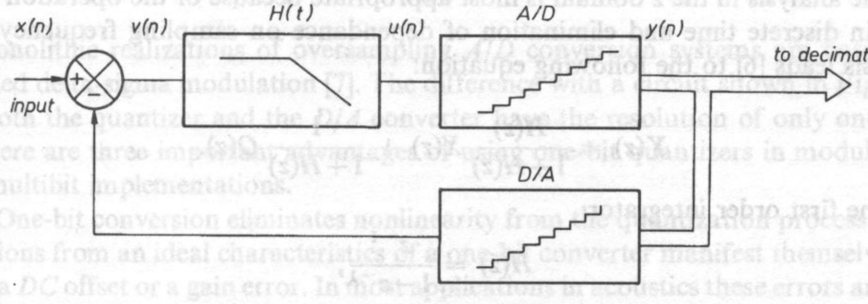

Fig. 5. Modulator

The diagram of modulator (from Fig. 2), where this process takes place, is shown in Fig. 5. The input of this circuit receives samples from the S/H circuit in discrete time, at the rate of Df_s. The discrete-time integrator (an analog accumulator) is described in the time domain by the following equation:

$$v(n+1) = v(n) + u(n). \tag{1}$$

The sequence of analog input samples $x(n)$ change their values slowly, because the input signal has been oversampled. The sequence $u(n)$ represents the approximation of quantization error of the N-bit A/D quantizer. This error varies strongly from sample to sample, while its long-term average is zero. The values of $y(n)$ represent a coarse approximation of $x(n)$, that oscillates over neighbouring values on the quantizer's scale around the slowly changing value of $x(n)$. The digital filter in a decimator block (Fig. 2) averages out the oscillation of $y(n)$, yielding a fine approximation of $x(n)$ [6, 12].

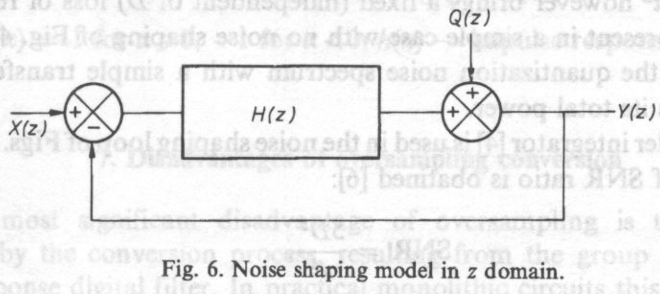

Fig. 6. Noise shaping model in z domain.

The analysis of the noise shaping loop is usually performed with a model, in which the A/D quantizer is replaced by a linear element plus a source of additive white noise (Fig. 6). The real D/A converter is replaced by an ideal converter with its output levels so trimmed, that it has no effect on the performance of the circuit and can be eliminated from the model. It should be emphasized, that the validity of such a model is limited by the character of the real quantization noise.

The analysis in the z domain is most appropriate because of the operation of the loop in discrete time and elimination of dependance on sampling frequency. This analysis leads [6] to the following equation:

$$Y(z) = \frac{H(z)}{1+H(z)} X(z) + \frac{1}{1+H(z)} Q(z). \tag{2}$$

For the first order integrator:

$$H(z) = \frac{z^{-1}}{1-z^{-1}}, \tag{3}$$

The comparison of power spectral density functions of the signal and the quantization noise in the passband of the decimator (Fig. 2) gives the following result [6]:

$$\mathrm{SNR} = \left[\frac{P_x}{P_q}\right] \frac{\frac{\pi}{2}}{\frac{\pi}{D} - \sin\left(\frac{\pi}{\mathrm{D}}\right)}, \tag{4}$$

where: SNR — the ratio of power of signal to the power of noise, $\frac{P_x}{P_q}$ — the SNR of the quantizer itself, operating on the same input signal $x(n)$ with no oversampling and noise shaping loop.

For $D \gg \pi$ it can be assumed, that

$$\mathrm{SNR} = \frac{3D^3}{\pi^2} \tag{5}$$

It can be shown [6], that the increase of resolution per octave of oversampling is 1.5 bits. The term $3/\pi^2$ however brings a fixed (independent of D) loss of resolution of about 1 bit, not present in a simple case with no noise shaping of Fig. 4. This is an effect of altering the quantization noise spectrum with a simple transfer function, which also boosts its total power.

If a second order integrator [4] is used in the noise shaping loop of Figs. 5 and 6, the following value of SNR ratio is obatined [6]:

$$\mathrm{SNR} = \frac{5D^5}{\pi^4}. \tag{6}$$

This corresponds to the increase of resolution of 2.5 bits per octave of oversampling with the fixed loss of resolution of about 2 bits.
Implementation of higher order loops is difficult because of their potential instability. Instead, more complex transfer functions are used, which exhibit a high order low-pass characteristics at low frequencies and more gentle response at higher frequencies, where the gain exceeds unity and a phase shift of π would cause oscillation [1]. Another solution are multistage noise shaping loops, called MASH [11].

6. The delta-sigma modulator

Monolithic realizations of oversampling A/D conversion systems are based on so-called delta-sigma modulation [7]. The difference with a circuit shown in Fig. 5 is that both the quantizer and the D/A converter have the resolution of only one bit.

There are three important advantages of using one-bit quantizers in modulators over multibit implementations.

1. One-bit conversion eliminates nonlinearity from the quantization process. Any deviations from an ideal characteristics of a one-bit converter manifest themselves as either a DC offset or a gain error. In most applications in acoustics these errors are not harmful and can be easily compensated. This elimination of nonlinearity opens way to very high potential resolution of this technique.

2. It is much easier to implement one-bit converters in monolithic technology.

3. One-bit signal on the decimator's input substantially simplifies the digital filter arithmetics and hence its monolithic implementation.

The obvious drawback is that higher order noise shaping loops or higher oversampling rates must be used to obtain the same resolution as those with multibit realizations.

The analysis of a delta-sigma modulator is difficult, because in the same circuit there are: a linear filter (integrator) and a strong nonlinearity — comparator (one-bit quantizer). If the model of Fig. 6 and (abstracted) equations (2–4) were used with one-bit quantizer, then the result obtained for the second order loop case would overestimate the SNR ratio by about 14 dB, as verified in near-ideal circuits and by careful simulation [6]. This model is also inappropriate, because the additive white quantization noise assumption is more likely to be not fulfilled in the one-bit case [6].

With an ideal D/A converter, the following nonlinear equation describes the performance of the one-bit loop in the time domain:

$$y(n+1) = \text{sgn}\,[h(n) * [x(n)-y(n)]] \tag{7}$$

where: $\text{sgn}\,(x) = 1$ for $x \geqslant 0$, -1 for $x<0$; $h(n)$ — impulse response function of the integrator.

7. Disadvantages of oversampling conversion

1. The most significant disadvantage of oversampling is the latency time introduced by the conversion process, resulting from the group delay of a finite impulse response digital filter. In practical monolithic circuits this latency is in the order of milliseconds and rises with lowering the sampling rate.

2. In classical conversion systems a maximum quantization error is comparable to the average quantization error. In oversampling conversion an error at individual samples can occasionally be noticeably higher than the average error, due to the stochastic nature of the conversion process.

3. An oversampling system must work with a fixed sample rate. It is not possible to acquire individual samples at arbitrary times.

4. The digital filter with very sharp cut-off slope generate oscillations when processing impulses. In classical conversion method of Fig. 1 this problem can be alleviated by increasing the system's sampling rate and using a filter with more gentle characteristics.

8. Basic tests of a monolithic delta-sigma converter

A monolithic implementation chosen for testing was a 16-bit two-channel converter from Crystal Semiconductor Co., type CS5336KP.

This particular integrated circuit was at the time of writing the only one that allowed for effective sampling rates to be as low as 1 kHz, kwhich is particularly useful in some acoustic research. The oversampling factor D in this circuit equals 64 and the manufacturer did not specify the type (order) of noise shaping loop used [13].

Out of many A/D converter parameters that can be specified [9], the author decided to measure and report only those having practical consequences in acoustic measurements, that is dynamic performance parameters.

Traditionally, such three basic parameters were: signal to noise ratio (SNR), total harmonic distortion (THD) and intermodulation distortion (IMD). In classical conversion methods nonlinear distortion introduced higher errors than quantization noise (except at low amplitudes of input signal). Excellent linearity of delta-sigma conversion reduces the THD and IMD errors to the level close to quantization noise. Therefore another, global parameter is usually used to assess these converters: total harmonic distortion + noise (THD+N). This parameter can be considered a measure of global dynamic error of a converter.

The author developed the following procedure for THD+N measurement: a spectrally pure sinusoid is presented to a converter and the output signal from the converter is compared to the ideal digitally generated sinusoid. The parameters of the ideal sinusoid are derived from the converter's output signal, by a least-squares method.

The following function was minimized:

$$E = \sum_{i=1}^{I} (c + a \cdot \cos(x \cdot i + \phi)) - s_i)^2, \tag{8}$$

where: I — the number of samples taken for this measurement (the band of analysis assumed was from 10 Hz to 22 kHz; this correspond to 4800 samples at 48 kHz sampling rate), c — the value of the DC component, a — amplitude, x — frequency, ϕ — phase, s_i — at value of the i-th sample of the signal at converter's output.

Having found the minimum of E function the THD+N coefficient was computed:

$$\text{THD+N} = 20 \log \left(\frac{A}{\sqrt{\frac{E_{\min}}{I}}} \right) \tag{9}$$

The constant A is the RMS value of full-scale sinusoid at the input, in the case of 16-bit long words it is 23170.

The initial values of parameters were derived by observation of digitally stored waveform. Then the minimum of E was found by a semi-automatic software procedure, which minimized E as a function of one of the parameters, fixing the remaining three. All four parameters were optimized in turn, and such iteration procedure was repeated three times, until E was minimized to the precision of three decimal digits, which corresponded to 0.02 dB of the final result of THD+N.

The practical measurement was difficult because of lack of the source of highly pure sinusoid. The purest sinusoidal signal that was available (from B&K 1022 oscillator) analysed with B&K 2134 analyser showed a noise floor of about -80 dB. Therefore the author decided to measure THD+N for a sinusoidal input signal that guaranteed to have all its distortion and noise components well below those of the converter, that is a -30 dB (related to full scale) signal.

An important advantage of oversampling converters is the lack of inherent sources of specific nonlinear distortion that can manifest strongly at low input signals [3], where quantization noise does not fulfill the assumptions of additive white noise. Therefore it was decided to test the converter's performance also at low amplitude (-80 dB) input signals.

The "idle channel noise" parameter (ICHN) was also measured, with input of the converter grounded. The computation algorithm was similar to the one shown in Eqs. (8) and (9), with a in (8) equal to zero.

The measurement setup was a special system interfacing two such monolithic double channel converters to the IBM PC AT computer, constructed by the author [8].

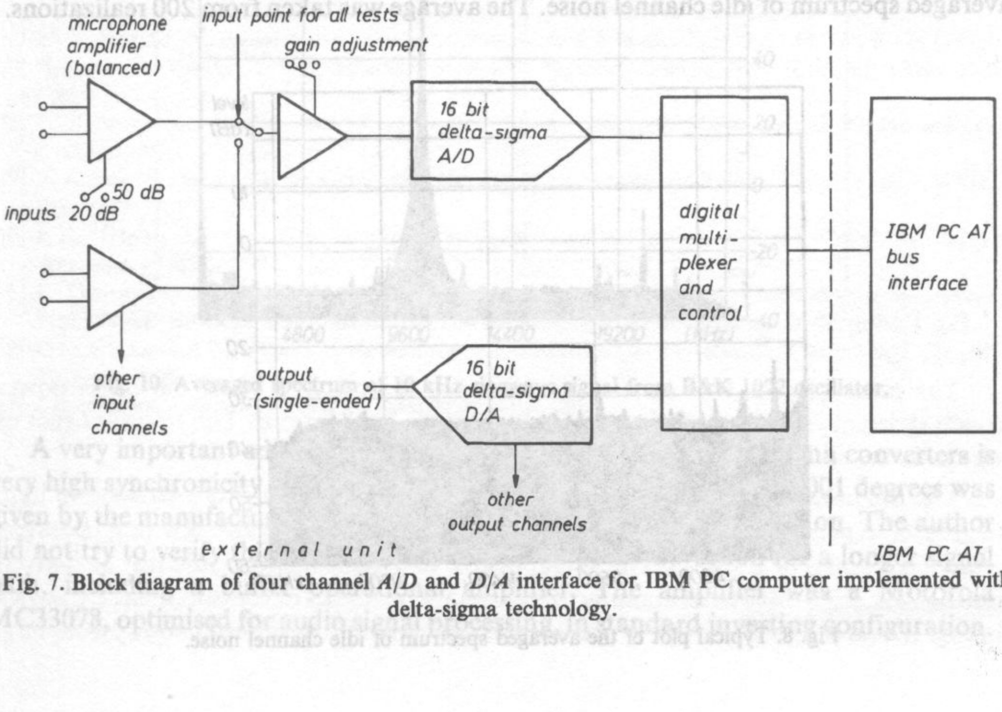

Fig. 7. Block diagram of four channel A/D and D/A interface for IBM PC computer implemented with delta-sigma technology.

This interface can receive four channels of simultaneously sampled analog signals and record them on computer hard disk. It can also reproduce analog signals through four D/A channels, using the similar (but inverted) delta-sigma method for D/A conversion. The block diagram of this interface is shown in Fig. 7. Input signal in all tests was from the B&K oscillator type 1022.

The tests were performed for three frequencies: 100 Hz, 1 kHz and 10 kHz. The results of tests [dB]:

	THD+N/−40 dB			THD+N/−80 dB			ICHN
Frequency [kHz]	0.1:	1:	10:	0.1:	1:	10:	—
Left channel	92.16	92.39	91.42	92.87	93.12	92.20	94.02
Right channel	93.71	93.53	92.88	93.27	93.45	93.04	94.11

The ICHN test was more representative statistically, as the results were averaged for three integrated circuits (ICs), each one of different manufacturing series. Ten realizations were taken on each IC. The results obtained in the ICHN test are better, probably because of shorter signal path in this case (direct grounding of the input pin on the IC instead of passing through a buffer operational amplifier).

In research of vibration it is often necessary to analyse a narrow band of low frequencies. The manufacturer's specification [13] gave 1 kHz as the lowest limit of effective sampling frequency. It was found that the converter still worked at the frequency of 187.5 Hz. The THD+N with a 30 Hz sinewave input signal and ICHN results were still at similar level.

The 1024-point FFT analysis was also performed. Fig. 8 shows a typical plot of the averaged spectrum of idle channel noise. The average was taken from 200 realizations.

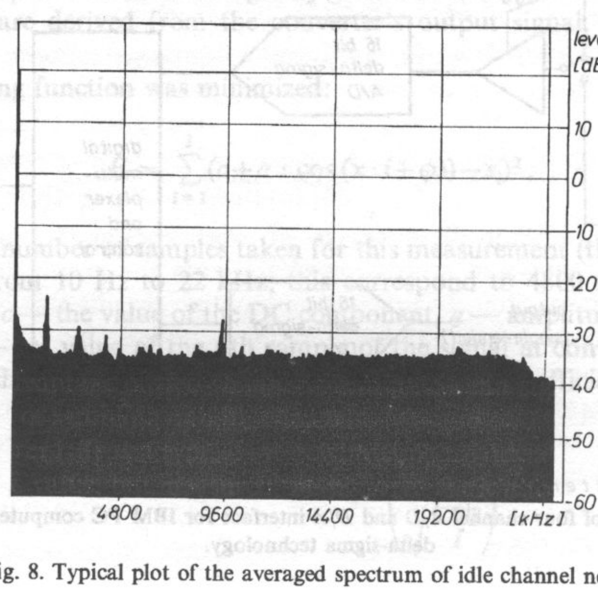

Fig. 8. Typical plot of the averaged spectrum of idle channel noise.

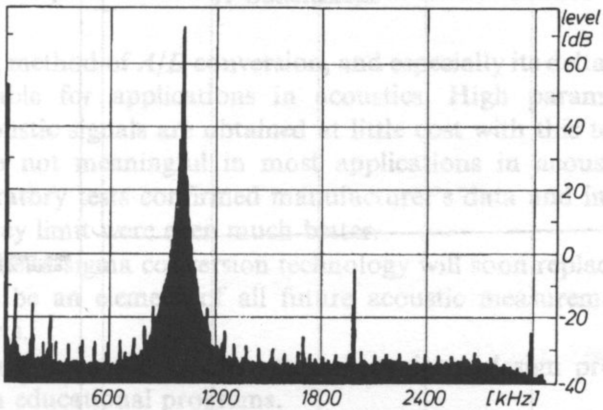

Fig. 9. Averaged spectrum of 1 kHz sinewave signal from B&K 1022 oscillator.

Fig. 9 shows the plot of averaged spectrum of a 1 kHz sinusoidal signal (from B&K 1022 oscillator). The sampling frequency for this case was 6 kHz (instead of 48 kHz for all other tests) and cosine window was used. The level of the input signal was deliberatery chosen so that its second and third harmonics can be seen but its noise floor does not exceed the noise floor of the conversion system (except for some disturbing peaks). A similar plot for 10 kHz sinewave is shown in Fig. 10.

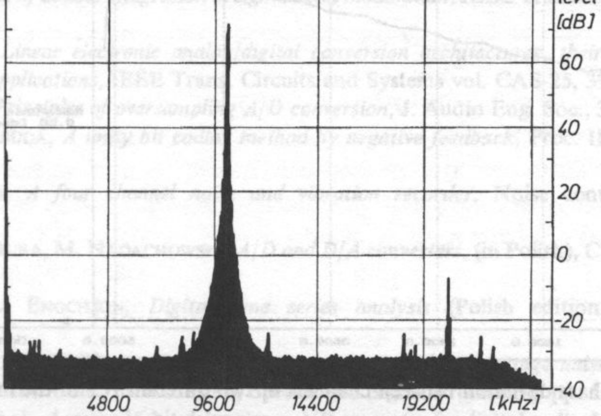

Fig. 10. Averaged spectrum of 10 kHz sinewave signal from B&K 1022 oscillator.

A very important advantage of two-channel monolithic delta-sigma converters is very high synchronicity of sampling in both channels. A figure of 0.0001 degrees was given by the manufacturer [13] as a typical interchannel phase deviation. The author did not try to verify this value, but measured the phase deviation for a longer signal path, including a buffer operational amplifier. The amplifier was a Motorola MC33078, optimised for audio signal processing, in standard inverting configuration.

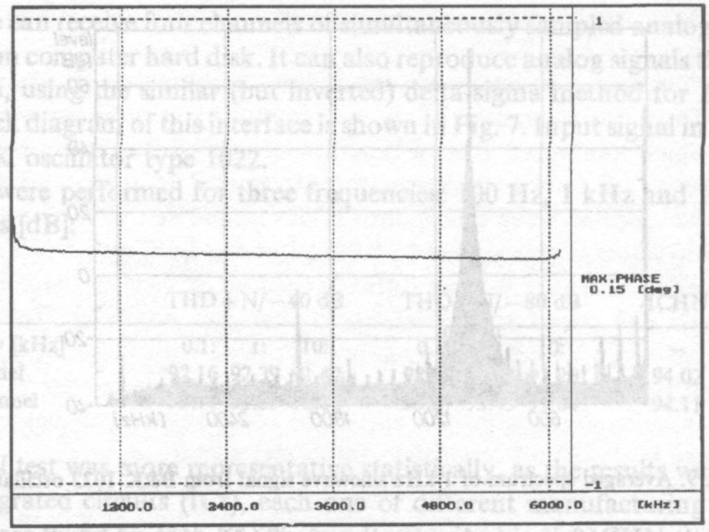

Fig. 11. Typical phase deviation between a pair of channels of the same IC.

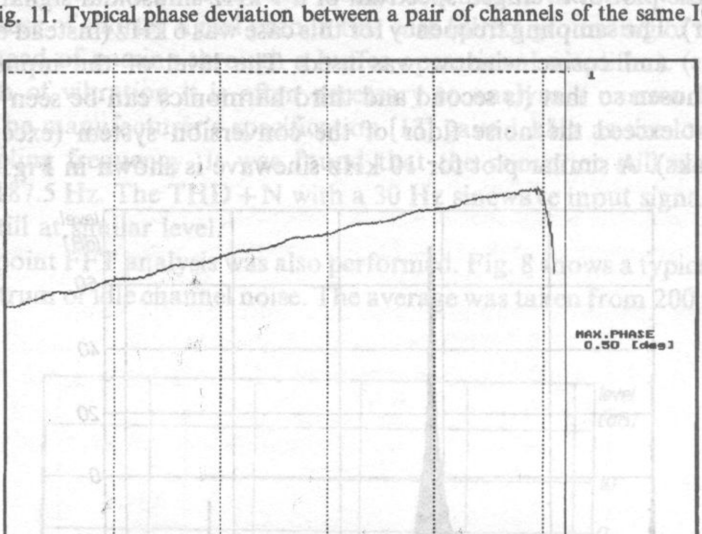

Fig. 12. Typical phase deviation between channels of two different ICs in the interface of Fig. 7.

The measurement was carried according to the method based on cross power spectral density spectrum [2, 10], with white signal on the input. The resultant plot of phase deviation versus frequency is shown in Fig. 11.

The architecture of the interface of Fig. 7 was designed in order to achieve high degree of synchronicity between channels belonging to different ICs (e.g. between channels 2 and 3 or 2 and 4). The actual degree of synchronicity could not be derived during the design stage from manufacturer's data. The results of measurement are shown in Fig. 12 and are good enough to consider all four channels as simultaneously sampling.

7. Conclusions

Oversampling method of A/D conversion, and especially its delta-sigma version is particularly suitable for applications in acoustics. High parameters needed in processing of acoustic signals are obtained at little cost with this technology. Their shortcomings are not meaningful in most applications in acoustics. The values obtained in laboratory tests confirmed manufacturer's data and in the case of low sampling frequency limit were even much better.

It seems that delta-sigma conversion technology will soon replace the traditional method and will be an element of all future acoustic measurement systems and consumer products.

This fact should be considered in research projects, design procedures of new equipment and in educational programs.

References

[1] R.W. Adams, P.E. Ferguson, A. Ganesan, *Design of single-bit noise-shaping loops with higher-order loop filters*, J. Audio Eng. Soc., **38**, 871 – 872 (1990).

[2] J.S. Bendat, A.G. Piersol, *Random data: analysis and measurement procedures*, Chapt. 1, Polish edition PWN, Warszawa 1976.

[3] B.A. Blesser, *Digitization of audio: A comprehensive examination of theory, implementation and current practice*, J. Audio Eng. Soc., **26**, 739 – 771 (1978).

[4] J.C. Candy, *A use of double integration in sigma delta modulation*, IEEE Trans. Comm., vol. COM-33, 249 – 258 (1985).

[5] B.M. Gordon, *Linear electronic analog/digital conversion architectures, their origins, parameters, limitations and applications*, IEEE Trans. Circuits and Systems vol. CAS-25, 391 – 418 (1978).

[6] M.W. Hauser, *Principles of oversampling A/D conversion*, J. Audio Eng. Soc., **39**, 1/2, 3 – 26 (1991).

[7] H. Inose, Y. Yasuda, *A unity bit coding method by negative feedback*, Proc. IEEE, **51**, 1524 – 1535 (1963).

[8] P. Kleczkowski, *A four channel noise and vibration recorder*, Noise control 92 Proceedings, Kraków 1992.

[9] Z. Kulka, A. Libura, M. Nadachowski, *A/D and D/A converters*, (in Polish), Chapt. 6 and 7, WKiŁ, Warszawa 1987.

[10] R.K. Otnes, L. Enochson, *Digital time series analysis* (Polish edition) Chapt. 6, WNT, Warszawa 1978.

[11] K. Uchimura et. al., VLSI *A to D and D to A converters with multi-stage noise shaping modulators*, Proc. 1986 IEEE Int. Conf. on Acoustics, Speech and Signal Processing, pp. 1545 – 1548.

[12] D.R. Welland et al., *A stereo* 16-bit *delta-sigma A/D converter for digital audio*, J. Audio Eng. Soc., **37**, 476 – 486 (1989).

[13] *Crystal Semiconductor Corporation Data Book* vol. 1, 1990.