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Enhanced Digital Signal Processing of Signal-dependently Sampled Data

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Introduction

The evolution of electronic circuits has been drastically advanced during the recent years. Introduction of new nano-materials and technologies allows to miniaturize devices considerably. As a result, engineers are developing much faster, more complex and smarter systems. This progress makes them attractive for expanded range of applications. A System-on-Chip (SoC) technology is considered as a revolutionary approach in electronics. It means the packaging of all the necessary electronic circuits and parts for a "system" on a single integrated circuit [1]. For example, a system-on-chip for a sound-detecting device might include an audio receiver, an analog-to-digital converter (ADC), a digital signal processor, necessary memory as well as the input/output logic control for a user. Consequently, SoC may contain digital, analog and often mixed-signal functions – all on one chip.

Most of today's electronics systems are based on synchronous paradigm - the work of a whole system is ruled by a clock. All the states in a design have to be changed at the same time. Since the systems constantly become more sophisticated and capacious, it is established that 10-30% of their resources (size, power consumption etc.) are wasted to circuits, which provide the synchronism of the device. Thus, it is relevant to explore other paradigms for development of future microelectronics based on alternative and innovative solutions.

One of the alternative approaches, which was introduced in the mid 50s and now receives increasing interest, is asynchronous logic. The key benefits of the asynchronous system are lower power consumption, absence of the clock screw, reduced heat elimination, lower EMI, automatic adaptation to physical properties, etc. [2]. Asynchronous design is characterized by the absence of a global clock, instead of that the system is driven by events, which arise during the execution of actual task. Such a change of paradigm leads to a drastic rearrangement in electronic designs by complete rethinking their architecture and signal processing methods.

Typically the information obtained from sensors has continuous time nature, while modern signal processing techniques are based on digital methods. The conversion from analog (continuous) to digital (discrete) presentation of the source data is being performed by ADC. In systems designed according to synchronous paradigm, ADC is driven by system clock. If uniform sampling is used the ADC clock frequency is determined by Nyquist theorem. If non-uniform sampling is involved, the mean sampling density can be below the Nyquist rate, however the sampling point flow is still derived from the clock. The asynchronous systems do not have a clock at all. That leads to the question - how to manage the analog-to-digital conversion in a case without clock? The asynchronous circuits on its merits are event-driven, so also the signal sampling process should be organized by events, which are got from the information presented in signal. New class of ADCs based on signal-dependent sampling schemes has to be developed.

The three most popular types of signal-dependent sampling – zero-crossing [3], reference signal crossing [4] and level-crossing [5] are illustrated in Fig. 1. Naturally, each of them has its own advantages and limitations, however there are some important joint features, which have to be taken into account performing the processing of signal-dependently sampled data:

1. in general case the signal samples are spaced non-uniformly,
2. it is impossible to determine the sampling time instants in advance,
3. local sampling density depends on local statistical characteristics of signal.

In this paper the samples obtained by level-crossing approach are chosen as an example of input data for digital signal processing. However, the developed methods are useful also for other cases of signal-dependently sampled data.

Non-uniformly spaced data

The techniques of non-uniform sampling are discussed in literature for many years [6]. In most cases the problem is stated from the following point of view - how to calculate the sampling time instants in a way, which allows to gain some advantages during the processing. The suppression of frequency aliasing effect can be quoted as an example. In this case the sampling point flow is deliberately pseudo-randomized, and it has to satisfy certain requirements. The statistical characteristics of such a sampling process are independent of input signal. In electronic design that leads to the task to clock the "classical" ADC at predetermined, non-uniformly spaced time instants with high accuracy.

The signal-dependent sampling also provides signal samples at the non-uniformly spaced time moments. Their nature and properties can considerably differ from deliberately pseudo-randomized case. Let us discuss, as an example, the level-crossing sampling approach. Higher frequency in a spectral presentation of signal provides faster changes in its waveform. If sampling is organized as events of levels crossing that leads to the higher density of signal samples (see Fig.1.c). Statistical characteristics of obtained sampling point flow directly depend on input signal. Furthermore, if a signal waveform has some regularity, the event flow of level-crossing sampling has the same regularity as well. As a result of this effect, the following problem can be stated - the methods developed for processing of deliberately non-uniformly sampled data can be impracticable for signal-dependently non-uniformly sampled data. This situation will be illustrated in the next section.

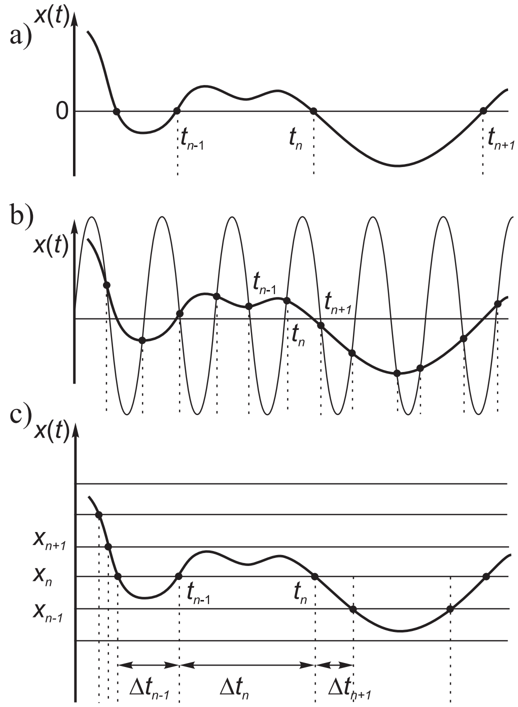

Fig. 1. Signal-dependent sampling schemes: a) zero-crossing; b) reference signal crossing; c) level-crossing

General Discrete Fourier transform

The classical method for digital signal analysis is a pair of Discrete Fourier Transform (DFT). The direct DFT allows to obtain the spectral values from N signal samples if they are equidistantly spaced in time:

$$X(m) = \sum_{n=0}^{N-1} x(n)\exp(-j2\pi mn/N)\,. \tag{1}$$

The inverse DFT allows to reconstruct signal values in time from signal spectral values at equally spaced frequencies:

$$x(n) = \frac{1}{N}\sum_{m=0}^{N-1} X(m)\exp(j2\pi mn/N)\,. \tag{2}$$

In practical applications typically exist two restrictions – the bandwidth of signal spectrum is limited to some higher frequency Ω, and the observation of signal waveform is limited to some duration Θ. The discrete signal samples $x_n = x(t_n)$, $n = \overline{0, N-1}$ are obtained by sampling procedure and they can be located arbitrary along the time axis. The general form of DFT, that enables the processing of non-equidistantly spaced data, can be expressed as:

$$X_m = \sum_{n=0}^{N-1} x_n \exp\left(-j2\pi \frac{m}{\Theta} t_n\right). \tag{3}$$

The band-limited nature of the signal provides the restriction for m: $\left|\frac{m}{\Theta}\right| \le \Omega$.

In the uniform sampling case expression (3) conforms to the formula (1). If appropriate deliberately non-uniform sampling scheme is used, expression (3) provides quite good spectral estimation for the cases, where sampling density is equal or higher than the Nyquist rate. Moreover, it is possible satisfactorily to estimate the spectral content of signal even if sampling density is below the Nyquist rate. The sampling point flow has to suppress the frequency aliasing for that. However, if formula (3) is applied to the data acquired by level-crossing sampling, the obtained result not always is adequate, despite the fact that signal always is oversampled. Let us illustrate this statement by a simple example. The mono-harmonic signal with frequency f_0 is sampled by 3-bit level-crossing ADC (7 levels). The spectral estimate obtained by formula (3) is shown in the Fig. 2. In addition to the spectral component at the true frequency (~0.0955 Hz) spurious components at the higher odd harmonics appear as well. Note, the issue of spurious components is not due to the frequency aliasing effect connected with insufficient sampling density, but due to specifics of signal and non-uniform samples obtained by level-crossing approach [5]. It leads to the conclusion that specialized enhancement of DFT should be derived to process the data acquired in such a way.

On the base of the Fourier series we can reconstruct the signal waveform from its spectral estimates by the following formula

$$\hat{x}(t) = \sum_{m=-M}^{M} X_m \exp(j2\pi t f_m)\,,\ t \in [0, \Theta]\,, \tag{4}$$

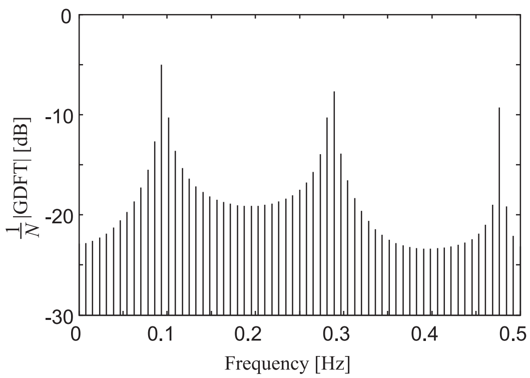

Fig. 2. Spectral estimate of sinusoidal signal with $f_0 = 12.22/\Theta$ ($\Theta = 128$ s.) sampled by 7 levels-crossings

where $f_m = \frac{m}{\Theta}$. Using the information about original signal $x(t)$, we can construct a reconstruction error as $\varepsilon(t) = x(t) - \hat{x}(t)$. The initial part of it is illustrated in the Fig. 3. The values $\varepsilon_n = \varepsilon(t_n)$ of reconstruction error at the sampling point instants t_n are shown by bold points. We can observe that the reconstruction error between samples used to be considerable higher than ε_n values. If sampling is based on the classical scheme, where sampling points are determined in advance, there are no restriction to the behavior of reconstructed signal between samples. In the case of level-crossing sampling the situation differs. Each level crossing is defined as an event and characterized by signal sample. Between them the waveform of reconstructed signal should not cross any level. Consequently the reconstruction error has to be minimized not only at sampling time instants, but also between them with the same accuracy. The minimization task

$$\int_0^{\Theta} \varepsilon^2(t)dt \to \min \qquad (5)$$

can be defined on the understanding that the signal values are known only at sampling points and the reconstructed signal is described by the expression (4). The minimization problem has to be solved with respect to spectral estimates $\{X_m\}$.

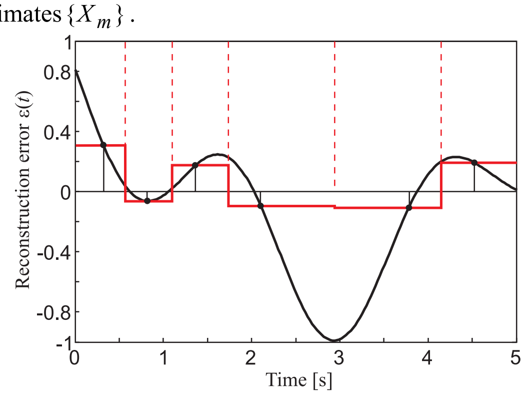

Fig. 3. Reconstruction error function (black curve) and zero-order interpolation (red piece-wise constant line) of ε_n (bold dots)

The solution of (5) can be based on two different approaches:

1) signal samples $\{x_n\}$ are interpolated within the time interval $[0\,\Theta]$ and the reconstruction error is expressed as $\varepsilon^{(x)}(t) = \tilde{x}(t) - \sum_m X_m \exp(j2\pi f_m t)$, where $\tilde{x}(t)$ is interpolated signal;
2) error samples $\varepsilon_n = x_n - \hat{x}_n$ are interpolated within the time interval $[0\,\Theta]$, and $\hat{x}_n = \sum_m X_m \exp(j2\pi f_m t_n)$.

Enhanced DFT based on signal interpolation.

Signal interpolation easily can be done by connecting the sample points with polynomials $p_n^k(t)$ of order k, or a band-limited interpolation can be performed as a sum of time-shifted sinc functions.

Let us rewrite (5) taking into account that signal samples are interpolated:

$$\int_0^{\Theta} \left(\tilde{x}(t) - \sum_{m=-M}^{M} X_m^{(x)} \exp\left(j2\pi f_m t\right) \right)^2 dt \to \min . \qquad (6)$$

To find the minimum, all the individual derivatives of X_m have to be considered as being equal to zero. We obtain $2M+1$ linear equations

$$2\int_0^{\Theta} \left(\tilde{x}(t) - \sum_{k=-M}^{M} X_k^{(x)} \exp(j2\pi f_k t) \right) \cdot \exp(j2\pi f_m t)dt = 0 \quad (7)$$

for $m = \overline{-M, M}$. Taking into account that $f_m = \frac{m}{\Theta}$, the $\{\exp(j2\pi f_m t)\}$ is a set of orthogonal functions into interval $[0\,\Theta]$. Thus we can write:

$$X_m^{(x)} = \frac{1}{\Theta} \int_0^{\Theta} \tilde{x}(t) \exp(j2\pi f_m t)dt . \qquad (8)$$

The expression (8) is similar with the formula of calculation of Fourier series coefficients for signal $\tilde{x}(t)$.

If signal samples $\{x_n\}$ are interpolated with zero-order polynomials (like a piece-wise constant line in the Fig. 3), we obtain

$$X_m^{(x0)} = \sum_{n=0}^{N-1} x_n \int_{\frac{t_n + t_{n-1}}{2}}^{\frac{t_n + t_{n+1}}{2}} \exp(j2\pi f_m t)dt = \frac{j}{2\pi f_m} \sum_{n=0}^{N-1} x_n \exp(j2\pi f_m t_n)(1 - \exp(-j2\pi f_m \Delta t')), \qquad (9)$$

where $\Delta t'_n = \frac{t_{n+1} - t_{n-1}}{2}$.

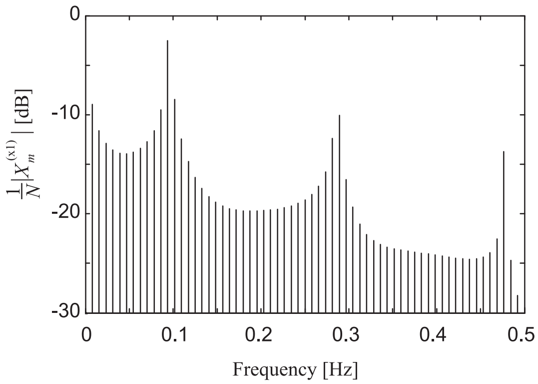

Fig. 4. Spectrum of mono-harmonic test-signal estimated by first-order signal interpolation algorithm

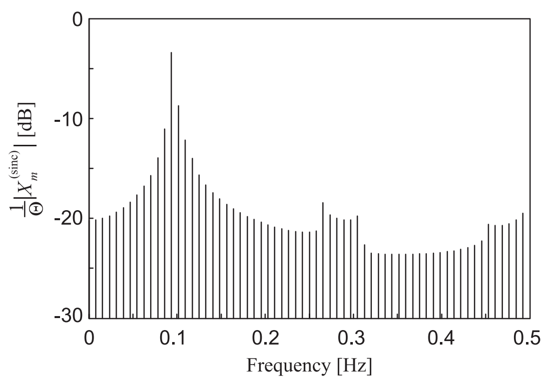

Fig. 5. Spectrum of mono-harmonic test-signal estimated by sinc interpolation algorithm

For piece-wise linear interpolation between the samples we can use polynomial $p_n^1 = \alpha_n(t - t_n) + x_n$, where $\alpha_n = \frac{\Delta x_n}{\Delta t_n}$, $\Delta x_n = x_n - x_{n-1}, \Delta t_n = t_n - t_{n-1}$, which gives:

$$X_m^{(x1)} = X_m^{(x0)} + \frac{1}{(2\pi f_m)^2} \cdot \sum_{n=0}^{N-1} \alpha_n \exp(j2\pi f_m t_n)(1 - \exp(-j2 f_m \Delta t_n)) + \frac{i}{2\pi f_m} \sum_{n=0}^{N-1} \alpha_n \Delta t_n \exp(j2\pi f_m t_n) \exp(-j2\pi f_m \Delta t_n). \quad (10)$$

Another approach is based on interpolation of $\{x_n\}$ with sinc functions. As the bandwidth of signal is limited to Ω, we can write:

$$\tilde{x}^{(\mathrm{sinc})}(t) = \sum_{k=0}^{K-1} c_k \,\mathrm{sinc}(2\Omega(t - kT)), \quad (11)$$

where $T = \frac{1}{2\Omega}, K : kT < \Theta$. Amplitudes c_k can be found from a linear equation system:

$$x_n = \tilde{x}^{(\mathrm{sinc})}(t_n). \quad (12)$$

In this case the enhanced discrete Fourier transform becomes

$$X_m^{(\mathrm{sinc})} = \sum_{k=0}^{K-1} c_k \exp(-j2\pi f_m kT). \quad (13)$$

Note, the last method, besides the DFT complexity of calculations, requires the solution of linear system with N equations and with $2M + 1$ unknowns.

Enhancement of Discrete Fourier transform based on interpolation of error samples.

Like the interpolation of signal samples the continues time reconstruction error function $\tilde{\varepsilon}(t)$ can be obtained from its values $\varepsilon_n = x_n - \hat{x}_n$ at time instants $\{t_n\}$. The estimates of reconstructed signal are calculated as $\hat{x}_n = \sum_m X_m \exp(j2\pi f_m t_n)$. The problem (5) in this case can be interpreted as minimization of area under the function $\tilde{\varepsilon}^2(t)$. If ε_n^2 is interpolated by zero–order polynomial the minimization task becomes as:

$$\sum_{n=0}^{N-1} \left(x_n - \sum_{m=-M}^{M} X_m^{(\varepsilon 0)} \exp(j2\pi f_m t_n) \right)^2 \cdot \Delta t'_n \to \min. \quad (14)$$

After the derivation the system of linear equations is formed

$$2 \sum_{n=1}^{N} \left(x_n - \sum_{k=-M}^{M} X_k^{(\varepsilon 0)} \exp(j2\pi f_k t_n) \right) \exp(j2\pi f_m t_n) \times \Delta t'_n = 0 \quad (15)$$

This system rewritten in matrix form is

$$\mathbf{\Psi x} = \mathbf{X}^{(\varepsilon 0)} \cdot \mathbf{\Phi} \cdot \mathbf{\Psi T}, \quad (16)$$

where $\varphi_{mn} = \exp(j2\pi f_m t_n)$ and $\psi_{mn} = \varphi_{mn} \cdot \Delta t'_n$.

Solution of (16) can be found as

$$\mathbf{X}^{(\varepsilon 0)} = (\mathbf{\Psi} \cdot \mathbf{x}) \cdot (\mathbf{\Phi} \cdot \mathbf{\Psi}^{\mathbf{T}})^{-\mathbf{1}}, \quad (17)$$

where $(\cdot)^{\mathbf{T}}$ and $(\cdot)^{-\mathbf{1}}$ denotes the transpose and inverse operation of matrix respectively.

The first-order polynomial interpolation of error function samples provides the following minimization task:

$$\frac{1}{2} \left(\sum_{n=0}^{N-2} \varepsilon_n^2 \cdot \Delta t_n + \sum_{n=1}^{N-1} \varepsilon_n^2 \cdot \Delta t_{n-1} \right) \to \min. \quad (18)$$

It looks like a sum of two zero-order interpolation problems. The solution in this case is similar to the previous one expressed as (17):

$$\mathbf{X}^{(\varepsilon 1)} = \left(\mathbf{\Psi}'' \cdot \mathbf{x}' + \mathbf{\Psi}'' \cdot \mathbf{x}' \right) \left(\mathbf{\Phi}'' \cdot \mathbf{\Psi}''^{\mathbf{T}} + \mathbf{\Phi}'' \cdot \mathbf{\Psi}''^{\mathbf{T}} \right)^{-\mathbf{1}}, \quad (19)$$

where $\mathbf{\Phi}', \mathbf{\Psi}', \mathbf{x}'$ and $\mathbf{\Phi}'', \mathbf{\Psi}'', \mathbf{x}''$ matrices are formed from $\mathbf{\Phi}, \mathbf{\Psi}, \mathbf{x}$ by using indexes $n = \overline{0, N-2}$ and $n = \overline{1, N-1}$ respectively.

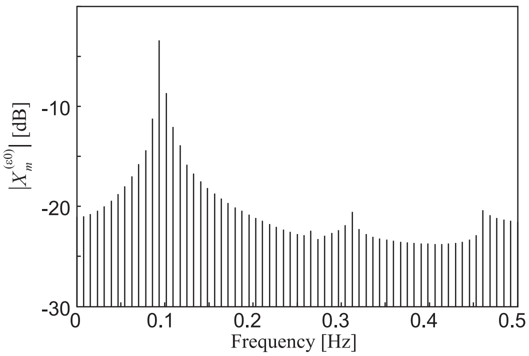

Fig. 6. Spectrum of mono-harmonic test-signal estimated by zero-order reconstruction error interpolation algorithm

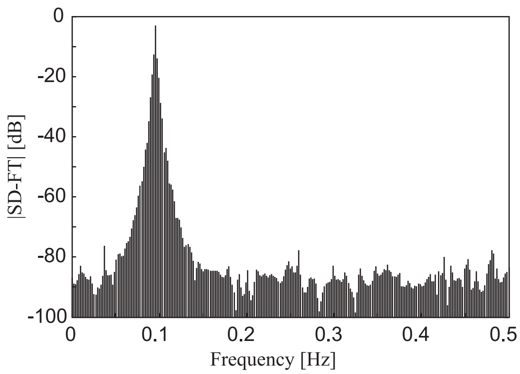

Fig. 7. Spectrum of mono-harmonic test-signal estimated by signal-dependent transformation

Simulation results

The performance of described algorithms has been investigated by computer simulations. The above mentioned single-tone sinusoid has been used as a test-signal. The input data are acquired from $\Theta = 128[\text{sec}]$ long observation as events of 7 levels-crossings. The frequency of test-signal is $12.22/\Theta \approx 0.0955[\text{Hz}]$ that does not lie on the frequency grid of Fourier series. The number of acquired samples is $N = 172$.

The spectral estimate obtained by General DFT (expression (3)) and illustrated in Fig. 2 can be used as a reference, which allows identifying the improvement of processing, when enhanced methods are exploited. The Fig. 4 shows spectral result obtained by first-order signal interpolation algorithm (expression (10)). Although the amplitudes of higher spurious harmonics are decreased for about 5dB, they are still considerable. The complexity of this algorithm is few times higher then DFT. The zero-order signal interpolation gives slightly worse result.

The band-limited interpolation with sinc functions (expression (13)) provides improved spectral presentation of test signal, which is shown in the Fig. 5. The spurious harmonics are suppressed. The drawback of this algorithm is that beside the DFT calculation it requires the solution of the linear equation system as well.

The similar mathematical complexity is also for methods, which use reconstruction error interpolation by zero- or first-order polynomials. The results obtained in both cases are quite similar, and therefore we present here only the simplest case – zero-order error interpolation, in the Fig. 6.

The common feature of all presented algorithms is spectral analysis on the grid of Fourier frequencies $f_m = m/\Theta$. For the expressions (9) and (10) the motivation is to build the orthogonal basis of transformation. The expressions (17) and (19) can be considered as unorthogonal transformations. That allows to use also frequency grid with higher density, which can improve spectral resolution. However, the necessity to solve the equations system limits the grid – the number of analysis frequencies has to be equal or less than the number of samples. Note, the number of samples in signal-dependent sampling case is not known in advance, because it depends on actual signal properties. The method which overcomes this problem is so called signal-dependent Fourier transform described in [7]. The frequency grid of this algorithm is independent on number of samples and provides high spectral resolution. The Fig. 7 demonstrates the spectral estimate obtained by this method, where analysis grid is four times frequenter than in the DFT case.

The quality of simulated algorithms can be characterized by integral value of squared reconstruction error. The estimated values, averaged over different signal phases, are summarized in the Table 1. It shows that all methods decrease the reconstruction error in comparison with GDFT. The obtained improvement can vary from insignificant to several thousand times.

Conclusions

One of the basic tools in digital signal processing is the Discrete Fourier Transform. In this paper we presented several enhancements of DFT to make it convenient for analysis of signal-dependently sampled data, particularly paying attention to the level-crossing sampling. In this case the samples are not only spaced non-uniformly in time, but also its distribution depends on signal properties. The periodical signal provides regularities into sampling point flow, which leads to the appearance of spurious components (harmonics) in the spectrum estimated on the bases of conventional DFT analysis.

The level-crossing sampling technique determines not only signal value at sampling time instance, but also the rule, that signal values between two samples should not cross any quantization level. This condition can be taken into account during development of processing methods. The proposed idea is to minimize the error between original and reconstructed by Fourier series signals not only at sampling time instants, but also between them with the same accuracy. The problem lies in the fact that we know original signal values only at sampling instants. Two different approaches are considered. The first one is based on obtaining continues time signal by interpolation of known signal samples. The expressions for enhanced DFT calculation are proposed for zero-order and first order polynomial interpolation as well as for band-limited interpolation with sinc functions. The second approach to minimize continues time reconstruction error is based on

Table 1. Estimated reconstruction errors for different methods

Method	GDFT	Zero-order x_n interpolation	First-order x_n interpolation	x_n interpolation with sinc	Zero-order ε_n interpolation	First-order ε_n interpolation	Signal-dependent DFT-like method
$\int_0^{\Theta} \varepsilon^2(t)dt$	835	811	803	3.00	2.35	2.29	0.07

interpolation of error samples. The expressions for calculation of spectral coefficients are proposed for zero-order and first-order polynomial interpolation. Simulation results show the improvement of data processing if enhanced algorithms are used instead of classical General DFT. It should be mentioned that better results are achieved by algorithms, which require performing more mathematical operations. The best result can be reached by the signal-dependent DFT-like method, however the high computing complexity makes it unpractical for real-time applications.

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Asynchronous circuits receive now an increasing interest. Their promising advantages can play a significant role in future electronics' development. One of the factors, which delay the wide spread of the clock-less systems, is incompatibility with classical clock-driven analog to digital converters and processing algorithms. The new ADC approaches, which are based on signal-dependent sampling, are required, as well as convenient processing methods have to be developed. The paper presents the enhanced DFT-like algorithms derived from the idea to minimize the signal reconstruction error not only at sampling points, but also between them with the same accuracy. Two approaches for creating continuous time error function from its discrete values are investigated: 1) interpolation of signal samples; 2) interpolation of error samples. Level-crossing sampling has been used as an example for signal digitizing. Achieved advantages are demonstrated by simulations. Ill.7, bibl. 7 (in English; summaries in English, Russian and Lithuanian).

М. Грейтанс, И. Хомяковс. Усовершенствованная цифровая обработка дискретизованных по уровню сигналов // Электроника и электротехника. – Каунас: Технология, 2006. – № 4(68). – С. 9–14.

В настоящее время все больше внимания уделяется асинхронным системам. Однако для асинхронных систем обработки сигналов существует потребность в новых видах аналого-цифрового преобразования и соответствущих им алгоритмах обработки. Рассматривается подход к аналого-цифровому преобразованию сигналов, основанный на их дискретизации по уровню, и предлагаются усовершенствованные, ДФТ-подобные алгоритмы, позволяющие значительно уменьшить погрешность результатов цифрового анализа сигналов. Ил.7, библ. 7 (на английском языке; рефераты на английском, русском и литовском яз.).

M. Greitans, I. Homjakovs. Patobulintas adaptyviai diskretizuotų signalų skaitmeninis apdorojimas // Elektronika ir Elektrotechnika.– Kaunas: Technologija, 2006. – Nr. 4(68). – P. 9–14.

Vis didesnį susidomėjimą kelia asinchroniniai grandynai. Dėl daug žadančių privalumų jie gali vaidinti reikšmingą vaidmenį ateityje, tobulinant elektroniką. Vienas iš veiksnių, lėtinančių platų sistemų, nenaudojančių taktinio dažnio, paplitimą, yra jų nesuderinamumas su klasikiniais taktinio signalo valdomais analoginiais-kodiniais keitikliais ir apdorojimo algoritmais. Reikalingi naujų tipų keitikliai, kurių veikimas būtų grindžiamas adaptyviu (nuo signalo priklausančiu) diskretizavimu, taip pat turi būti sukurti patogūs apdorojimo metodai. Pateikiami patobulinti DFT algoritmai, kurie sukurti remiantis idėja minimizuoti signalo atkūrimo paklaidą ne tik diskretizavimo taškuose, bet ir tarp jų, tuo pačiu tikslumu. Nagrinėti du metodai, skirti tolygiai laikinei klaidos funkcijai sukurti iš jos diskretinių verčių: 1) diskretinių signalo verčių interpoliavimas; 2) diskretinių klaidos verčių interpoliavimas. Lygio kirtimo diskretizavimo būdas naudotas signalui skaitmenizuoti. Pasiekti patobulinimai demonstruojami naudojant modelius. Il. 7, bibl. 7 (anglų kalba; santraukos anglų, rusų ir lietuvių k.).

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