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APPLICATION NOTE

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HMC661LC4B and HMC760LC4B THA Noise Characteristics and Analysis

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INTRODUCTION

There are two key noise components to the track-and-hold (THA) output noise: sampled noise and output buffer amplifier noise. This application note explores both components.

SAMPLED NOISE COMPONENT

The first component of noise is the sampled noise produced by the sampling process, which heterodynes the front-end noise of the THA into every Nyquist interval of the frequency domain. The noise resulting from the entire front-end bandwidth is captured in each time domain sample. This noise is then distributed across each Nyquist interval approximately uniformly. This noise, which consists of both front-end thermal noise and sample jitter noise, cannot be filtered out unless a low-pass filter corner frequency is used at the output that reduces the Nyquist bandwidth significantly. This filtering is not usually implemented because it defeats the available bandwidth provided by the clock rate and causes degraded settling times in the output waveform.

OUTPUT BUFFER AMPLIFIER NOISE COMPONENT

The second component of noise is the THA output buffer amplifier noise contribution. The THA does not sample this noise; however, filtering reduces the noise. The amount of output filtering that can be tolerated is determined by the settling time requirements for the particular clock rate used. An approximate guideline for tolerance limits is that the bandwidth of the output path (including the analog-to-digital converter (ADC) input bandwidth) is at least 2× the clock rate to support accurate (for example, linear) settling of the THA waveform as sampled by the downstream ADC. The input bandwidth of a high speed ADC is usually in the vicinity of the 2× clock rate guideline; therefore, additional filtering is not usually necessary when working with high speed ADCs.

NOISE DENSITY FOR SAMPLED AMPLIFIER

The effective equivalent input noise spectral density in the frequency domain for the THA depends on the bandwidth of the output filtering prior to any analog-to-digital conversion, unlike a conventional nonsampled amplifier. For this reason, sampled devices often do not specify noise in these terms because the actual output noise is a complex function of the sampled input buffer noise, which aliases the noise throughout the entire input bandwidth down into the first Nyquist interval and the output buffer amplifier noise, which responds to output bandlimiting similar to a conventional amplifier. For a sampled system, the important quantity is the time domain output noise in the held output samples (see the HMC661LC4B data sheet) because the ADC converts this output noise.

The best definition of the input referred, frequency-domain noise density is determined by dividing the output time domain sample noise by the square root of the input sampling bandwidth × π/2.

The result of this definition gives the same input referred noise density as a unity-gain continuous wave (CW) amplifier (not sampled) with the same single-pole bandwidth and output time domain noise. The pi divided by 2 (π/2) is derived because the effective noise bandwidth of a single-pole, low-pass transfer function is BW3dB × π/2. For the HMC661LC4B with no output bandlimiting (for example, a full 7 GHz bandwidth of an output buffer amplifier), this noise bandwidth corresponds to an equivalent input noise density of about 6.2 nV/(√Hz) when using a 1.05 mV rms time domain sample noise and an 18 GHz, 3 dB input bandwidth. Because the thermal noise floor is at 0.64 nV/(√Hz), the effective noise figure is about 19.7 dB. This noise figure is high because there are several stages in a THA, all of which are operating at unity gain; consequently, each stage adds noise.

NOISE DENSITY FOR NONSAMPLED AMPLIFIER

This definition of an effective sampler noise figure is a reasonable comparison to a normal nonsampled amplifier in terms of equivalent input noise performance. This does not take into account the sampling induced noise folding that might be used for a typical mixer noise figure definition. To arrive at the mixer noise figure definition, add a noise folding correction factor, given by the ratio of the input sampling noise bandwidth and the Nyquist bandwidth, as shown in the following equation:

$$NF_{CORRECTION} = \textit{Noise Figure Sampling Folding Correction} = 10\log(BW_{N\_INPUT}/(f_{CLK\_TH}/2))$$

where BW_{N_INPUT} is the effective noise bandwidth of the input sampling bandwidth.

For example, when the HMC661LC4B operates at a 4 GHz clock rate, the additional degradation due to folding (18 GHz × π/2) of noise into one Nyquist interval of 2 GHz is a mixer defined, additional noise figure degradation of about 11.5 dB, producing an overall mixer defined noise figure of 19.7 dB + 11.5 dB = 31.2 dB.

TABLE OF CONTENTS

REVISION HISTORY

2/2018—Revision 0: Initial Version

ESTIMATING OUTPUT NOISE SPECTRUM

For an estimate of the output noise spectrum, use the fact that all the front-end noise is heterodyned or folded into one Nyquist interval, while the output buffer noise is spread over approximately 7 × π/2 GHz of the output buffer noise bandwidth. Simulations indicate that the effective noise bandwidth of the combined output buffer amplifier stages in both the HMC661LC4B and the HMC760LC4B is about 12.6 GHz, corresponding to an effective −3 dB noise density bandwidth of 8 GHz, despite the 7 GHz small signal output buffer bandwidth. This slight discrepancy appears to result from the distributed contributions of the noise at various bandwidth points in the signal chain. Table 1 and Table 2 show the breakdown for output time domain and frequency domain noise components for the HMC661LC4B and the HMC760LC4B, operating at a 1 GHz clock rate, for various output noise bandwidth filtering scenarios.

DERIVING FULL OUTPUT BANDWIDTH

The full output bandwidth data is derived from detailed chip simulations with full wiring parasitics; however, the results are in fairly good agreement with lab data as well (for the HMC661LC4B, the measured integrated noise voltage (V_{NT}) = 1.05 mV rms). The reduced output bandwidth scenarios are calculated assuming constant output buffer noise spectral density because the filtering is typically external. Table 1 and Table 2 simulate V_{NT_SAMPLE}, V_{NT_OUT}, and V_{NF_OUT} for the full output bandwidth scenario. These fundamental parameters directly derive all other quantities.

In the process of sampling, the downstream ADC heterodynes all of the THA noise over its input bandwidth (a combination of filtering and ADC input bandwidth as shown in Table 1 and Table 2) into one ADC Nyquist interval. Therefore, the total folded noise accounts for the total THA output amplifier time domain noise over the ADC noise bandwidth.

For reference, the data using the HMC661LC4B to drive the National Semiconductor ADC12D1600 ADC shows reasonable agreement with the simulated values in Table 1 and Table 2. In particular, a THA noise component of about 37 nV/(√Hz) is measured for the spectral density in the ADC fast fourier transient (FFT). The input noise bandwidth of the National Semiconductor converter is estimated to be about 2.8(π/2) = 4.4 GHz. For this scenario, the total THA output time domain noise is about 0.98 mV rms corresponding to a noise spectral density (after sampling by the ADC) of 43.9 nV/√Hz. This value is within 1.5 dB of the measured THA component in the ADC digitized noise spectrum.

Table 1. Summary of HMC661LC4B 18 GHz Bandwidth THA Simulated and Calculated Noise Components for 1 GHz Clock Frequency

Output Noise Bandwidth (GHz)[1]	Sampled Noise V_{NT_SAMPLE} (mV)	Output Buffer Noise V_{NT_OUT} (mV)	Total Output Noise V_{NT} (mV)	Sample Noise Spectral Density V_{NF_SAMPLE} (nV/√Hz)	Output Buffer Noise Spectral Density V_{NF_OUT} (nV/√Hz)	Total Output Noise Spectral Density at Baseband V_{NF} (nV/√Hz)	Total Folded THA Noise after ADC Sampling $V_{NF\_TH\_ADC}$ (nV/√Hz)
Full (~12.6)	0.92[2]	0.614[2]	1.106	41.1	5.46[2]	41.5	49.5
3	0.92	0.3[2]	0.97	41.1	5.46[2]	41.5	43.4
1	0.92	0.17[2]	0.94	41.1	5.46[2]	41.5	42

[1] For any filtering and the ADC bandwidth.
[2] Simulated Values, all other values are calculated from simulated values.

Table 2. Summary of HMC760LC4B 5.5 GHz Bandwidth THA Simulated and Calculated Noise Components for 1 GHz Clock Frequency

Output Noise Bandwidth (GHz)[1]	Sampled Noise V_{NT_SAMPLE} (mV)	Output Buffer Noise V_{NT_OUT} (mV)	Total Output Noise V_{NT} (mV)	Sample Noise Spectral Density V_{NF_SAMPLE} (nV/√Hz)	Output Buffer Noise Spectral Density V_{NF_OUT} (nV/√Hz)	Total Output Noise Spectral Density at Baseband V_{NF} (nV/√Hz)	Total Folded THA Noise after ADC Sampling $V_{NF\_TH\_ADC}$ (nV/√Hz)
Full (~12.6)	0.67[2]	0.61[2]	0.90	30	5.46[2]	30.5	40.5
3	0.67	0.3[2]	0.73	30	5.46[2]	30.5	32.6
1	0.67	0.17[2]	0.69	30	5.46[2]	30.5	30.9

[1] For any filtering and the ADC bandwidth.
[2] Simulated values, all other values are calculated from the simulated values.

CONCLUSION

The user can estimate the output noise spectral density of the THA by spreading the sample time domain noise over one Nyquist bandwidth and by filtering the output buffer noise spectral density over the effective noise detection bandwidth of the downstream ADC. Therefore, the following estimations must be obtained:

$$V_{NF\_SAMPLE}(f) = V_{NT\_SAMPLE}/(f_{CLK}/2)^{1/2}$$

$$V_{NF\_OUTPUT}(f) = 5.46 \text{ nV}/\sqrt{\text{Hz}} \ (\textit{Out to 7 GHz Bandwidth})$$

$$V_{NF} = [(V_{NF\_SAMPLE})2 + (V_{NF\_OUTPUT})2]^{1/2}$$

$$V_{NF\_TH\_ADC} = V_{NT}/(f_{CLK}/2)^{1/2}$$

where:
V_{NT} and V_{NT_x} are the time domain noise quantities.
V_{NF} and V_{NF_x} are the frequency domain spectral densities.

This calculation assumes that the spectral content of only the hold mode portion of the output waveform is measured. If an ADC samples the THA waveform at the same clock rate, the total time domain noise occurring over the input bandwidth of the ADC (including any additional output filtering of the THA output) is spread over one ADC Nyquist interval. In principle, these computations can be performed for any arbitrary clock frequency. It is clear that the THA sample noise dominates; therefore, the impact and benefit of output filtering is limited.

At higher signal frequencies, the jitter of the clock and the signal create an additional noise component to sample noise. At these higher clock frequencies, the jitter noise is not negligible and must be included in the total noise. Jitter noise is usually quantified by quoting the jitter specification in the data sheet because the noise created by jitter is easily calculated and depends on the input frequency and value of the jitter. In general, the rms value of the noise created by the jitter during the sampling process is approximately

$$V_{NT\_JITTER} \sim SR \times tj$$

where:
SR is the signal slew rate at the sample point.
tj is the rms jitter.

Calculate the peak value of the slew rate (SR) for a sinusoidal signal by

$$V_{IN} \times 2\pi \times f_{SIGNAL}$$

where:
V_{IN} is the zero-to-peak signal level.
f_{SIGNAL} is the signal frequency.

After statistical averaging, the effective slew rate to use for this computation is based on the rms value of V_{IN}, the effective slew rate ($SR_{EFFECTIVE}$) = $(V_{IN}/2^{1/2}) \times 2\pi \times f_{SIGNAL}$. Therefore, the total jitter noise (in the time domain samples) is

$$V_{NT\_JITTER} = SR_{EFFECTIVE} \times tj = (V_{IN}/2^{1/2}) \times 2\pi \times f_{SIGNAL} \times tj$$

This unavoidable component of noise increases linearly with frequency. Therefore, the jitter limited signal-to-noise ratio (SNR) is

$$SNR_{JITTER} \sim -20\log[1/(2\pi \times f_{SIGNAL} \times tj)].$$

To compute the total noise at a given frequency, add the jitter noise power to the thermal noise power. The jitter in the HMC661LC4B THA is specified at <70 fs in the HMC661LC4B data sheet from specialized jitter measurements on the THA alone. Typical values measured in THA and ADC assembly measurements are in close agreement with the measurement of the THA alone and of the order of 65 fs. This noise tends to have a relatively flat spectrum in a given Nyquist sampling interval. To average this noise down in level, use multiple independent data records. To achieve this level of total subsystem jitter, good signal and clock generators must be used and phase-locked to each other, and the outputs of the signal and clock generators must be filtered to remove nonharmonic spurious signals.

Even state-of-the-art, low phase noise, synthesized signal generators can contribute significant jitter to a sampling system that incorporates the HMC661LC4B, particularly when the phase locking jitter between the signal and clock generators is incorporated. Observe the impact of the generator noise induced jitter by band-pass filtering the generator output signal that is applied to the THA. In this case, phase noise sidebands corresponding to the band-pass filter bandwidth are observed on the THA output signals and any ADC FFT-processed output spectrums. Obtain optimal performance by using small filter bandwidths to remove broadband noise from the generator. In addition, a reasonable clock slew rate must be maintained. Use the 2 V/ns to 4 V/ns for each clock differential half circuit input to achieve the jitter as detailed in the HMC661LC4B data sheet. If the THA is used in front of an ADC, the THA sets the jitter, and the jitter of the ADC is essentially negligible because it is sampling a stable held waveform from the THA. The jitter noise can also be processed to improve the SNR by multiple record averaging or spread spectrum processing techniques. The THA jitter noise component is also spread over one Nyquist interval because it tends to be broadband. Therefore, the jitter spectral noise density is

$$V_{NF\_JITTER} \sim V_{NT\_JITTER}/(f_{CLK}/2)^{1/2}$$

The three noise contributions of sample thermal noise, sample jitter noise, and output buffer noise are uncorrelated and their powers add linearly.

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