Statistical Spectra and Distortion Analysis of Time-Interleaved Sampling Bandwidth Mismatch

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***Abstract*—Time interleaving is one of the most efficient techniques employed in the design of high-speed sampled-data systems. However, the mismatches appearing among different channels will create distortion tones that will degrade the system performance. This paper presents a detailed analysis of the effect of sampling bandwidth mismatches in the statistical behavior of time-interleaved sampling systems. Closed-form and handy signal-to-noise distortion ratio formulas will also be derived which allow quick evaluation of the system performance, and the MATLAB simulations are provided in order to verify the effectiveness of those formulas. Finally, the corresponding design procedure extracted from the formulas will be further addressed through a design example.**

***Index Terms*—Bandwidth mismatches, sampled-data systems, time-interleaved (TI) analog-to-digital converter (ADC).**

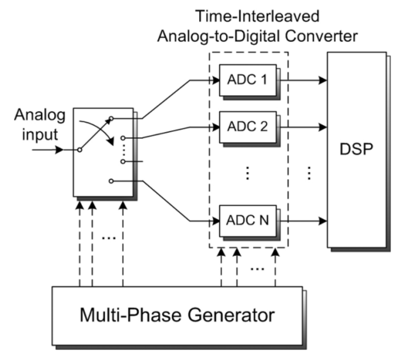

Fig. 1. Time-interleaved ADC.

I. Introduction

HIGH-SPEED sampled-data systems, e.g., analog-to-digital converters (ADCs), have found increasing importance in various electronics circuits and systems, such as in wireless communications, flat-panel displays, video and imaging signal processing applications. Time-interleaving (TI) in ADCs [1], as shown in Fig. 1, is one of the effective ways to boost the speed of sampled-data systems beyond the limitation imposed by the technology. However, special attention must be paid to avoid various types of mismatches arising in parallel identical channels, e.g., offset, [2], gain, [2], sampling-time [2]–[5], as well as bandwidth mismatches (from the sampling *RC* time constant) [2], [6]–[8]. These mismatches create distortion tones at frequency locations which are multiples of f_s/M, where $f_s = 1/T$ is the overall sampling frequency of the system and M the number of TI channels.

To characterize conveniently the performance degradation due to various types of mismatches, handy and closed-form expressions for the signal-to-noise-and-distortion ratio (SNDR) are highly desirable. Thus, the mismatches can be predicted and characterized in the very early design phase, allowing an accurate estimation of sampling circuit components, otherwise post-calibration (either analog or digital) [6]–[8] is unavoidable, which may lead to additional power and silicon consumption with complex hardware and software (algorithms) for measurement and calibration.

In this paper, a thorough spectra-domain analysis of the bandwidth mismatch effect will be presented and based on this a closed-form SNDR expression will be derived. Also, it will be proven that the most important contribution for the bandwidth mismatch is the phase mismatch originated from the phase-shift of the *RC*-filtering by the sampling branch, which causes performance degradation even for signal frequencies smaller than the bandwidth of such sampling branch. Although the analysis of deterministic mismatches had already been addressed in [2], [6] and [7], this paper will present a comprehensive and thorough study with a new statistical approach that comprises the following improved features over previous works.

a) A handy closed-form SNDR expression is derived.
b) The formula can be generalized for any number (M) of TI channels.
c) Since intrinsic circuit imbalance (being highly circuit and architecture dependent) or inadequate layout design will introduce systematic mismatches, and considering that only standard deviations of various mismatch parameters are given by the foundries process datasheets, a statistical method dealing with random mismatches is derived.

The formula derived here will simplify the performance evaluation of TI sampling systems under bandwidth mismatches, and will also allow the prior determination of the specifications of sampling switches and capacitors in the design phase for a specifically targeted value of SNDR. The analysis in [8] focuses on the design methodology for digital post calibration assuming small input signal bandwidth, while this paper is targeting a prior analysis to suppress the bandwidth mismatch within a full

Manuscript received February 6, 2007; revised December 22, 2007. First published May 7, 2008; last published July 16, 2008 (projected). This work was supported by the University of Macau under Research Grant RG058/06-07S/MR/FST. This paper was recommended by Associate Editor G. Manganaro.

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Digital Object Identifier 10.1109/TCSII.2008.921600

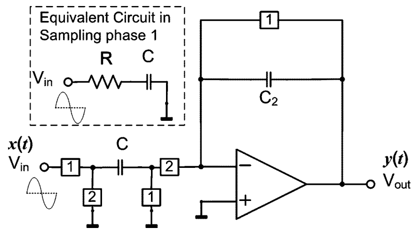


Fig. 2. Typical front-end S/H circuit.

input signal bandwidth at a much earlier design phase that is well suited for usual top-down analog design flows.

II. Spectra Analysis of Bandwidth Mismatches

A typical sample-and-hold (S/H) circuit widely used in the front-end of sampled-data systems is shown in Fig. 2, which must also be part of each channel of the ADC front-end of Fig. 1. In phase 1 the continuous-time input signal is passively sampled into the capacitor *C* and also *RC*-filtered by the finite bandwidth imposed by the sampling switch's on-resistance *R* and the sampling capacitor *C*. The mismatches occurring in the sampling switches and the capacitors among various channels will lead to the overall bandwidth mismatch. This will only modulate the continuous-time input signal, and after the signal being sampled into *C* the operation of the remaining part of the S/H is in discrete-time and all the mismatches will appear as offset and pure gain-mismatches that were already fully characterized before, [2]. Assuming that R_m and C_m are the switch's on-resistance and the sampling capacitance of the mth channel, respectively, then, the sampling time constant would be $\tau_m = R_m C_m = \tau(1+r_m)$ where τ corresponds to its mean/nominal value and r_m is the percentage of the time-constant mismatch. Based on this, the frequency response in the mth channel, from the input $x(t)$ to the sampled signal $y(t)$, can be defined as

$$H_m(\omega) = \frac{1}{1+j\omega\tau_m} = \frac{1}{1+j\omega\tau(1+r_m)} \tag{1}$$

containing both magnitude and phase contributions. On the other hand, the sampled time-domain signal in capacitor *C* can be represented as

$$y(t) = \sum_{n=-\infty}^{\infty} \{[x(nT) * h_m(nT)]\delta(t-nT)\} \tag{2}$$

where $h_m(t)$ is the impulse response of the *RC*-filter in the mth channel and $*$ denotes the convolution operation. For a TI system with M channels the following will hold:

$$nT = kMT + mT, \qquad \text{for integer } k \tag{3}$$

and (2) becomes

$$y(t) = \sum_{m=0}^{M-1}\sum_{k=-\infty}^{\infty}\Big\{[x(kMT+mT) * h_m \times (kMT+mT)]\delta(t-kMT-mT)\Big\}. \tag{4}$$

This leads to an output spectrum that can be expressed by [7]

$$Y(\omega) = \frac{1}{T}\sum_{k=-\infty}^{\infty} A_k(\omega) X\left(\omega - k\frac{2\pi}{MT}\right) \tag{5}$$

where

$$A_k(\omega) = \frac{1}{M}\sum_{m=0}^{M-1} H_m\left(\omega - k\frac{2\pi}{MT}\right) e^{-jkm\frac{2\pi}{M}}. \tag{5a}$$

Equations (5) and (5a) completely describe the spectrum of a TI sampling system with bandwidth mismatches. Furthermore, they reveal that the characteristic of the bandwidth mismatch is similar to a pure gain - and timing mismatch, also creating weighted modulation sidebands at multiples of $f_s/M = 1/MT$ [with weights specified by A_k in (5a)].

III. Closed-Form Formula for SNDR

In this section, a closed-form formula for SNDR will be derived for TI systems with bandwidth mismatches. Recalling that (5a) represents the weights of the various modulation sidebands which allows SNDR's computation. For an input sinusoidal signal with frequency $\omega_0 = 2\pi f_0$, the sidebands contain tones located at $\omega = \pm\omega_0 + k(2\pi)/(MT)$. Due to the symmetry property of the Fourier Transform of real signals, the calculation can be simplified by evaluating the sideband components at only $\omega = \omega_0 + k(2\pi)/(MT)$ over the range $[-f_s/2, f_s/2]$ (or equivalently $[0, f_s]$) to find the SNDR (with respect to the signal only centered at ω_0) over the range of $[0, f_s/2]$.

Simplifying (5a) using $\omega = \omega_0 + k(2\pi)/(MT)$ and (1) it yields

$$A_k\left(\omega_0 + k\frac{2\pi}{MT}\right) = \frac{1}{M}\sum_{m=0}^{M-1}\frac{e^{-jkm\frac{2\pi}{M}}}{1+j\omega_0\tau(1+r_m)} \tag{6}$$

where for $k=0$ $A_0(\omega_0)$ corresponds to the signal component, and for $k = 1$ to $M-1$ A_k correspond to $M-1$ different sideband components. The SNDR, over the range of $[0, f_s]$ or equivalently for $k=1$ to $M-1$, can be expressed as [9]

$$\text{SNDR} = 10\log_{10}\left[\frac{E[|A_0(\omega_o)|^2]}{E\left[\sum_k \left|A_k\left(\omega_o + k\frac{2\pi}{MT}\right)\right|^2\right]}\right] \text{dB}. \tag{7}$$

In order to evaluate the SNDR by statistical means it would be necessary to assume now $r_m (m = 0, 1, 2, \ldots, M-1)$ as M independent identically distributed (i.i.d) random variables with Gaussian distribution of zero mean and standard deviation σ_r. Then, the expected values of the signal component $|A_0(\omega_o)|^2$ can be evaluated by substituting $k=0$ into (6) and multiplying it by its complex conjugate as follows:

$$E[|A_0(\omega_0)|^2] = \frac{1}{M^2}E\left[\left|\sum_{m=0}^{M-1}\frac{1}{1+j\omega_0\tau(1+r_m)}\right|^2\right] \approx \frac{1}{1+\omega_0^2\tau^2} \tag{8}$$

which shows that the signal component is approximately scaled by the *RC*-filter for small values of r_m $(r_m \ll 1)$. The expected value of the sideband components can also be calculated from

(6) leading to (9), shown at the bottom of the page, where, in this case, the second-order term $r_m r_n$ in (9) cannot be neglected since in the following derivations it will be required to use the statistical values of $E[r_m] = 0$ and $E[r_m^2] = \sigma_r^2$ and, besides this, it will constitute the most significant contribution to the sideband components. However, in (8) such approximation is valid because the amplitude of the signal component is, in principle, much larger than that of the sideband. The evaluation of the expected value of the sideband components in (9) is quite complex and only key steps will be addressed here. The double summation in (9) can be analyzed in two parts for $m = n$ (containing M terms) and $m \neq n$ (contain $M^2 - M$ terms) in the following way:

$$\text{Part 1} = \frac{1}{M^2}\sum_{\substack{m=0\\m=n}}^{M-1}\sum_{n=0}^{M-1} E\left[\frac{1}{1+\omega_0^2\tau^2\left(1+2r_m+r_m^2\right)}\right]$$

$$\approx \frac{1}{M^2}\sum_{\substack{m=0\\m=n}}^{M-1}\sum_{n=0}^{M-1} \frac{1}{1+\omega_0^2\tau^2}E[1-A+A^2] \qquad (10)$$

$$\text{Part 2} = \frac{1}{M^2}\sum_{\substack{m=0\\m\neq n}}^{M-1}\sum_{n=0}^{M-1} e^{-jk(m-n)\frac{2\pi}{M}}$$

$$E\left[\frac{1}{1+j\omega_0\tau(r_m-r_n)+\omega_0^2\tau^2(1+r_m+r_n)}\right]$$

$$\approx \frac{1}{M^2}\sum_{\substack{m=0\\m\neq n}}^{M-1}\sum_{n=0}^{M-1} \frac{e^{-jk(m-n)\frac{2\pi}{M}}}{1+\omega_0^2\tau^2}E[1-B+B^2] \qquad (11)$$

where

$$A = \frac{\omega_0^2\tau^2\left(2r_m+r_m^2\right)}{1+\omega_0^2\tau^2}$$

$$B = \frac{\omega_0^2\tau^2(r_m+r_n)+j\omega_0\tau(r_m-r_n)}{1+\omega_0^2\tau^2}$$

should be small for $r_m \ll 1$ leading to a valid Taylor Series expansion in (10) and (11). The expansion can also be evaluated up to the second order because A^3 and B^3 contain at least a third-order term like r_m^3 which is negligible compared to the main contribution from r_m^2. The expected values can be calculated as

$$E[A] = \frac{\omega_0^2\tau^2}{1+\omega_0^2\tau^2}\left(2E[r_m]+E\left[r_m^2\right]\right)$$

$$= \frac{\omega_0^2\tau^2\sigma_r^2}{1+\omega_0^2\tau^2} \qquad (12)$$

$$E[B] = \frac{\omega_0^2\tau^2}{1+\omega_0^2\tau^2}(E[r_m]+E[r_n])$$

$$+\frac{j\omega_0\tau}{1+\omega_0^2\tau^2}(E[r_m]-E[r_n]) = 0 \qquad (13)$$

since r_m and r_n are random variables with zero means. Furthermore, for the second-order term

$$E[A^2] = \left(\frac{\omega_0^2\tau^2}{1+\omega_0^2\tau^2}\right)^2 \times\left(4E\left[r_m^2\right]+4E\left[r_m^3\right]+E\left[r_m^4\right]\right)$$

$$\approx \left(\frac{\omega_0^2\tau^2}{1+\omega_0^2\tau^2}\right)^2\left(4\sigma_r^2\right) \qquad (14)$$

$$E[B^2] = \left(\frac{\omega_0\tau}{1+\omega_0^2\tau^2}\right)^2 E\left[\omega_0^2\tau^2(r_m+r_n)^2 + 2j\omega_0\tau\left(r_m^2-r_n^2\right)-(r_m-r_n)^2\right]. \qquad (15)$$

Since r_m and r_n have zero mean and are independent (uncorrelated) (15) can be evaluated considering $E[r_m^2] = E[r_n^2] = \sigma_r^2$ and $E[r_m r_n] = 0$ which will lead to

$$E[B^2] = \left(\frac{\omega_0\tau}{1+\omega_0^2\tau^2}\right)^2\left[2\sigma_r^2\left(\omega_0^2\tau^2-1\right)\right]. \qquad (16)$$

Then, substituting (12) and (14) into (10) it yields

$$\text{Part 1} = \frac{1}{M}\frac{1}{1+\omega_0^2\tau^2} \times\left[1+\left(\frac{\omega_0\tau}{1+\omega_0^2\tau^2}\right)^2\sigma_r^2\left(3\omega_0^2\tau^2-1\right)\right] \qquad (17)$$

and, similarly, by substituting (13) and (16) into (11)

$$\text{Part 2} == \frac{1}{M^2}\frac{1}{1+\omega_0^2\tau^2} \times\left[1+\left(\frac{\omega_0\tau}{1+\omega_0^2\tau^2}\right)^2\left[2\sigma_r^2\left(\omega_0^2\tau^2-1\right)\right]\right] \times\sum_{\substack{m=0\\m\neq n}}^{M-1}\sum_{n=0}^{M-1} e^{-jk(m-n)\frac{2\pi}{M}} \qquad (18)$$

where

$$\sum_{\substack{m=0\\m\neq n}}^{M-1}\sum_{n=0}^{M-1} e^{-jk(m-n)\frac{2\pi}{M}} = \sum_{m=0}^{M-1}\sum_{n=0}^{M-1} e^{-jk(m-n)\frac{2\pi}{M}} - \sum_{\substack{m=0\\m=n}}^{M-1}\sum_{n=0}^{M-1} e^{-jk(m-n)\frac{2\pi}{M}} = 0 - M \qquad (19)$$

leading to the following simplification of part 2:

$$\text{Part 2} = -\frac{1}{M}\frac{1}{1+\omega_0^2\tau^2}\left[1+\left(\frac{\omega_0\tau}{1+\omega_0^2\tau^2}\right)^2\left[2\sigma_r^2(\omega_0^2\tau^2-1)\right]\right]. \qquad (20)$$

$$E\left[\left|A_k\left(\omega_0+k\frac{2\pi}{MT}\right)\right|^2\right] = \frac{1}{M^2}\sum_{m=0}^{M-1}\sum_{n=0}^{M-1} e^{-jk(m-n)\frac{2\pi}{M}} E\left[\frac{1}{1+j\omega_0\tau(r_m-r_n)+\omega_0^2\tau^2(1+r_m+r_n+r_m r_n)}\right] \qquad (9)$$

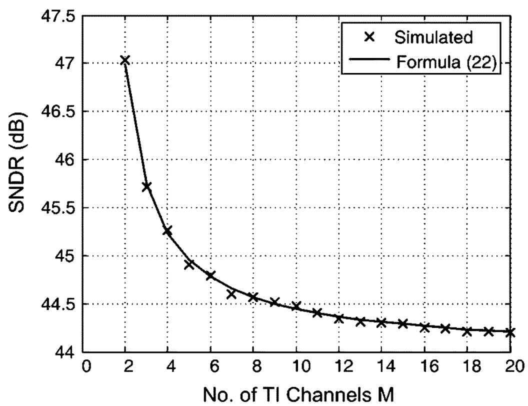


Fig. 3. Plot of number of TI channels M versus SNDR [simulated and calculated from (22)] with $\omega_o\tau = 0.8$ and $\sigma_r = 1\%$.

Finally, the sideband components can be calculated by the addition of both parts 1 and 2

$$E\left[\left|A_k\left(\omega_0 + k\frac{2\pi}{MT}\right)\right|^2\right] = \frac{1}{M}\frac{1}{1+\omega_0^2\tau^2}\frac{\omega_0^2\tau^2\sigma_r^2}{1+\omega_0^2\tau^2}. \quad (21)$$

From (21) it can be concluded that the expected values of the various sidebands are independent of the frequency index k, i.e., all the sidebands have the same expected distortion power. As a result, the total distortion power should be equal to (21) multiplied by $M-1$, and the final SNDR can be calculated by substituting (8) and (21) into (7)

$$\mathrm{SNDR} = 10\log_{10}\left[1+\omega_0^2\tau^2\right] - 20\log_{10}[\omega_0\tau\sigma_r] - 10\log_{10}\left[1-\frac{1}{M}\right]\mathrm{dB}. \quad (22)$$

Although the previous analysis can be considered relatively complex, the final SNDR formula is simple enough for quick evaluation of the performance of TI sampled-data systems under the influence of bandwidth mismatches. Important information can be extracted directly from (22), namely that SNDR is inversely proportional to the signal frequency, as expected. The 1st term in (22) is reflecting the *RC* filtering effect by the finite bandwidth, since this will degrade the signal amplitude and thus the SNDR. Also, the formula can be generalized for any number (M) of TI-channels, and actually the SNDR only degrades by an amount of 3 dB from $M = 2$ to ∞, as shown in Fig. 3, a fact that was not explained before when the phenomenon of bandwidth mismatch was previously analyzed [2], [6], [7] (actually a similar conclusion can be found in the gain- and timing-mismatch analysis, e.g., from, [3], and [5]).

The first-order filtering effect caused by the sampling branch will usually have more influence when the signal frequency is near or greater than its bandwidth, since before the -3-dB frequency the filter provides only negligible attenuation. However, the bandwidth mismatches among TI channels can have a large impact on the performance even when the signal frequency is well below the corner frequency (quite unexpectedly), since the phase shift starts to increase at $1/10$ of the corner-frequency. To demonstrate this effect it should be considered only the magnitude response of an idealized *RC*-filter

$$H_{m\_\mathrm{magnitude\_only}}(\omega) = \frac{1}{\sqrt{1+[\omega\tau(1+r_m)]^2}}. \quad (23)$$

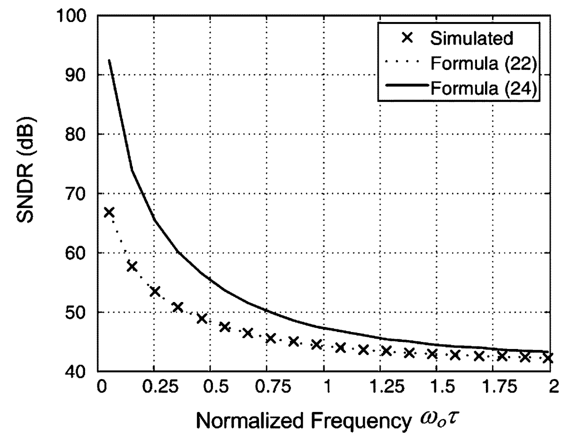


Fig. 4. Plot of normalized frequency $\omega_o\tau$ versus SNDR [simulated and calculated from (22) and (24)] with $M = 4$ and $\sigma_r = 1\%$.

Then, following the same procedure it can be shown that :

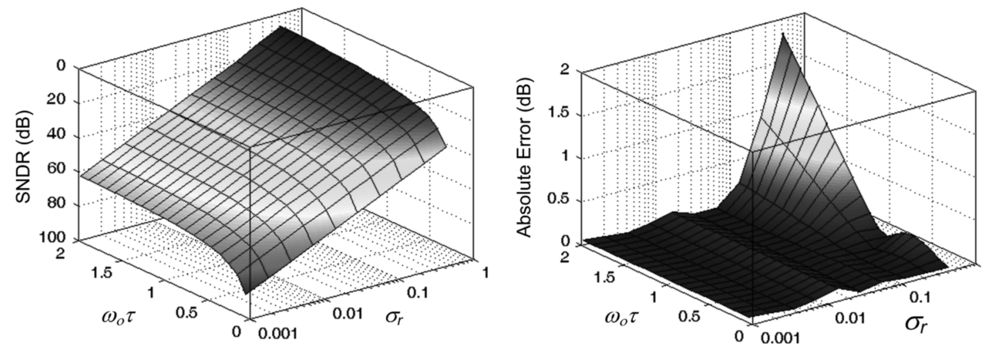
$$\mathrm{SNDR}_{\mathrm{magnitude\_only}} = 10\log_{10}\left[1+\omega_0^2\tau^2\right] - 20\log_{10}[\omega_0\tau\sigma_r] - 10\log_{10}\left[1-\frac{1}{M}\right] - 10\log_{10}\left[\frac{\omega_0^2\tau^2}{1+\omega_0^2\tau^2}\right]\mathrm{dB} \quad (24)$$

which differs from (22) only in the last term. Fig. 4 presents a plot of (22) and (24) as a function of $\omega_0\tau$ with $M = 4$ and $\sigma_r = 1\%$. It clearly shows that the effect of the phase mismatch is dominant especially at low frequency, which will lead to (e.g., 25 dB) overestimation of SNDR for the TI systems if only gain mismatch is considered as a dominating factor around the corner frequency.

IV. Simulation Results and Design Flow

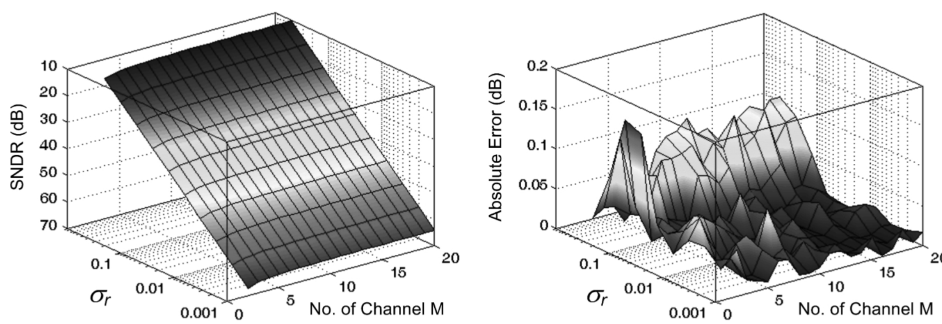
To verify the accuracy of the formula, MATLAB FFT model simulations will be presented. Fig. 5(a) shows plots of the simulated SNDR versus normalized signal frequency $\omega_0\tau$ and σ_r by 10000 times Monte Carlo simulations under $M = 4$ and $\omega_0 T = 2\pi \times 0.2495$, with the absolute error (in decibels) between the simulated and calculated SNDR presented in Fig. 5(b). From this figure the absolute error between the simulated and calculated SNDR is well below 0.5 dB for the normal range of $\omega_0\tau$ and σ_r. The error starts to increase with higher values of the mismatch σ_r and high input frequency, since an approximation of $\sigma_r \ll 1$ is considered in the previous formula derivations, but such a large value of mismatches (e.g., >10%) are not typical in CMOS which will lead to an SNDR of only <20 dB. Fig. 6(a) shows another 3-D plots of the simulated SNDR versus the number of TI channels M and σ_r by 10000 times Monte Carlo simulations under $\omega_o\tau = 0.8$ and $\omega_0 T = 2\pi \times 0.2495$, with the absolute error (in decibels) between the simulated and calculated SNDR presented in Fig. 6(b). It is also demonstrated that the formula can accurately predict systems' performance under any number of TI channels M.

On the other hand, to demonstrate the corresponding design flow using (22), a 10-bit 6-channel 360 MHz pipelined ADC was implemented in 0.18-μm CMOS [10], where the SNDR specification for distortions imposed by signal-dependent modulation sidebands was designed to be 60 dB (including all error sources), and the SNDR contribution from pure bandwidth mismatches should be below 65 dB. With the signal frequency of



(a) (b)

Fig. 5. (a) Simulated SNDR. (b) Absolute error between the simulated and calculated SNDR of TI systems under bandwidth mismatches versus normalized frequency $\omega_o\tau$ and mismatch standard derivation σ_r by 10000 times Monte Carlo Simulations $(M = 4, \omega_o T = 2\pi \times 0.2495)$.



(a) (b)

Fig. 6. (a) Simulated SNDR. (b) Absolute error between the simulated and calculated SNDR of TI systems under bandwidth mismatches versus no. of TI channels M and mismatch standard derivation σ_r by 10000 times Monte Carlo Simulations $(\omega_o\tau = 0.8, \omega_o T = 2\pi \times 0.2495)$.

25.2 MHz and sampling bandwidth of 100 MHz, the mismatch in the sampling bandwidth should be smaller than $\sigma_r < 0.3\%$ from (22), which is achievable in current standard CMOS technology. According to this specification and the statistical parameters from the foundry's datasheet (with design margins), 1-pF sampling capacitor and 1.6-kΩ resistance were chosen which will imply $\tau = 1.6$ ns. This chip was successfully implemented with good silicon performance, and without any type of bandwidth mismatch calibration, showing the importance of minimizing the bandwidth mismatches according to the formula and following an analog design flow leading to precise values of the sampling circuit components.

V. Conclusion

This paper presents a comprehensive spectrum analysis of the bandwidth mismatch in time-interleaved sampling systems. Due to the statistical nature of mismatches a statistical analysis is presented together with the derivation of a handy and closed-form SNDR formula which can be used to quickly evaluate the system performance. The analysis was further generalized to any number of TI channels with an elegant and simple expression. To verify the accuracy of the formula MATLAB simulations were presented showing that the absolute error in performance prediction was as low as 0.5 dB only for a reasonable range of bandwidth mismatches and signal frequencies. An IC design example is also briefly referred to show the correctness of following an analog design procedure (based on the formula), in the dimensioning of the sampling network components, thus leading to the suppression of such mismatches.

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