Broadband Sampling Oscilloscope Characterization with the 'Nose-to-Nose' Calibration Procedure: a Theoretical and Practical Analysis

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Broadband Sampling Oscilloscope Characterization with the “Nose-to-Nose” Calibration Procedure: 
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Abstract—In the past the “nose-to-nose” calibration procedure has been introduced as probably the most accurate method to determine the impulse response of broadband sampling oscilloscopes like the HP54124T. The method is based on the hypothesis “sampler kick-out equals oscilloscope impulse response.” This hypothesis was originally based on an intuitive approach and was later verified experimentally (comparing with power measurements) as well as with SPICE simulations. Until now, however, there was no generalized mathematical evidence supporting this basic hypothesis. In this paper a mathematical theory is developed, which starts from a generalized sampler equivalent scheme, and which shows that, under conditions which are valid in practice, the sampler kick-out indeed equals the sampler impulse response. Experimental results are reported concerning the accuracy and precision of the calibration procedure. These experiments involve the investigation of experiment repeatability, noise, sampler linearity and timebase effects.

I. INTRODUCTION

Many problems arise when one wants to calibrate broadband sampling oscilloscopes. With the present technology sampling oscilloscopes are available with bandwidth exceeding 50 GHz, as for example the Hewlett-Packard HP54124T. Several articles were published in the past concerning the characterization of such instruments, more specifically concerning the determination of the impulse response [1]–[5]. Probably the most accurate and practically realizable method to determine the impulse response of a broadband sampling oscilloscope is the “nose-to-nose” calibration procedure [3], [5]. This method is based on the fact that a sampling oscilloscope sampling direct current (dc) acts as a pulse generator. This pulse, named “kick-out” and launched from the sampler head towards the input connector, contains information on the oscilloscope’s linear characteristic. By digitizing this pulse with a second identical oscilloscope and using deconvolution techniques, the linear characteristic of the oscilloscope can be extracted from the waveform of the pulse finally appearing on the screen of the pulse-receiving oscilloscope. By the use of three oscilloscopes the assumption that the two oscilloscopes are identical can be omitted [5]. The basic principle of the method can be formulated as “sampler kick-out equals oscilloscope impulse response.” At first intuitively formulated by Ken Rush, the validity of the principle was proven by comparison with swept-sine amplitude measurements [3] and by SPICE [6] simulations. A more general mathematical proof was based on a simplified model of the sampler [5]. In this simplified model, the sampler is replaced by a time-varying conductance.

In this article the validity of this simplified model will theoretically be proven, based on an extended large-signal model. The corresponding theoretical analysis describes the effects of sampler asymmetry, strobe pulse feedthrough, nonperfect matching of the input circuitry and relates the shape of the time varying conductance of the simplified model to the sampling diode characteristics and the strobe pulse shape.

Experimental results are finally given concerning experiment repeatability, sampler linearity, noise and timebase effects.

II. MATHEMATICAL MODEL OF THE “NOSE-TO-NOSE” CALIBRATION PROCEDURE

Mathematical Notations

\[ f * g \] convolution of the functions \( f \) and \( g \).

\[ G' \] derivative of the function \( G \).

\[ \delta(x) \] Dirac-delta distribution.

\[ U(x) \] unit step function, equal to 0 if \( x < 0 \) and equal to 1 if \( x \geq 0 \).

Generalized Sampler Equivalent Scheme

The general sampler model on which all calculations are based, is depicted in Fig. 1. Filled arrow heads indicate voltage definitions, and unfilled arrow heads indicate current definitions. The source with an output impedance of 50 \( \Omega \) models the signal source at the input connector of the sampler; the general scattering parameter network models all physical structures between input connector and the sampling diodes, including the connection towards the terminating 50 \( \Omega \) resistor. The incident and scattered voltage waves are indicated by \( a_1(t) \), \( b_1(t) \), \( a_2(t) \), and \( b_2(t) \). In practice the output of the signal source will be limited to a magnitude such that no significant nonlinear distortions of the sampled signal are introduced. The sampling diodes are modelled by nonlinear conductances, with a current-voltage relationship given by the functions \( G_1 \) and \( G_2 \). Before the actual sampling takes place, both hold capacitors \( C_1 \) and \( C_2 \) are charged such that the sampling diodes are reverse biased. This is expressed in (1)
and (2), where $v_{B1}$ and $v_{B2}$ stand for the reverse bias voltages:

\[ v_{C1}(-\infty) = v_{B1} > 0 \]
\[ v_{C2}(-\infty) = v_{B2} > 0. \]

The sampling circuitry is designed to be as symmetric as possible. Sampling is performed by applying a strobe pulse, modelled by the voltage sources $v_{S1}(t)$ and $v_{S2}(t)$, to the circuit. This strobe pulse has a large negative amplitude which turns the inverse bias voltage into a forward bias condition during a small time span (small compared to all other time constants involved). During this time span, the sampling diodes are forward biased, and a charge will be transferred from the input circuitry to the hold capacitors. This charge is detected by a charge amplifier and will be proportional to the value of the input signal at the sampling instant. Note that, if the sampling circuit is symmetric, no charge will be transferred from the strobe pulse towards the hold capacitors.

In the following sections the mathematical expressions will be derived describing the effects mentioned above.

**Derivation of the “Sampler Drive” Equivalent Scheme**

First some basic network equations will be derived, describing the large-signal behavior of the sampling diodes. The equations are given by

\[ i_{D1}(t) = G_1(v_{D1}(t)) \]
\[ i_{D2}(t) = G_2(v_{D2}(t)) \]
\[ v_{D1}(t) + v_{C1}(t) + v_{S1}(t) = -v_2(t) \]
\[ v_{D2}(t) + v_{C2}(t) + v_{S2}(t) = v_2(t) \]
\[ v_{C1}(t) = \frac{1}{C_1} \int_{-\infty}^{t} i_{D1}(u) \, du + v_{B1} \]
\[ v_{C2}(t) = \frac{1}{C_2} \int_{-\infty}^{t} i_{D2}(u) \, du + v_{B2}. \]

In what follows, four new functions and two new constants are introduced, with subscripts $A$ (“average”) and $\Delta$ (“delta”) replacing the subscripts “1” and “2” for $v_{S1}(t), v_{S2}(t), v_{D1}(t), v_{D2}(t), v_{B1}$ and $v_{B2}$. The new functions and constants are defined as the average and half of the difference of the original functions and constants. The idea is illustrated for $v_{D_A}(t)$ and $v_{D_\Delta}(t)$ by (9) and (10), and has to be applied to all functions and constants mentioned:

\[ v_{D_A}(t) = \frac{v_{D1}(t) + v_{D2}(t)}{2} \]
\[ v_{D_\Delta}(t) = \frac{v_{D1}(t) - v_{D2}(t)}{2}. \]

Summing (5) and (6), substituting (3) and (4) and using the new functions results in

\[ v_{D_A}(t) + \int_{-\infty}^{t} \left( \frac{G_1(v_{D_A}(u)) + v_{D_\Delta}(u)}{2C_1} \right) \, du \]
\[ + \frac{G_2(v_{D_A}(u) - v_{D_\Delta}(u))}{2C_2} \, du + v_{B_A} + v_{S_A}(t) = 0. \]

In practice $v_{D_\Delta}(t)$ will be much smaller than $v_{D_A}(t)$. This is the case because the sampler is, by design, very symmetric and because the input signal is limited to a level where the circuit is behaving as a linear sampler. It will then be possible to expand $G_1$ and $G_2$ into a Taylor series around $v_{D_A}(t)$, and to neglect second and higher order terms of $v_{D_\Delta}(t)$ without causing any significant error. There is, in fact, an equivalence between the conditions which allow neglecting these higher order terms and the conditions needed to insure that the circuit acts as a linear sampler. This results in

\[ \int_{-\infty}^{t} \left( \frac{C_2 G_1(v_{D_A}(u)) + C_1 G_2(v_{D_A}(u))}{C_1 + C_2} \right) \, du \]
\[ + v_{D_A}(t) + v_{B_A} \]
\[ + v_{S_A}(t) + \int_{-\infty}^{t} \left( \frac{G_1'(v_{D_A}(u)) - G_2'(v_{D_A}(u))}{2C_1} \right) \, du = 0. \]

In (12) the last term of the left-hand side can be neglected because of two reasons: $v_{D_A}(t)$ is small and the circuit is practically symmetric, such that the integrand is the product of two small quantities, making it a second-order effect. When this term is neglected, the equation can be associated with the equivalent scheme of Fig. 2, which is called the “sampler drive” equivalent scheme. The diode $D$ has the following voltage-current characteristic:

\[ i = \frac{C_2 G_1(v) + C_1 G_2(v)}{C_1 + C_2}. \]

Note the definition of $C_H$:

\[ C_H = \left( \frac{1}{2} \left( \frac{1}{C_1} + \frac{1}{C_2} \right) \right)^{-1} \]

and that the initial charge on the capacitor $C_H$ equals 0. The solution of this electrical network results in $v_{D_A}(t)$. 
Derivation of the “Signal Sampling” Equivalent Scheme

In what follows, an equivalent scheme will be derived, describing the input signal sampling process and the kick-out generation. A proof of the validity of the basic nose-to-noise principle “kick-out equals impulse response” [3] results.

To simplify notations, two time-variant conductances are defined by

\[
g_A(t) = \frac{G_1'(v_{DA}(t)) + G_2'(v_{DA}(t))}{2},
\]

\[
g_\Delta(t) = \frac{G_1'(v_{DA}(t)) - G_2'(v_{DA}(t))}{2}.
\]

Subtracting (6) from (5), dividing by 2, expanding \(G_1\) and \(G_2\) in a first-order Taylor series around \(v_{DA}(t)\) results in

\[
v_{DA}(t) + \int_{-\infty}^{t} \left( \frac{1}{C_1} + \frac{1}{C_2} \right) \frac{g_A(u)}{2} v_{DA}(u) du
\]

\[+ \int_{-\infty}^{t} \left( \frac{1}{C_1} - \frac{1}{C_2} \right) \frac{g_\Delta(u)}{2} v_{DA}(u) du\]

\[= -v_2(t) - s_{SA}(t) - v_{BA}\]

\[+ \int_{-\infty}^{t} \left( \frac{G_1(v_{DA}(u))}{2C_1} - \frac{G_2(v_{DA}(u))}{2C_2} \right) du.
\] (17)

Note that the third term of the left-hand side of (17) is the product of two small terms because of symmetry, and may be neglected compared to the second term. Next, the Thévenin equivalent of the input circuitry will be calculated, including the 50 \(\Omega\) termination. As can be easily verified, the resulting equation is

\[
v_2(t) = s_{21}(t) * a_1(t) - 25(\delta(t) + s_{22}(t)) * i_3(t).
\] (18)

In this equation \(s_{21}(t) * a_1(t)\) is the open-circuit voltage, and \(25(\delta(t) + s_{22}(t))\) is the time-domain representation of the input circuit output impedance. For ease of notation this will be called \(z(t)\) in what follows. Next, \(i_3(t)\) is written as a function of \(v_{DA}(t)\) and \(v_{DA}(t)\):

\[
i_3(t) = G_2(v_{DA}(t) - v_{DA}(t)) - G_1(v_{DA}(t) + v_{DA}(t)).
\] (19)

Taking into account the same arguments as above, \(G_1\) and \(G_2\) are very well approximated by a first-order Taylor series, resulting in

\[
i_3(t) = 2(i_{AS}(t) - G_A(t)v_{DA}(t)),
\] (20)

with

\[
i_{AS}(t) = \frac{G_2(v_{DA}(t)) - G_1(v_{DA}(t))}{2}.
\] (21)

If (20) is substituted in (18), and this into (17) with the third term of the left-hand side eliminated, an equation in the unknown \(v_{DA}(t)\) results in

\[
v_{DA}(t) + \int_{-\infty}^{t} \frac{g_A(u)}{C_H} v_{DA}(u) du + 2z(t) * (g_A(t)v_{DA}(t))
\]

\[= v_{AS}(t) - s_{21}(t) * a_1(t),
\] (22)

with \(v_{AS}(t)\) independent from the input signal \(a_1(t)\) and \(v_{DA}(t)\) and defined as follows:

\[
v_{AS}(t) = -v_{SA}(t) - v_{BA}\]

\[+ \int_{-\infty}^{t} \left( \frac{G_1(v_{DA}(u))}{2C_1} - \frac{G_2(v_{DA}(u))}{2C_2} \right) du
\]

\[+ 2z(t) * i_{AS}(t).
\] (23)

The equivalent scheme corresponding to (22) and (20) is illustrated in Fig. 3. The initial charge on the capacitor is equal to zero. Note that \(v_{AS}(t)\) and \(i_{AS}(t)\) are equal to zero for a perfect symmetric sampler with a perfect symmetric strobe pulse. As such, \(v_{AS}(t)\) and \(i_{AS}(t)\) are the mathematical representation of the effects of the asymmetry of the sampler.

The Effect of Sampler Circuitry Asymmetry

The description of the sampling process and the creation of a kick-out when sampling a dc-voltage is based on the equivalent scheme of Fig. 3 and the scattering-parameter description of the input circuitry. The pulse coming out of the input connector when sampling, will equal \(b_1(t)\). It can be easily verified that it is given by

\[
b_1(t) = s_{11}(t) * a_1(t) - s_{12}(t) * (25i_3(t)).
\] (24)

When sampling occurs, the value that will be displayed on the screen of the oscilloscope will be proportional to the charge stored on the hold capacitors. This charge is detected by charge amplifiers, and the corresponding output value is given by

\[
Q = L \int_{-\infty}^{\infty} \frac{i_3(t)}{2} dt
\] (25)

with \(L\) a constant. The value of \(L\) is determined by a dc measurement.

First the effects of asymmetry of the circuit on the creation of a kick-out and the sampling process will be described.

The equivalent scheme of Fig. 3 represents a linear system.

This means that \(i_3(t)\) can always be written as the sum of a linear functional of the input signal \(a_1(t)\), which will be noted \(2i_{DF}(t)\), and of a linear functional of the two sources \(v_{AS}(t)\) and \(i_{AS}(t)\), noted \(2i_{CO}(t)\), which represents the effect of the asymmetry of the circuit. Looking at (24) reveals that this
The sampler impulse response, noted \( h(t) \), equals \( Q_{DI} \) when \( a_1(u) \) equals \( \delta(u + t) \). Using (29), (26), (27) and some trivial substitutions of the integrand variables, the following expression can be derived:

\[
h(t) = \int_{-\infty}^{\infty} \left( \int_{-\infty}^{\infty} K(-v, u) \, dv \right) s_{21}(t - v) \, dv. \tag{31}
\]

In what follows, the function \( K(u, t) \) will be calculated. \( K(u, t) \) will be equal to the current waveform \( i_{DI}(t) \) when \( v_{DI}(t) \) equals \( \delta(t - u) \). Using the equivalent scheme of Fig. 4, the following equation can be derived:

\[
\frac{K(u, t)}{2} = \frac{p(t)}{25} \delta(t - u) - p(t) \cdot \int_{-\infty}^{\infty} \left( s_{22}(t - v) + \frac{U(t - u)}{50C_H} \right) K(u, v) \, dv. \tag{32}
\]

In (32) \( p(t) \) is defined by

\[
p(t) = \frac{25g_A(t)}{1 + 25g_A(t)}. \tag{33}
\]

Note that \( p(t) \) is dimensionless, has an asymptotic value of \( 1 \) for \( g(t) \) going to plus infinity and that \( g(t) \) will always be positive, since it is the derivative of a voltage-current relationship of a stable diode. For a good sampler, the second term on the right-hand side of (32) is small compared to the first term of the right-hand side. For a perfect sampler, with \( C_H \) equal to infinity and \( s_{22} \) equal to 0, the second term of the right-hand side of (32) disappears completely. A good first-order approximation to \( K(u, t) \) can then be constructed by substituting the solution for a perfect sampler into the right hand side of (32). The result is

\[
K(u, t) \approx \frac{2p(t)}{25} \delta(t - u) - \frac{2p(t)}{25} p(u) \cdot \left( s_{22}(t - u) + \frac{U(t - u)}{50C_H} \right). \tag{34}
\]

Substitution of (34) into (30) and (31) results in the kick-out waveform \( k(t) \) and the sampler impulse response \( h(t) \):

\[
k(t) = \frac{2p(t)}{25} (1 - p(t) \ast f(t)) \ast s_{12}(t) \tag{35}
\]

and

\[
h(t) = \frac{2p(-t)}{25} (1 - p(-t) \ast f(t)) \ast s_{21}(t). \tag{36}
\]

In these equations \( f(t) \) is defined by

\[
f(t) = s_{22}(t) + \frac{U(t)}{50C_H}. \tag{37}
\]

Because of reciprocity of the input circuitry \( s_{12}(t) \) will equal \( s_{21}(t) \), such that the validity of the statement “kick-out equals impulse response” will be determined by the symmetry of \( p(t) \). Indeed, for the statement to be valid, \( p(t) \) needs to be equal to \( p(-t) \). In practice this situation will probably be approximated very well. Arguments are based on the approximate knowledge of the waveforms \( v_{S1}(t) \) and \( v_{S2}(t) \), which are the sampler “strobe pulses,” and on computer simulations. Because (33) shows that \( p(t) \) is a compressed version of \( g(t) \), since during
the diode forward bias the differential resistance of the diode is only about 10 Ω, a large part of the asymmetry of \( g(t) \) will be masked in \( p(t) \). Even if nothing is assumed concerning the symmetry of \( p(t) \), some quantitative limit exists concerning the maximum error in the frequency domain. The existence of this limit is based on the fact that \( p(t) \) is positive and limited in time (about 10 ps) [7]. Nevertheless, the determination of the amount of systematic error introduced by the possible asymmetry of \( p(t) \) is probably the most important remaining unknown in the "nose-to-noise" calibration procedure.

III. EXPERIMENTAL RESULTS
CONCERNING ACCURACY AND PRECISION

Introduction

As mentioned in [5] several measurement errors limit the accuracy and precision of the "nose-to-noise" calibration procedure. For the experimental results a setup was built with two different oscilloscope types. The first oscilloscope is an HP-54121T (bandwidth 20 GHz) and the second oscilloscope an HP-54124T (bandwidth of 35 GHz for the channel used). The fact that we use a 20 GHz system limits our conclusions to a frequency range not much higher than this frequency. In order to be able to make the same kind of conclusions for a frequency range of about 50 GHz, it is necessary to do the same type of measurements again but with two 54124T scopes, by using the 50 GHz bandwidth channels. At the time of the experiments two of these scopes were, however, not available.

Sampler Linearity

As mentioned above (see derivation of (12)) the signal levels need to be small in order for the sampler to behave linearly. This means that the dc-voltage applied (in practice the oscilloscope internal "OFFSET" function is used, such that we will call this value "offset" in what follows, cf. [5]) in order to create a kick-out pulse needs to be limited. To determine the amount of nonlinear distortion the following experiment is done. Several "nose-to-noise" measurements are performed, with the following offset voltages: 50, 75, 100, 125, 150, and 200 mV. The amplitude and phase of the Fourier transform of all experiments are calculated, each time normalized to the 50 mV offset experiment (with normalizing it is meant that the kick-out pulse with 200 mV offset is divided by 4). In an ideal case, all curves should be identical. The difference between the normalized amplitude characteristics and the 50 mV offset characteristic is shown in Fig. 5. Note that a curve which is closer to 0 corresponds to a smaller offset voltage. For an offset voltage of 100 mV there is a difference with the 50 mV offset experiment smaller than 100 mDB for the amplitude and smaller than 1 degree for the phase, and this for all frequencies smaller than 25 GHz.

Repeatability and Noise

To have an idea about the repeatability and the uncertainty due to noise, the following experiment is done. The oscilloscopes are connected and disconnected three times, and each time 1000 kick-out pulses are measured, with an offset voltage of 75 mV. Note that the third measurement was performed 24 hours later than the first. For the three measurements the average and standard deviation are calculated of both amplitude and phase (for the averaging "logarithmic spectral averaging" [5] is applied to avoid the effects of small timebase drifts). Next, the difference is calculated between the second and third measurement sequence and the first. From the standard deviation a 99% confidence interval of this difference is also calculated. The results of this are shown in Figs. 6 and 7. The figures show that, for an average of 1000 pulses with 75 mV offset, the 99% confidence interval is better than ±100 mDB for the amplitude and ±1 degree for all frequencies smaller than 35 GHz. The figures also show that the repeatability is better than these values. Note that the phase characteristics are aligned by applying appropriate delays (to compensate for small timebase drifts).

Timebase Drift and Jitter

Two stochastic timebase effects have an influence on the accuracy of the calibration procedure. First there is a timebase drift, with a typical time constant much higher than the time it
takes to do the acquisition of one kick-out pulse. This timebase drift is highly correlated with temperature. It appears on the screen as a smooth drifting of the whole pulse. A second effect is timing jitter. The time constant of this effect is much smaller than the time it takes to take one individual sample. Timing jitter appears on the screen as an additive noise with a standard deviation that is high where the slope of the pulse is high. Both effects introduce an error when regular averaging techniques are being used.

In order to avoid timebase drift, the experiments are done in thermostatic conditions, and "logarithmic spectral averaging" [5] is being used.

The timing jitter has a standard deviation of about 1.6 ps. This is estimated using the method as described in [8] and taking into account the amount of timebase drift. Note that "logarithmic spectral averaging" cannot compensate for the errors introduced by time jitter. For a jitter standard deviation of 1.6 ps, this error is about 175 mDB for a frequency of 20 GHz and 538 mDB for a frequency of 35 GHz. As explained in [9] this error can be compensated by multiplying all frequency components with the frequency-dependent function $C(f)$ defined as follows:

$$C(f) = e^{(2\pi f \tau)^2/2}, \text{ with } f \text{ the frequency and } \tau \text{ the jitter standard deviation.}$$

(38)

**Symmetry of p(t)**

Although the assumption that $p(t)$ approximately equals $p(-t)$ itself cannot be checked by doing "nose-to-nose" measurements, some consequences of this assumption can. One such consequence of the assumption to be valid is the fact that the phase characteristic of the Fourier transform of the measured kick-out pulse is almost invariant towards changes in the sampling diode bias voltages $v_{B1}$ and $v_{B2}$, nevertheless the amplitude characteristic will change drastically. As can be seen by looking at (35)–(37) this phase invariance should be perfectly achieved when $v_{B2}$ equals zero. In practice, the bias voltages can be changed by switching both oscilloscopes between "LOW" and "HIGH BANDWIDTH" mode. The resulting amplitude difference of the Fourier transform of the measured kick-out is about 5 dB for a frequency of 20 GHz. As can be seen in Fig. 8 the phase difference, however, is smaller than 1 degree for all frequencies lower than 20 GHz. Note the phase 99% confidence interval in the figure. It is assumed that the larger difference for higher frequencies is mainly due to the fact that a 54121T scope is used with a specified bandwidth of 20 GHz.

**IV. CONCLUSION**

The validity of the statement "sampler kick-out equals sampler impulse response" was theoretically proven starting from a generalized equivalent scheme of sampling heads. It was shown what approximations are needed, and what assumptions need to be made. The analysis shows the mathematical relationship between general sampler parameters and the kick-out waveform and impulse response. The sampler parameters under consideration are the sampling diode voltage-current characteristics, the hold capacitor values, the strobe pulse waveforms, the reverse bias values and the scattering parameters of the input circuitry. It was explained that the determination of the amount of systematic error introduced by the possible asymmetry of $p(t)$ is probably the most important remaining unknown in the "nose-to-nose" calibration procedure.

Several experimental results were given concerning the accuracy and precision of the "nose-to-nose" calibration procedure. In the future the same type of measurements with 54124T oscilloscopes could extend all conclusions to a bandwidth of 50 GHz.

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