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RESEARCH PAPER

The accurate comparison (or calibration) of solid-state noise sources made simple

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The accurate comparison (or calibration) of conventional solid-state noise sources is often thought to be a difficult process that only specialized laboratories can perform. In this paper, the uncertainties due to the impedance differences between the sources and their on and off states are emphasized and a simple formula is developed that makes them negligible. It requires an isolator being used in front of the receiver. In the experimental section, the errors associated with common practice are evidenced and compared to what can be achieved through the correction formula.

Keywords: Low-noise and communication receivers, Microwave measurements

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I. INTRODUCTION

This paper deals with the comparison of two (solid-state) matched noise sources characterized by their excess noise ratio *ENR*o. The proposed procedure is simple enough for a non-specialized lab engineer to reach transfer accuracies below 0.1 dB using standard measurement equipment. The procedure can be used:

(1) to check the consistency of two calibrated noise sources;
(2) to verify their evolution over time; and
(3) to calibrate one of the noise sources if the *ENR*o of the other is renowned to be accurate.

II. THE CLASSICAL COMPARISON/CALIBRATION FORMULA

The comparison/calibration of two noise sources is basically a double noise temperature measurement of a (theoretically) arbitrary receiver, first with source no. 1 (we will call the reference) and then with source no. 2 (we will call the unknown). The setup for this kind of measurement is described in details in Section V, with Fig. 5 visualizing the symbols used.

Let: T_n = receiver noise temperature;
T_r^{on} = reference source temperature in its on state;
T_x^{on} = unknown source temperature in its on state;
T_p = physical (ambient) temperature.

For solid-state noise sources the off temperatures T_r^{of} and T_x^{of} are supposed to be equal to the ambient temperature T_p. The ratio y_r of the available noise powers at the receiver output with the reference source on and off then writes [1]:

$$y_r = \frac{T_n + T_r^{on}}{T_n + T_p}. \tag{1}$$

Similarly, with the unknown source, we have:

$$y_x = \frac{T_n + T_x^{on}}{T_n + T_p}. \tag{2}$$

By eliminating the receiver noise temperature T_n between (1) and (2), it follows that [2]:

$$T_x^{on} - T_p = \left(\frac{y_x - 1}{y_r - 1}\right)(T_r^{on} - T_p). \tag{3}$$

Accordingly, the on temperature of the unknown source T_x^{on} can be computed from the measured y-factors y_r and y_x, the on temperature of the reference source T_r^{on} and the physical or ambient temperature T_p that are supposed to be known.

The conventional excess noise ratio of the newly calibrated source is [1]:

$$ENRo_x = 10\log\left(\frac{T_x^{on} - T_o}{T_o}\right), \tag{4}$$

with $T_o = 290$ K.

Formula (3) involves two assumptions that are only approximately true:

The receiver noise temperature T_n is the same during all four steps of the measurement (with reference source on, reference source off, unknown source on, and unknown source off). This is not perfectly true because the source impedance changes with the device and its state and the receiver noise temperature is a function of this impedance.

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The receiver available power gain is the same during all four steps. This is not perfectly true for the same reasons and the fact that the receiver available power gain is a function of the source impedance.

In the experimental part of this work, we will show that errors up to 0.4 dB typically follow because the above assumptions are imperfectly fulfilled. This is much more than for example the instrumentation error of modern noise figure meters, making impedance related errors the most important error cause when comparing noise sources.

III. AN IMPROVED CALIBRATION FORMULA

It is possible to overcome these problems if a receiver with an isolator in front of it is used. With an isolator indeed the noise temperature and the available power gain are closely related (this property is true for all passive devices, the corresponding formula being given in the Appendix). Thanks to it, the errors introduced by the source impedance variations can be eliminated almost completely in a simple way. As shown in the Appendix, formula (3) only has to be modified by a g_a factor:

$$T_x^{on} - T_p = g_a\left(\frac{y_x - 1}{y_r - 1}\right)(T_r^{on} - T_p), \tag{5}$$

$$g_a = \frac{G_{a1}(S_r^{on})}{G_{a1}(S_x^{on})}, \tag{6}$$

where $G_{a1}(S_r^{on})$ is the available power gain of the isolator connected to the reference source in its on state and $G_{a1}(S_x^{on})$ is the same with the unknown source.

Computing the g_a factor is a simple matter as:

(1) the S-parameters of the isolator are easily measured on a vector network analyzer (VNA);
(2) the available power gain G_a depends on the source reflection factor S by [1]:

$$G_{a1}(S) = |S_{21}|^2 \frac{1 - |S|^2}{|1 - S_{11}S|^2 - |S_{22} - \Delta_S.S|^2}, \tag{7}$$
$$\Delta_S = S_{11}S_{22} - S_{12}S_{21}.$$

IV. MEASURING THE S-PARAMETERS

A) The setup

All S-parameters were measured using an HP 8722C VNA and a Maury Microwave 8050M TOSL PC 3.5 calibration kit with sliding load. The reason why a sliding load is recommended will appear later.

B) The reflection factors of the noise sources

The noise sources are a Micronetics NS346AK ($ENR_0 \approx$ 16 dB) that is used as the reference source and a Micronetics NS346C ($ENR_0 \approx$ 16 dB) that is used as the unknown source. Their reflection factors in both states are shown in Figs 1 and 2. The first source has a stronger internal attenuator that reduces the reflection factor differences between the on and off states. With the second source these differences are much larger and can significantly affect the calibration accuracy.

C) The S-parameters of the isolator

The isolator is a DiTom D418018 dual junction unit with typically 40 dB isolation. Its S-parameters are shown in Figs 3 and 4 and indicate what performance is desirable.

V. THE NOISE MEASUREMENTS

A) The setup

The setup is presented in Fig. 5.

The noise sources and the isolator were already defined. The amplifier is a Miteq AFS5-08001800 low-noise device with about 35 dB gain and less than 2 dB noise figure. The SSB downconverter is a custom design with YIG preselector, fundamental mixing and frequency control by an external locking counter. The NFA is an Agilent N8973A instrument. To closely follow all frequency related variations, 401 points are made in the conventional 8–18 GHz band. The noise sources, isolator, and amplifier are all

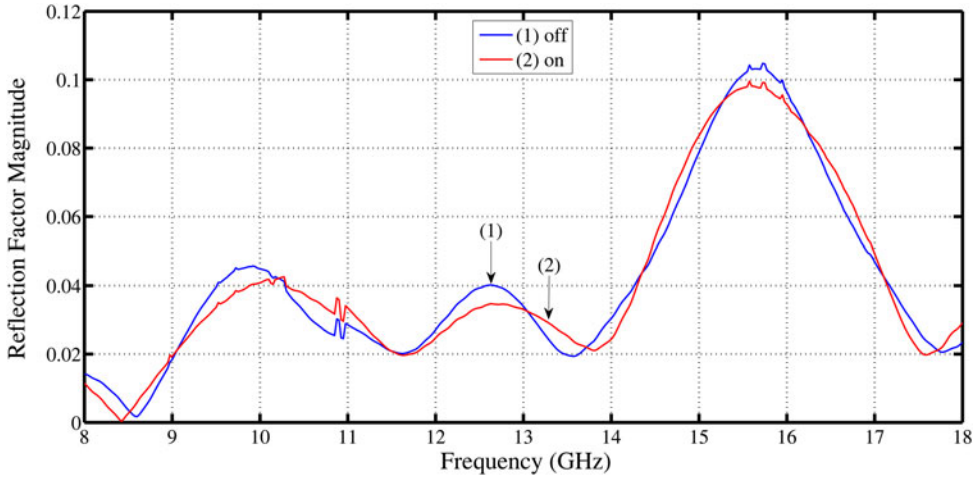


Fig. 1. The reflection factor of the reference source versus frequency. As this source has a 6 dB nominal ENR_0 (higher internal attenuator), the reflection factor differences between the on and off states are very small.

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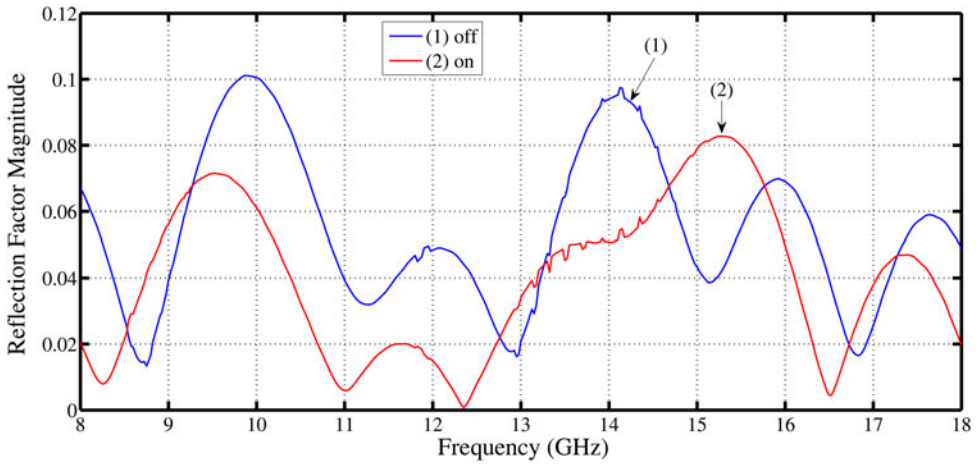


Fig. 2. The reflection factor of the unknown source versus frequency. As this source has a 16 dB nominal *ENR*o (lower internal attenuator), the reflection factor differences between the on and off states are much larger.

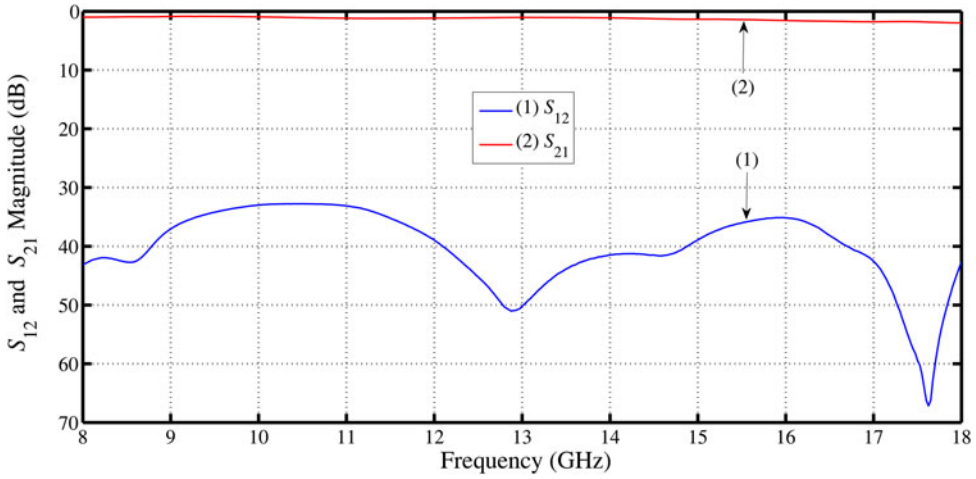


Fig. 3. The forward and reverse transmission factors of the isolator.

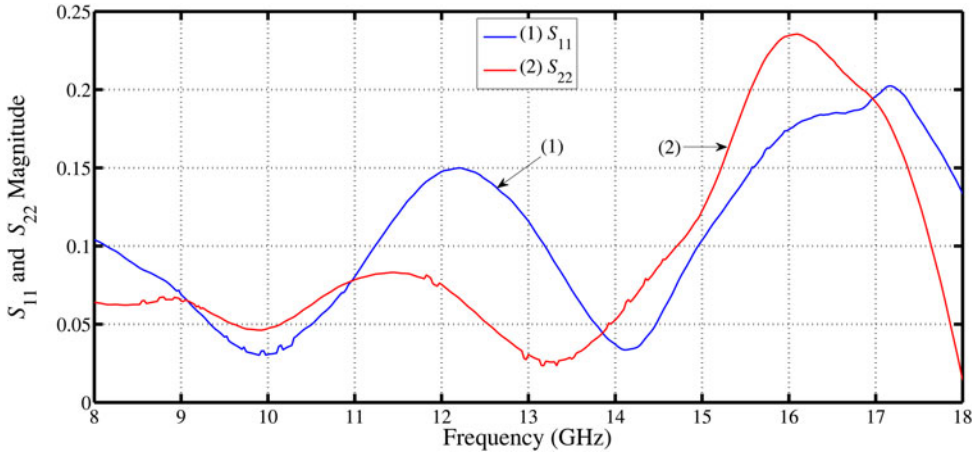


Fig. 4. The reflection factors of the isolator.

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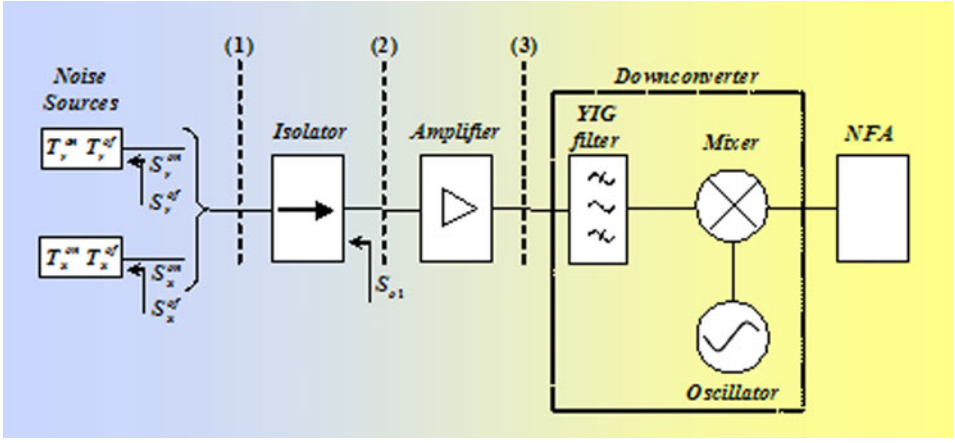


Fig. 5. The measurement setup.

broadband devices with a smooth noise and gain behavior versus frequency. Moreover, the lowest frequency is 8 GHz. So it seemed possible to use the largest bandwidth available on the NFA (4 MHz). To further reduce the random errors, all *y*-factor measurements were performed with an averaging factor of 16.

B) Comparing the noise sources in different ways

The noise setup can be used as a receiver in three ways by connecting the noise sources at the downconverter input, the amplifier input or the isolator input. We performed *y*-factor measurements with both sources at all three inputs. For the reference *ENR*o, we chose the calibration data by the manufacturer at every integer frequency in GHz and linear interpolation between these points. Based on the classical formula (3), an excess noise ratio can be computed in all three cases. They are called $ENRo_{xd}$ (downconverter input), $ENRo_{xa}$ (amplifier input), and $ENRo_{xi}$ (isolator input). The improved formula (5) of course only applies to the isolator input and leads to $ENRo_{xic}$ (isolator corrected). We now compare the set of results based on the classical formula to the results from the improved formula by considering the differences:

$$\Delta ENRo_{xd} = ENRo_{xd} - ENRo_{xic}, \quad (8)$$

$$\Delta ENRo_{xa} = ENRo_{xa} - ENRo_{xic}, \quad (9)$$

$$\Delta ENRo_{xi} = ENRo_{xi} - ENRo_{xic}. \quad (10)$$

They are plotted versus frequency in Fig. 6. The highest spread (up to ≈ 0.4 dB) occurs on the $ENRo_{xd}$ derived from the measurements at the downconverter input (plane 3 in Fig. 5). This is the input port with the poorest match and the highest noise temperature due to switches, the YIG filter and mixer. Also note the quick variation with frequency that follows from the fact the downconverter is an electrically long device in the 8–18 GHz band. Clearly better (spread below 0.2 dB) are the measurements at the amplifier input (plane 2 in Fig. 5) that has a closer match and a lower noise temperature. A step further is to use the measurements at the isolator input (plane 1 in Fig. 5) that differ from the corrected measurements by 0.15 dB at most. This legitimates the recommendation made in [3] to use isolators as a first stage when calibrating noise sources.

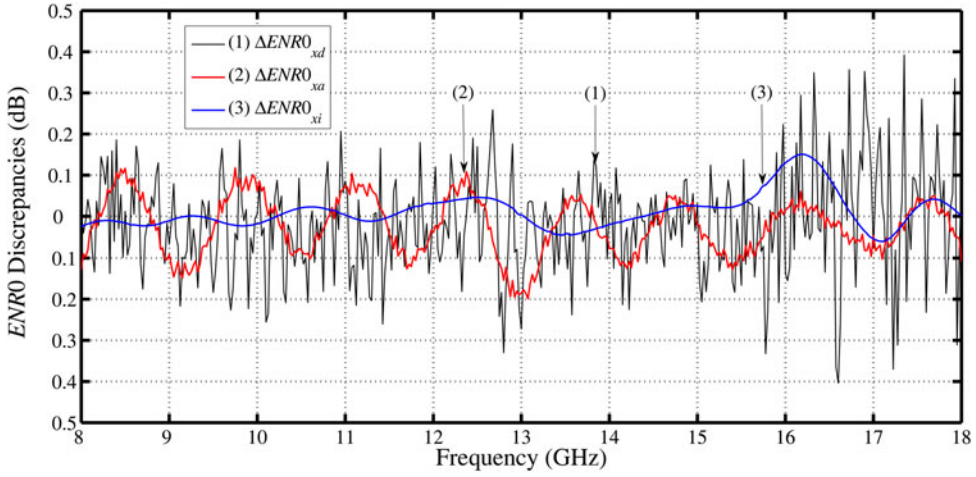


Fig. 6. The discrepancies between the *ENR*o evaluations.

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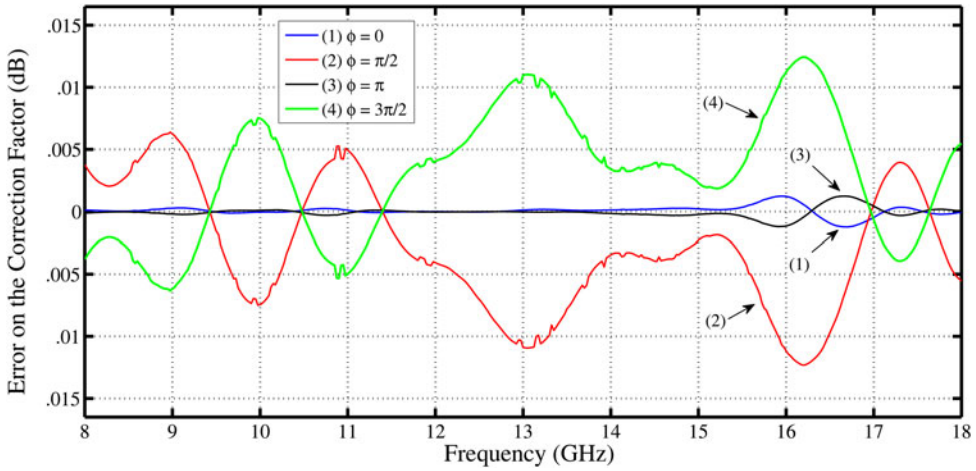


Fig. 7. If state-of-the-art *S*-parameter measurements are performed, the error on the correction (<0.015 dB) is expected to be much smaller than the correction itself (up to 0.15 dB).

Now the question arises what is the residual error when using the correction. Provided the isolator has a reverse transfer that is low enough, the associated error mainly depends on how accurately the correction can be computed. This subject is discussed in the next section.

VI. ABOUT THE CORRECTION ACCURACY

From formulas (3) and (5) it follows that the only difference between the classical and improved formula is the ratio g_a. Replacing the available power gain by its expression (7) and considering that for the isolator S_{12} is very small, we get:

$$g_a \approx \frac{1-\left|S_r^{on}\right|^2}{1-\left|S_x^{on}\right|^2}\left|1+S_{11}(S_r^{on}-S_x^{on})\right|^2. \qquad (11)$$

So the correction factor mainly depends on the source reflection factors S_r^{on} and S_x^{on} and the isolator input *S*-parameter S_{11}. Let us assume that all reflection factors are affected by a vector error:

$$\Delta S = \Delta s\, e^{i\phi}. \qquad (12)$$

We think that $\Delta s = 0.01$ is representative for the directivity error of a VNA calibrated by a sliding load, which is the dominant error when measuring small reflection factors [4]. To see the influence on the *ENR*o of such an error, we compute ENR_{xic} with all *S*-parameters modified by ΔS for $\varphi = 0$, $\pi/2$, π, and $3\pi/2$. The differences between these ENR_{xic} (with modified *S*-parameters) and the values with the "true" *S*-parameters are plotted versus frequency in Fig. 7. They show that the uncertainty on the correction factor is always much smaller than the correction itself.

As a consequence, it can be assumed that the impedance related error on the *ENR*o measurement of a noise source has been reduced from several tenths of a dB to several hundreds of a dB. At this level, the error becomes of the same order or smaller than the other errors occurring when two sources are compared (instrumentation *y*-factor error, random *y*-factor error, drift error of the setup, connector reproducibility error, receiver linearity error, physical temperature error inside the sources and isolator, etc.)

VII. CONCLUSION

Without caution, impedance-related errors are often the dominant contribution to the uncertainty when comparing two noise sources. Thanks to the use of an isolator combined with the correction formula presented in this paper, these errors can on the contrary be reduced to a level where they become negligible. Moreover, the formula is very simple and rests on easy measurements, namely the *S*-parameters of an isolator. There is no need to define a noise model for the receiver or to measure its input reflection factor, which is much more delicate.

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APPENDIX

DERIVATION OF THE IMPROVED CALIBRATION FORMULA

The improved calibration formula (5) only applies to the special case of solid-state noise sources but in theory it is equivalent to the radiometer correction procedures in use at NIST or NPL [5–10]. Moreover, it can be derived rather easily through concepts that are very common in noise theory such as noise temperature, y-factor, available power gain, and cascaded two-ports.

The exact expressions of the y-factors

In its on state the reference source has a reflection factor S_r^{on} and an available noise power in a bandwidth Δf given by:

$$kT_r^{on}\Delta f, \tag{5}$$

where k designates Boltzmann's constant.

Hereafter, the receiver is described as a cascade of two two-ports, the first being the isolator and the second being the rest of the chain. Subscript 1 corresponds to the isolator and subscript 2 to the rest of the chain. The subscript 12 stands for the whole receiver. If the receiver has an available power gain G_{a12} (S_r^{on}) and a noise temperature T_{n12} (S_r^{on}), the available output power is:

$$G_{a12}(S_r^{on})k(T_r^{on} + T_{n12}(S_r^{on}))\Delta f. \tag{6}$$

In the off state, this power becomes:

$$G_{a12}(S_r^{of})k(T_p + T_{n12}(S_r^{of}))\Delta f. \tag{7}$$

From (6) and (7), it follows that the y-factor writes:

$$y_r = \frac{G_{a12}(S_r^{on})}{G_{a12}(S_r^{of})}\frac{T_r^{on} + T_{n12}(S_r^{on})}{T_p + T_{n12}(S_r^{of})}. \tag{8}$$

Similarly, for the unknown source, replacing the r subscripts by x, we get:

$$y_x = \frac{G_{a12}(S_x^{on})}{G_{a12}(S_x^{of})}\frac{T_x^{on} + T_{n12}(S_x^{on})}{T_p + T_{n12}(S_x^{of})}. \tag{9}$$

The receiver noise temperature and gain

Using the Friis formula for the noise temperature of cascaded two-ports [1], the overall noise temperature T_{n12} is:

$$T_{n12}(S) = T_{n1}(S) + \frac{T_{n2}(S_{o1}(S))}{G_{a1}(S)}. \tag{10}$$

In this relation S designates the source reflection factor and S_{o1} the first stage output reflection factor, which is a function of the source reflection factor S given by [1]:

$$S_{o1}(S) = S_{22} + \frac{S_{12}S_{21}}{1 - S_{11}S}S, \tag{11}$$

where the S_{ij} are the first stage (isolator) S-parameters.

The isolator is a linear passive device at physical temperature T_p. Accordingly, the following relation holds between its noise temperature T_{n1} and its available power gain G_{a1} [1]:

$$T_{n1}(S) = T_p\left(\frac{1}{G_{a1}(S)} - 1\right). \tag{12}$$

Moreover, the available power gain G_{a12} of the cascade is the product of the available power gains of the first and second stages G_{a1} and G_{a2} provided they are computed for the right reflection factors:

$$G_{a12}(S) = G_{a1}(S)G_{a2}(S_{o1}(S)). \tag{13}$$

We now introduce a fundamental assumption which is of course an approximation but a much better approximation than simply assuming the overall noise temperature and available power gain do not depend on the source impedance. For an isolator (especially a two-stage design) the reverse transfer S_{12} is very small (in the experiments it was typically -35 dB or less). It then follows from equation (11) that the dependence of S_{o1} on S is very small. As a consequence, S_{o1} can be considered the same for all source reflection factors S_r^{on}, S_r^{of}, S_x^{on}, and S_x^{of} and both $T_{n2}(S_{o1}(S))$ and $G_{a2}(S_{o1}(S))$ can be replaced by the constant values T_{n20} and G_{n20}.

The overall result

Putting the expressions (10), (12), and (13) in (8) and (9) and making the above assumptions leads to:

$$y_r = 1 + G_{a1}(S_r^{on})\frac{T_r^{on} - T_p}{T_p + T_{n20}}, \tag{14}$$

$$y_x = 1 + G_{a1}(S_x^{on})\frac{T_x^{on} - T_p}{T_p + T_{n20}}, \tag{15}$$

Eliminating T_{n20} between (14) and (15) produces the improved calibration formula:

$$T_x^{on} - T_p = \left(\frac{G_{a1}(S_r^{on})}{G_{a1}(S_x^{on})}\right)\left(\frac{y_x - 1}{y_r - 1}\right)(T_r^{on} - T_p). \tag{16}$$

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