Time-interleaved oversampling convertors

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Indexing terms: Sigma-delta modulation, Data convertors, Integrated circuits

A new architecture is proposed which exploits the time-interleaving concept to increase the oversampling ratio in delta-sigma modulators. It is shown that the effective oversampling ratio is increased by a factor M through the use of M interconnected modulators. Although a high speed sample-and-hold circuit is still required for an analogue-to-digital convertor, speed constraints are significantly reduced for the majority of analogue parts such as loop filters, A/D and D/A blocks.

Introduction: One approach for realising high-speed Nyquist-rate convertors is the use of time interleaving [1]. In this approach, M independent convertors operate in a time-interleaved fashion to increase the effective conversion rate. In other words, each convertor operates at 1/M times the overall sampling frequency thus reducing their speed requirements. This approach was recently used to realise a 6 bit A/D convertor operating at 1GHz where the input sample-and-hold circuit was fabricated in GaAs, and the array of convertors was implemented in silicon [2]. Also recently developed was an approach to reduce harmonic distortion due to branch mismatches through the use of a quadrature mirror filter bank [3].

Oversampling convertors have become a popular method for realising high resolution convertors due to their reduced analogue circuit complexity [4]. The purpose of this Letter is to apply the time-interleaving concept to oversampling convertors resulting in an increase in the effective sampling rate. It should be noted here that a straightforward application of the time-interleaving concept to oversampling convertors results in only a small increase in resolution, ~3dB for every doubling of components. The approach described maintains the noise-shaping behaviour through the use of interconnections between modulators.

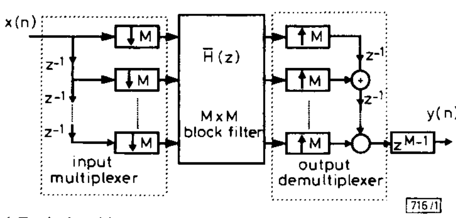

Fig. 1 Equivalent block filtering structure for single-input single-output transfer function H(z)

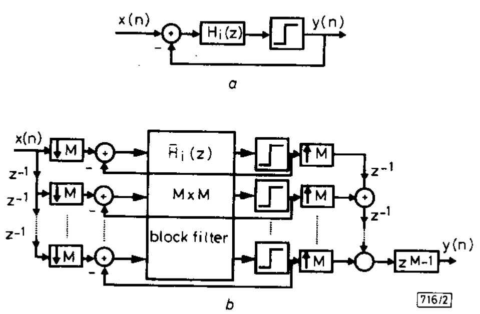

Fig. 2 Interpolative ΔΣ structure

a Conventional structure
b Time-interleaved structure

Proposed architecture: In time-interleaved Nyquist-rate convertors, the desired goal is to convert the largest possible input signal bandwidth assuming a fixed convertor resolution. In time-interleaved oversampling convertors, however, the desired goal can also be interpreted as maximising the overall resolution for a fixed input signal bandwidth. No matter which goal is being considered, the obstacle that limits the overall performance is the speed constraints of the analogue blocks.

To realise time-interleaved oversampling modulators, we use the concept of block digital filtering [5]. Consider the single-input single-output transfer function $Y(z) = H(z)X(z)$. An equivalent system with the same input-output transfer function is depicted in Fig. 1, in which $\bar{H}(z)$ is an M × M transfer function matrix, where $\bar{H}_{ij}$ represents the contribution of the jth input into the ith output. The general structure of $\bar{H}(z)$ is as follows:

$$\bar{H}(z) = \begin{bmatrix} E_0(z) & E_1(z) & E_2(z) & \dots & E_{M-1}(z) \\ z^{-1}E_{M-1}(z) & E_0(z) & E_1(z) & \dots & E_{M-2}(z) \\ z^{-1}E_{M-2}(z) & z^{-1}E_{M-1}(z) & E_0(z) & \dots & E_{M-3}(z) \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ z^{-1}E_1(z) & z^{-1}E_2(z) & z^{-1}E_3(z) & \dots & E_0(z) \end{bmatrix} \quad (1)$$

in which the elements of the first row of $\bar{H}(z)$ are type 1 polyphase components of H(z), or mathematically,

$$H(z) = \sum_{l=0}^{M-1} z^{-l} E_l(z^M) \quad (2)$$

Now consider the interpolative ΔΣ architecture as depicted in Fig. 2a. By inserting the equivalent block filter instead of $H_i(z)$ and applying several topological identities on the structure (using a linear model for the quantiser), we end up with the proposed architecture shown in Fig. 2b. Here, $\bar{H}(z)$ is the transfer function matrix derived based on eqns. 1 and 2. Note that the internal parts of the proposed architecture are operating at f_s/M where f_s is the overall sampling frequency, or if the two structures are operating at the same internal speed, the oversampling ratio of the proposed architecture is M times larger than that of the original structure. It is also worth mentioning that the time-interleaving concept is easily applicable to other ΔΣ structures such as the error-feedback or cascade-of-integrators topology. In fact, a second-order modulator based on the cascade-of-integrators architecture is used as an example for the simulations described below.

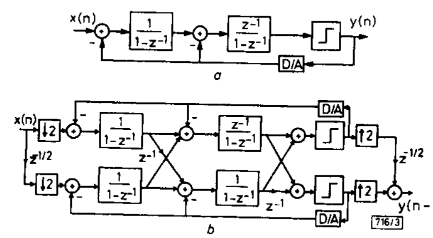

Fig. 3 Second-order ΔΣ modulator

a Conventional structure
b Time-interleaved structure

Simulation: A second-order ΔΣ modulator based on the cascade-of-integrators topology, shown in Fig. 3a, is used to demonstrate the time-interleaving concept. Its equivalent time-interleaved structure for M = 2 is depicted in Fig. 3b. The internal clock rates for both the conventional and time-interleaved modulators operate at a normalised frequency of 1, whereas only the input and output multiplexers for the time-interleaved structure operate at twice this frequency. A sinusoidal input with a frequency of 13/1024 was applied and the simulated output spectra are shown in Fig. 4. These results are obtained by averaging eight periodogram estimates where Hanning windows were first applied. Careful analysis shows that the signal-to-noise ratio for the time-interleaving structure is 15dB better than that for the conventional structure which is equivalent to increasing the oversampling ratio by a factor of 2 as expected. Also, it is worth mentioning that the input-output path in Fig. 3b has one more half delay ($z^{-1/2}$) in comparison to the same path in Fig. 3a.

It should also be noted that the proposed structure is less vulnerable to leaky integrators and hence less sensitive to finite gain-bandwidth of the operational amplifiers. Assuming that $H(z) = 1/(1-\alpha z^{-1})$, using eqns. 1 and 2 we find that $\tilde{H}(z)$ has elements with a common denominator of $1/(1-\alpha^2 z^{-1})$. Thus for example, if we realise the time-interleaved structure with leaky integrators having $\alpha^2 = 0.81$, then the effective overall transfer function would be $H(z) = 1/(1-0.9z^{-1})$ which is more ideal.

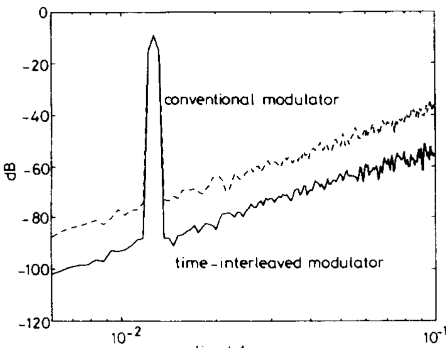

Fig. 4 Simulation results for output spectrum of conventional and time-interleaved second-order modulator

Finally, it is worth mentioning that, unlike time-interleaved Nyquist-rate convertors any mismatches between the branches do not cause harmonic distortion. However, mismatches do cause an increase in the quantisation noise floor, an effect which is often more favourable than having harmonic distortion.

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Optimal design of contiguous-band output multiplexers (COMUX)

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Indexing terms: Satellites, Telecommunications, Filters, Multiplexers

A general multiplexer design procedure without dummy channels is described. The formulas are applied to the optimal design of practical contiguous or non-contiguous band multiplexers consisting of multicavity filters distributed along a waveguide manifold. All design parameters can be directly optimised. An example of a practical 12 channel, 12GHz contiguous band multiplexer is presented.

Introduction: Advanced satellite communication systems require optimal use of the available frequency bands. For this reason, the number of transmit channels will increase, while the distance between channels will decrease. This means that future payloads will use contiguous channel systems with up to 12 channels. The use of optimisation techniques to determine the best multiplexer parameters is an indispensable part of the design procedure. A significant reduction in CPU time is obtained using powerful gradient-based minimax optimisation with a good starting point

Filter and multiplexer design: High performance in multiplexer design can only be obtained by using computer-aided design and advanced qualified technologies. The precondition for efficient CAD is the exact modelling of all physical multiplexer elements (i.e. T-junctions, filter couplings, resonance frequencies and manifold distances) by appropriate equivalent circuits [1]

The design procedure consists of a quasianalytical method first introduced by Rhodes and Levy [2]. This method uses filters which are derived from the same normalised Chebyschev lowpass equivalent prototype filter [3], although they have different centre frequencies and bandwidths. These filters are not sufficient for contiguous multiplexer applications because it is impossible to synthesise certain useful transmission characteristics (transmission zeros). Therefore, at first, we extend this method to the case of symmetrical quasi-elliptical filters with folded configuration (Fig. 1a) We then apply rotations [4] to convert a coupling matrix from a folded configuration to an in-line dual-mode configuration (input on the first resonator, output on the last). We obtain the coupling diagram shown in Fig. 1b.

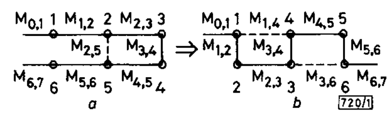

Fig. 1 Folded and inline networks coupling diagrams

a Folded
b Inline

The multiplexer under consideration consists of asynchronously tuned multicavity filters distributed along a waveguide manifold. The channel bandpass filters are six-pole dual-mode quasi-elliptic designs. Rhodes and Levy have shown that to considerably improve the multiplexer performance, it is sufficient to modify only the first two shunt susceptances, the first two admittance inverters of each channel filter and the physical separation between filters and the short-circuited manifold termination (phase lengths), all other elements remaining unchanged [1, 2]. Also, for each channel, we apply these modifications directly to the first two direct coupling coefficients (M_{01} and M_{12}), the first two resonance frequencies and the phase lengths. Fig. 2 illustrates the prototype manifold multiplexer.

This quasianalytical method is sufficient for a multiplexer with a small number of channels (<5) but not for more channels. Therefore, we need to use an optimisation procedure to optimise all design parameters. However, this method constitutes a good starting point for reducing the CPU time.

Formulation of problem: All design parameters can be directly optimised using a powerful gradient-based minimax algorithm [5]. The error function to be minimised is given by:

$$F(x) = \max f_j(x, \omega_j) \quad \text{with} \quad 1 \leq j \leq m \qquad (1)$$

where x is a k order vector of optimisation variables (we have 16 optimisation variables for each six-pole dual-mode multiple-coupled cavity filter: e.g. spacing lengths, channel input and output couplings, filter coupling parameters and filter resonance frequencies), m is the number of nonlinear functions $f_j(x,\omega_j)$.

The nonlinear function $f_j(x, \omega_j)$ is the return loss (in dB) of the common gate at the jth frequency point. All frequency points ω_j