Noise Figure Measurement – A Reality Check

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Many LNAs now seen in EME stations and tested at conferences have a Noise Figure (NF) that is significantly lower than the measurement uncertainty of the test equipment. This raises some very serious questions, not only about the measurements but also about claimed specifications, system performance calculations and bragging rights!

This paper will identify some of the major uncertainties in NF measurements, including some little-known problems in the older generation of test equipment (HP8970/HP346). We will also see that even modern Noise Figure Analysers have not made a major impact on measurement uncertainties, because the problems are inherent in the measurement technique.

Every measurement technique has its own characteristic problems. The problems with conventional NF measurements are discussed at length in Agilent AN57-2*, Noise Figure Measurement Accuracy – The Y-Factor Method,* and in the other references provided.[1] AN57-2 also warns against many types of "avoidable operator errors"; but this paper concentrates on the measurement uncertainties that are inherent in the technique itself.

Another problem about NF measurement is that amateur radio has a very poor collective memory. Even though most of the information in this paper has been known for 20 years or more, relatively few amateurs seem to be actively aware of it. This paper will provide a much-needed 'memory refresh' about the reality of NF measurements.

The Y-Factor Measurement Method

Almost all noise figure measurements use the 'Y-Factor method', so first we need to explain how this method *should* operate. Then we can move on to the problems.

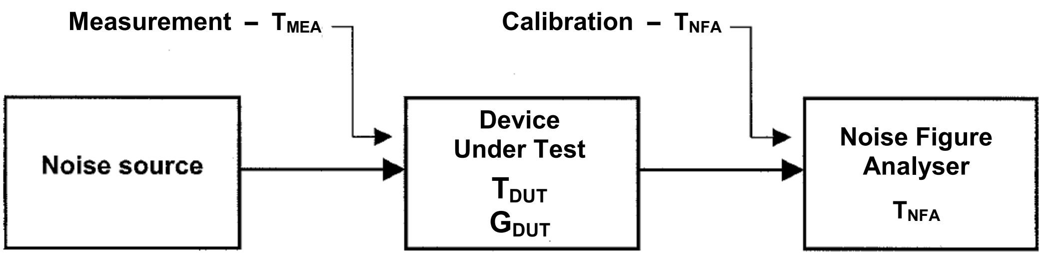

Figure 1: Test setup for Y-factor noise measurements

The Y-factor technique uses a **Noise Source** and a **Noise Figure Analyser (NFA)**. Initially the two items are connected directly together to calibrate the system. Then the Device Under Test (**DUT**) is inserted between them to measure its NF and gain.

The noise source can be switched either ON or OFF. The ON state generates noise at a calibrated power level; and even when switched OFF, the noise source still generates a

1. Most of the key references are provided as PDF files on the Conference DVD.

smaller amount of thermal noise power. The equivalent thermal noise temperatures are usually written as T_{ON} and T_{OFF}, or alternatively T_{HOT} and T_{COLD}.

The Noise Figure Analyser is an instrument for measuring *relative* noise power levels. The NFA does not need to be calibrated in terms of absolute power levels (that calibration resides in the noise source) but the NFA must be able to measure noise power *ratios* with very high accuracy and over a wide dynamic range.

The **Y factor** is simply the ratio of two noise power levels, one measured with the noise source ON and the other with the noise source OFF:

$$Y = N_{ON}/N_{OFF} \qquad (1a)$$

Because noise power is proportional to noise temperature, we can also say:

$$Y = T_{ON}/T_{OFF} = T_{HOT}/T_{COLD} \qquad (1b)$$

Other variants of the Y-factor method involve switching between two different noise sources, one 'hot' and the other 'cold', or even changing the physical temperature of the source. We will discuss some of these options later. For now, though, we will concentrate on the most common kind of Y-factor measurements using a single switchable noise source and the companion NFA (either commercial or home made).

The advantages of this technique for a general-purpose test instrument are that the noise source can provide two different, calibrated power levels covering the entire frequency range. The use of broadband noise also makes it unnecessary to measure the noise bandwidth of either the NFA or the DUT (in most cases; but see later).

The penalty of using noise as a test signal is that noise is a statistical phenomenon, so this introduces large and unavoidable uncertainties. Accurate measurements of noise power take time. To compute an averaged value of the Y factor with a reasonably low statistical uncertainty will typically require several hundred sample measurements of N_{ON} and N_{OFF}. A modern NFA will control the repeated on/off switching of the noise source, and then do all the necessary computations to determine the NF and gain of the DUT.

NF definitions

Noise factor F is a dimensionless ratio:

$$F = \text{(S/N at input port) / (S/N at output port)} \qquad (2a)$$

Noise Figure NF is measured in dB:

$$NF = 10\log_{10} \{\text{(S/N at input port) / (S/N at output port)}\} \qquad (2b)$$

where the S/N quantities are linear ratios between signal power and noise power.

In other words, **NF is defined as a ratio between two Y-factors**.

The engineering definitions of F and NF, derived from the basic definitions above, are:

$$F = 1 + (T_E/T_0) \qquad (3a)$$

$$NF = 10 \log_{10}\{1 + (T_E/T_0)\} \quad \text{dB} \qquad (3b)$$

where T_E is the effective (or equivalent) input noise temperature of the device.

Equation 3 also introduces the important reference temperature T_0 which is defined as 290 K (16.8°C, 62.2°F).

Calculations in Y-factor measurements

NF is defined at the input of the DUT but the noise power is being measured at the output, so any measurement of NF must include an accurate measurement of the device gain G_{DUT}. We can see this by developing some equations for the two-step measurement method shown in Figure 1:

Calibration (NFA only): $Y_{CAL} = N_{NFA\ ON} / N_{NFA\ OFF}$ (4)

Measurement with DUT: $Y_{MEA} = N_{MEA\ ON} / N_{MEA\ OFF}$ (5)

To find the true gain and noise temperature of the DUT, we now need to remove the effects of the NFA. Full details are given in AN57-2.

The noise source parameters T_{HOT} and T_{COLD} are available from the ENR (Excess Noise Ratio) calibration of the noise source, plus a measurement of the physical temperature T_{COLD}. ENR is formally defined as:

$$ENR = (T_{HOT} - T_{COLD}) / T_0 \tag{6a}$$

$$ENR = 10 \log_{10} \{(T_{HOT} - T_{COLD}) / T_0\} \quad \text{dB} \tag{6b}$$

where T_0 is the standard reference temperature of 290 K. ENR calibration data are normalized to T_{COLD} = 290 K, so T_{HOT} must always be corrected to account for the actual physical temperature T_{COLD} which can only be measured inside the noise source itself.

$$T_{HOT} = (ENR \times T_0) + T_{COLD} \tag{7}$$

From the calibration step which measured Y_{CAL} the NFA can calculate:

$$T_{NFA} = (T_{HOT} - Y_{CAL} T_{COLD}) / (Y_{CAL} - 1) \tag{8}$$

And the same again for the measurement with the DUT in place:

$$T_{MEA} = (T_{HOT} - Y_{MEA} T_{COLD}) / (Y_{MEA} - 1) \tag{9}$$

From the four different noise power measurements the NFA can calculate the gain of the DUT:

$$G_{DUT} = (N_{MEA\ ON} - N_{MEA\ OFF}) / (N_{NFA\ ON} - N_{NFA\ OFF}) \tag{10}$$

And finally:

$$T_{DUT} = T_{MEA} - (T_{NFA} / G_{DUT}) \tag{11}$$

The NFA will normally convert T_{DUT} into NF, and display NF_{DUT} and G_{DUT} in the normal engineering units of dB.

That is how it's all done, through quite a long chain of calculations. So where do the problems arise?

Generic Problems with Y-factor Measurements

The problems with Y-factor measurements are largely hidden in the algebra, so let's try a different view, and look at this problem through some simple graphics. Then we can go back to those equations with a better understanding of what is really happening.

The simplified case

To make this even simpler, let us temporarily assume a noise-free NFA. This removes the need for a calibration step, so we can obtain both T_{DUT} and G_{DUT} from just a single Y-factor measurement. Figure 2 illustrates this simplified case. The horizontal axis is the 'input power' to the DUT using a scale of noise temperature. The vertical axis is the 'output power' measured within the NFA, on a linear scale.

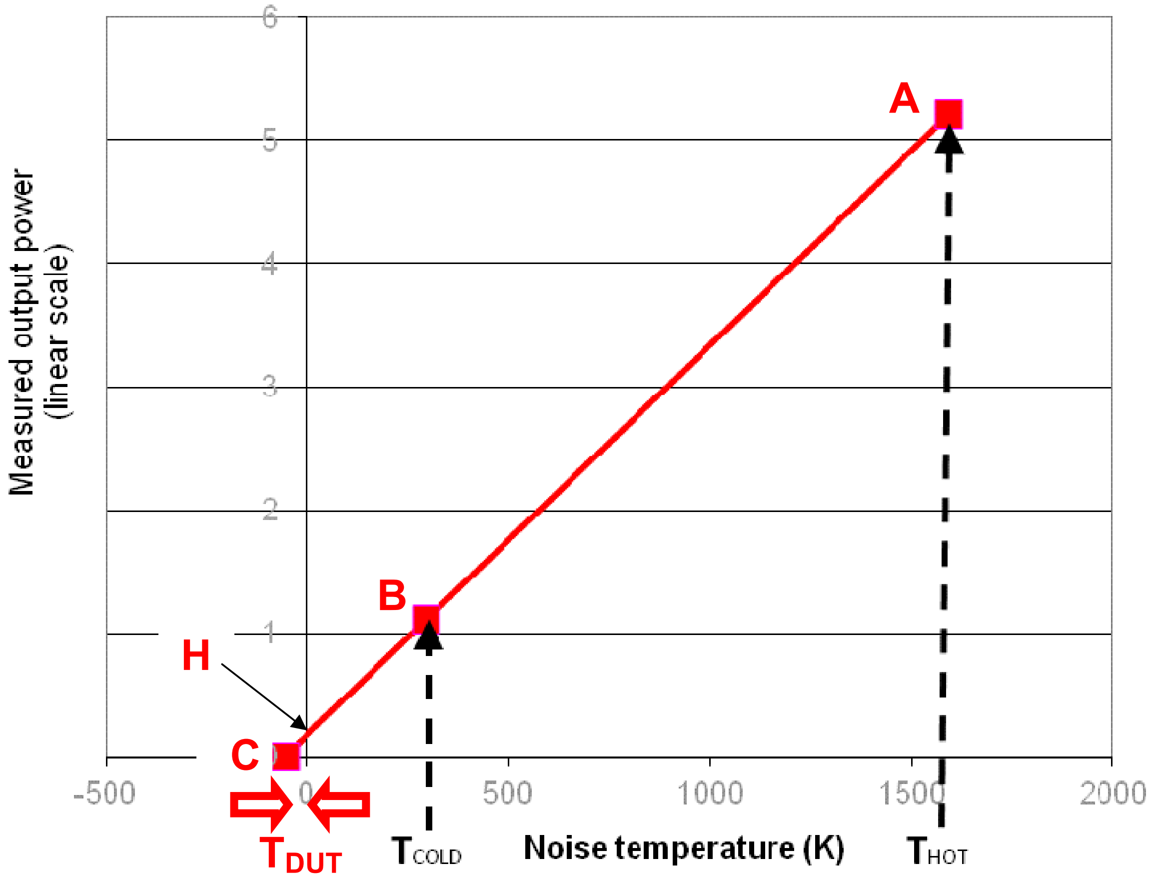

*Figure 2: Y-factor measurement requires a long extrapolation to find T_{DUT} at point **C**. (Simplified case with a noiseless NFA.)*

The Y-factor method always requires *pairs* of measurements. With the noise source switched on, the input power into the DUT corresponds to T_{HOT} and the measured output gives us point **A** in Figure 2. With the noise source switched off, the noise input power falls to T_{COLD} and the output power falls to give us point **B**.

Now extrapolate line **A–B** backward towards the origin at (0, 0). Because the DUT is generating internal noise, there would be some output power even when the input noise temperature was zero (this is point **H**). Extrapolating even further backward, the line eventually reaches zero output power at point **C**. From the geometrical construction, point **C** corresponds to $-(T_{DUT})$. [2]

The slope of the line **A–B** gives us the gain of the DUT as shown by equation 10. We will need that value of G_{DUT} later in the calculations.

Figure 2 is drawn to scale using realistic values, so now we can see the fundamental problems with the accuracy of this method.

We are trying to measure T_{DUT} as a very small distance along the noise temperature axis, by making a long extrapolation from two measurement points **A** and **B** that are a very long way away. Any small errors in the co-ordinates of points **A** and **B** will be

2. Notice the minus sign ahead of T_{DUT}! Obviously there is no such physical thing as negative noise temperature.

greatly magnified at point **C**, causing very significant errors in the final measured value of T_{DUT}.

Here is a very brief summary of the potential errors in drawing that line, **A**–**B**–**C**.

Error in noise source ENR

The **Excess Noise Ratio** (ENR) of the noise source is defined as:

$$ENR = (T_{HOT} - T_{COLD}) / T_0 \qquad (6a)$$

This value defines the horizontal distance between points **A** and **B** in Figure 2. ENR data are obtained from a factory calibration, and have a typical uncertainty of **± 0.2 dB or worse** (see later). Any error in the ENR calibration will affect the slope of line **A**–**B** and hence the intercept at point **C**.

Error in T_{COLD}

T_{COLD} is the physical temperature of the noise source. Through the definition of ENR, any error in T_{COLD} will affect T_{HOT} as well, so the whole line **A**–**B**–**C** moves sideways. Every 1°C error in T_{COLD} will cause an approximately equal error in the measured value of T_{DUT}.

When attempting to measure noise figures on the order of 0.2 dB, which is a noise temperature of only 13 K, we simply cannot afford to guess at T_{COLD}. We need to measure it accurately, **inside** the noise source if at all possible.

Gain errors in the DUT

Non-linearity in the DUT is a very serious error because it would mean that **A**–**B** is not a straight line but a curve. We also assume perfect linearity when we customarily refer all of the physical noise sources within the DUT to a single *equivalent* noise source at the input, so that the correct amount of magnified input noise will appear at the output. Curvature in the DUT transfer characteristic would mean that the linear extrapolation is no longer valid… but we don't know by how much.

Linearity is a reasonable assumption for a single-stage DUT handling very small signals; but it begins to fall apart with multi-stage LNAs that have a very high gain, and is certainly suspect when measuring complete receivers. When the overall gain is very high, some gain compression could easily occur in later stages of the DUT. We always tend to underestimate the range of peak values that are present in noise signals, and underestimate the amount of 'headroom' required to avoid peak compression that will then alter the mean value.[3]

Gain change in the DUT can be a very serious error when measuring certain types of LNA. The gain of the DUT is calculated from the measured slope of line **A**–**B**; but this depends on the hidden assumption that G_{DUT} does not change when the noise source is switched on and off. This point requires more detailed discussion later.

NFA and calibration errors

Even the simplified case of Figure 2 has revealed several serious sources of error or uncertainty. When we add in a real-life NFA, we find even more.

As shown in Figure 1, before making a real-life measurement we have to do a calibration: a hot/cold Y-factor measurement with the noise source connected directly to the

3. If any AGC system is active in the DUT, the measurement will be ruined – so that is another operator error to be avoided.

NFA. Figure 3 shows this in graphical form – and once again, everything is drawn to true scale.

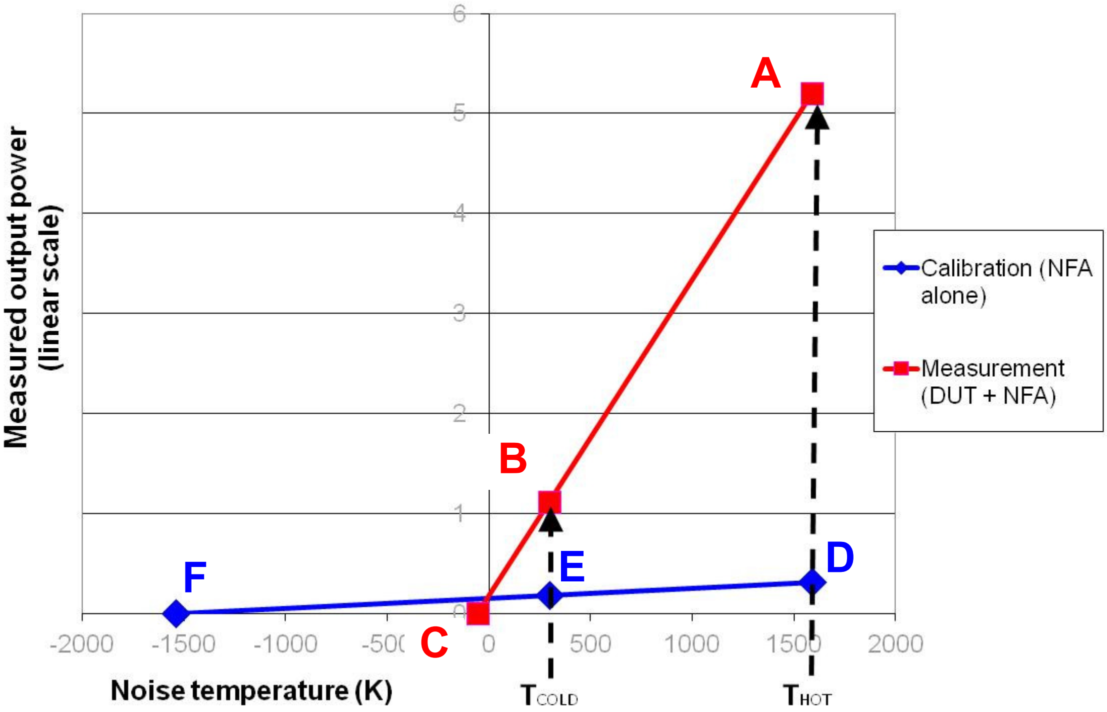

*Figure 3: A real-life measurement also includes the calibration **D–E–F**.*

The calibration step gives us a new line **D–E–F**, where point **F** represents the noise temperature of the NFA itself.[4] Because the NFA is a general-purpose broadband instrument, its NF is quite high, typically about 8 dB or 1500 K. This measurement of T_{NFA} is subject to quite large errors because of the very shallow slope and the long extrapolation to point **F**, but in most cases it plays a relatively small part in the final calculation of T_{DUT}.[5]

We are now ready to look at the errors inside a real-life NFA. Many of these errors apply particularly to the older-generation NFAs like the HP8970 series.

Non-linearity errors in the NFA

We can see from Figure 3 that measurements **A** and **B** take place at two quite different output levels, while **D** and **E** are at a lower level still. Non-linearity errors within the NFA can therefore be quite important, especially with a high-gain DUT which pushes point **A** even higher. Professional NFAs are designed to extreme standards of linearity. Typically they divide their IF amplification into blocks, with attenuators switched in and out automatically to extend the dynamic range. If the user makes reasonable efforts to avoid unnecessary demands on the dynamic range, the linearity error in a professional NFA should be less than ± 0.05 dB across the range required. Amateur NFAs are often much worse that that.

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4. Point **F** represents $-(T_{NFA})$. Once again, notice the minus-sign!
5. In the final calculation of T_{DUT} using equation 11, T_{NFA} (and any associated error) will be divided by G_{DUT} which is usually a large number.

Noise power measurement errors

NF measurements are based on the accurate measurement of noise power. The older HP8970 series of analysers used diode detectors which had some small errors in measuring noise signals with a high crest factor. The more modern Agilent NFA series use DSP which is much more accurate. An ADC takes samples of the noise voltage, which are then accurately converted to noise power density using a root-mean-square algorithm.

Frequency errors

The HP8970 and similar NFAs of that generation are liable to considerable frequency errors and drift, which can be important if the DUT is a narrowband device.

The receiver core of the HP8970 is essentially the same as the HP8558 1.5 GHz spectrum analyser, with a special RF stage and a new 20 MHz IF strip added. It uses a free-running YIG first oscillator in the frequency range of 2.0–3.5 GHz in the earlier HP8970 series and 3.9–5.9 GHz in the HP8970B. The frequency readout is not a counter display, but is merely calibrated from the YIG magnet current. Even if the instrument has been recently calibrated there can be significant errors in the frequency readout, and possibly quite serious errors in an older instrument that has not be calibrated for many years.

Also there can be serious errors due to frequency drift, which again will not appear on the readout. There is a system which periodically tunes the LO to 'zero hertz' to correct for LO breakthrough and to partially correct for frequency drift, but between these corrections I have seen drift as rapid as 1 kHz/second.

Frequency errors and drift of this magnitude were not a problem for testing the wideband military/aerospace equipment which was the instrument's main market all those years ago, but they are certainly an issue when testing narrowband receivers, and may also become significant for narrowband preamps. If you have both a HP8970 and a microwave synthesizer, you may wish to modify the 8970 for external LO operation, and possibly also add a stable second LO.

The more modern N8973 family fixed these problems with a synthesized LO. Its frequency accuracy is proportional to whatever 10 MHz frequency standard you use, plus or minus a further few hundred hertz contributed by the two free-running crystal oscillators in the 21.4 / 6.25 MHz IF system.

Bandwidth

The HP8970, the Maury/Ailtech equivalents and the (obsolete) Agilent base model N8972 only make measurements in a fixed 4 MHz bandwidth. This is fine for measuring wideband DUTs because the wide IF bandwidth gives rapid averaging of the noise power levels and little jitter in the results.

But amateur-band preamps and receivers may be much narrower than 4 MHz, and the problem comes when you calibrate the NFA at its own bandwidth of 4 MHz and then *change* the measurement bandwidth when you insert the narrowband DUT. This produces a measurement error because it defeats the assumptions of the NFA calculations.[6]

The error cannot be corrected by inserting two simple numbers to represent the respective equivalent noise bandwidths of the DUT and the NFA, because the composite noise bandwidth of the two units in cascade is not the simple difference between the

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6. For the purposes of this paper we regard this as an avoidable 'operator error'.

individual equivalent bandwidths. It is the integral of the product of the two actual pass-band shapes, and has to be evaluated by a series of point-by-point measurements.

This bandwidth error will be significant unless the instrument's own noise contribution (the T_{NFA} / G_{DUT} term in equation 11) is so small that any uncertainty becomes relatively unimportant in the final result. If that isn't possible, you would need to calibrate the system with a narrowband filter inside the loop, and even here than can be an error because the new noise bandwidth is not completely defined by that filter alone.

One big change in the N8973 family of NFAs was the facility for switchable narrow filters implemented in DSP. The NFA bandwidth can be reduced to 100 kHz which will be significantly narrower than the bandwidth of most DUTs and will thus define the composite bandwidth of the measurement. Narrower filters will require proportionally more averaging time to achieve the same level of statistical uncertainty, but the DSP can help to reduce the time required for swept measurements by processing several different frequency slots at the same time. However, there was no point in having narrower filters in the HP8970 family because it did not have the necessary frequency accuracy and stability to use them.

Summary so far

We have looked at the Y-factor measurement method from two different viewpoints, equations and graphics. The equations are definitive, but in some ways the graphics are more revealing about the elusive nature of T_{DUT} – hiding down in a corner, close to the origin (0,0) but far away from the points **A**, **B**, **D** and **E** that we actually measured.

Specific Problems with Some DUTs

All the problems identified above are generic, and in principle they can affect measurements on all different types of DUT. But the practical question is: ***How big are these errors for this LNA that I'm measuring?***

In this section we will identify specific problems that affect some DUTs much more than others.

Noise and gain contours

Noise figure, noise factor, and noise temperature are all simplifications of reality. They are simplified models of what is happening in our devices. The model takes all the different noise-creating mechanisms in our device and references them to an **equivalent** value at the input port. We see 'equivalent' and 'T_E' in the formal definition of noise temperature, but these are often omitted in everyday engineering. The problems start when we forget *why* they were necessary.

We are actually blaming all the device noise on a *fictitious* noise source applied to a *fictitious* duplicate input to *fictitious* noiseless device. This is useful and necessary fiction, because it allows us to handle something which would otherwise become too complicated. But it isn't real! Like any 'Black Box' model that represents the internal workings of a device to the outside world, it does not describe what is going on *inside* the real device… and sometimes it tells lies.

The main thing that we lose in referring noise parameters to the input port is that the device NF is affected by the source and load impedances. We 'patch' this problem by making the fictitious input noise source dependent on the source impedance. Device manufacturers usually construct a contour map on a Smith chart showing lines of constant NF depending on the input impedance presented to the device, and yet again we simplify reality by assuming that the contours are circular (Figure 4). On the map of complex impedance, the NF contours look like a shallow circular depression. This also neglects the lesser effect of the load impedance on NF, which is a reasonable approximation *except* when working at frequencies where available gains are low and S_{12} is large.

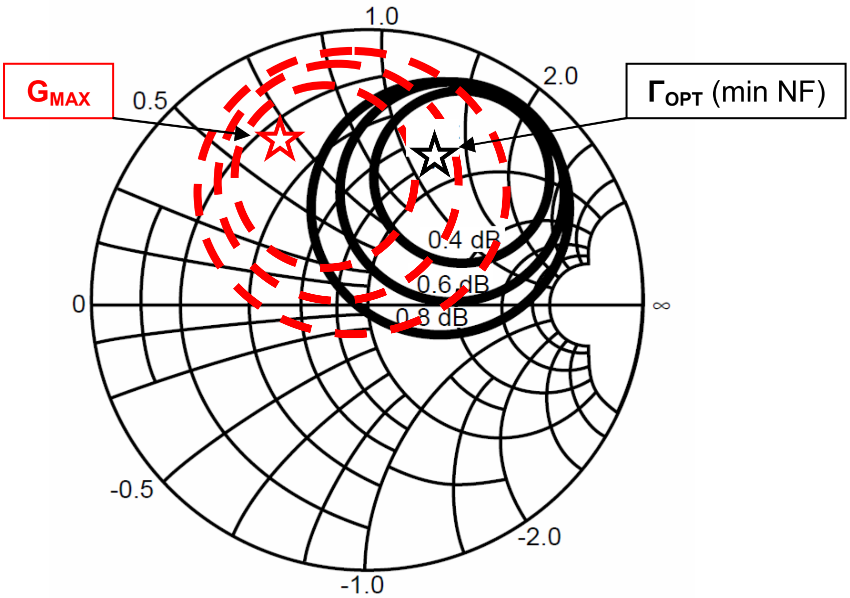

Figure 4: Typical contours of NF (assumed to be circular) on the Smith chart.

But on that same map of input impedance we also need to notice the contours of gain. Where NF contours look like a shallow depression, gain contours often look like quite a steep hill. Maximum gain does not occur at the same input impedance as minimum NF, so whenever we are adjusting the input circuit to find Γ_{OPT} at the bottom of the NF depression, we are also moving across the *slopes* of the gain peak which are much steeper. Small changes in the input circuit can often lead to quite large changes in gain; and this makes the NF measurement quite vulnerable to gain-related errors.

DUT-related gain errors

In *DUBUS* issues 4/1988 and 4/1990, Rainer Bertelsmeier DJ9BV[7] explained very clearly how NF measurements can be affected by the difference in the output impedance of a noise source between its ON and OFF states. The output impedance of the noise source becomes the input source impedance for the DUT, so any changes will affect both the internal noise and the gain of the DUT. This in turn will falsify the computation of NF.

DJ9BV developed an equation which showed the effects of these parameters:

$$\Delta G = \frac{1 - 2\,|S_{11}|\,|\Gamma_{OFF}|\cos(\Phi_{11} + \Phi_{OFF})}{1 - 2\,|S_{11}|\,|\Gamma_{ON}|\cos(\Phi_{11} + \Phi_{ON})} \qquad (8)$$

where ΔG is the gain error (G_{ON} / G_{OFF}, error-free value = 1)

$|S_{11}|$ is the magnitude of the DUT's input reflection coefficient

Φ_{11} is the phase of S_{11}

$|\Gamma_{OFF}|$ is the magnitude of the noise source reflection coefficient when the noise source is OFF

7. SK 2008. Rainer's presence at these EME Conferences is sadly missed.

Φ_{OFF} is the phase of the noise source reflection coefficient when the noise source is OFF

$|\Gamma_{ON}|$ and Φ_{ON} are the changed values when the noise source is ON.

The gain error ΔG (when converted into dB) translates directly into the error in NF. This error can be minimized by choosing both the correct kind of noise source and the correct kind of DUT. In the noise source, ΔG is minimized when Γ_{ON} and Γ_{OFF} are both small and equal, and this is achieved in a noise source like the HP346A which contains a larger value of built-in attenuator. In the DUT, ΔG is minimized by using a design with a low value of S_{11} – in other words, by choosing an LNA design that has reasonably good impedance matching at the input.

DJ9BV produced some horrifying graphs showing what can go wrong if these precautions are ignored when using the HP346B noise source to measure a 1980s-style LNA with a high-Q tuned input (and $|S_{11}|$ close to 1 at all phase angles). Simply by tweaking the input tuning, which varies the phase angle, it is easy to falsify the *indicated* value of NF by as much as 0.4 dB in either direction! It isn't hard to guess which direction people preferred…

Most people now understand that HP346B is an unsuitable noise source for measuring LNAs because of the large change in output impedance between the ON and OFF states (and also because the ENR is far too high, 15 dB). Less well understood is that this isn't a complete cure: the errors are reduced but they didn't go away completely. Barely understood at all is the role of the DUT design itself.

Figure 5 shows how the *indicated* NF of two different types of LNA can be made to vary by adjusting the input tuning. These graphs are calculated for a 'good' noise source like the HP346A.The solid red line is for a typical old-style preamp with a high-Q tuned input circuit and very poor input match, and as can be seen, the measured NF can easily be 'sweetened' by 0.02 dB or more. When modern ultra-low-noise devices are used in a 1980s-style LNA design, this kind of error begins to be significant.

However, when the same modern device is used in a modern LNA design, where attention has been given to good input matching, the phase-angle effect remains very small and well under control.

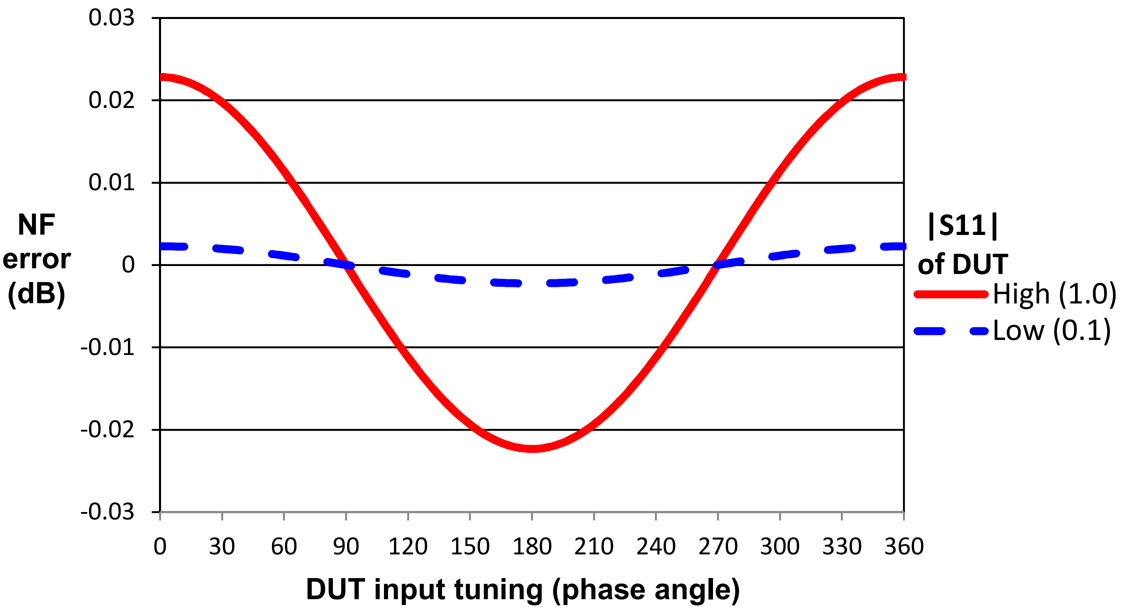

Figure 5: Variation of indicated NF with input tuning

It is vital to understand the consequences of this error, if it is not controlled:

1. **The minimum indicated NF is not the true minimum.**

   The NFA is no longer helping you to find the optimum noise performance – most likely, it is leading you **away** from the true minimum. By tuning the preamp input, the displayed value of NF can easily be 'sweetened' by 0.02 dB, while at the same time making the true NF *worse*.

2. **Professional noise figure analysers can deliberately tell lies**.

   Although negative values of NF are physically impossible, in some cases a negative error should genuinely drive the *displayed* value of NF below 0 dB. It is very unhelpful that some NFAs quietly suppress any negative values in the NF readout! The user needs to see those impossible results, as a warning that the measurement has gone seriously wrong.

This potentially fatal error been known by EMEers since the 1980s, or even before… but 30 years later it seems to have been forgotten.

When the a modern ultra-low-noise device is used in a modern LNA design that has paid good attention to input matching, the phase angle effect once again becomes small and well under control. NF measurements will still be subject to all the *other* errors and uncertainties described in the rest of this paper, but at least the indicated minimum value will be very close to the true minimum.

ENR Calibration

ENR calibration is probably the largest unavoidable source of uncertainty in NF measurements because it isn't something you can do for yourself. Even the big players like Agilent, Anritsu and Rohde & Schwartz obtain their primary ENR standards by cross-calibration from other, better standards which are ultimately traceable to the world network of national standards laboratories.

Excess Noise Ratio is a concept from the older, simpler days of NF measurement when small errors were either ignored or completely unthought-of. Today these errors are comparable with the small values of NF that we are trying to measure, so the simple equations of the old days will *always* require quite detailed corrections.[8] Technically it would be better to calibrate noise sources directly in terms of noise temperatures, T_{ON} and T_{OFF} but for reverse compatibility with existing industry standards, all noise sources are calibrated in terms of ENR (dB) and the calibration uncertainties are expressed in the same units.

During the development of the Agilent NFA (Noise Figure Analyzer) family of instruments we planned a new family of noise sources to automate some of the things that users often forgot to do. With the old HP8970 instruments it was easy to use a set of ENR values for a different noise source from the one you had connected. It was also easy not to bother about measuring the true value of T_{COLD}, and to let the NFA use its 290 K default. These shortcuts can result in large errors which justified the development of a new family of 'Smart Noise Sources'. At the same time we wanted to improve the ENR uncertainty as much as possible and this forced a review of the traceability chain right back to national standards laboratories.

8. You cannot know *which* of the errors are important until you have calculated *all* of them.

The ENR calibration business

The calibration of your noise source traces back to a different primary noise standard for each of its frequency points. Commercial noise sources are routinely calibrated against noise standards at 10 MHz, 100 MHz and then at integer GHz frequencies above that. Each of these primary noise standards contains precision termination resistors held at carefully measured temperatures, connected to a noise measuring receiver. Every possible source of error is measured and corrected with the utmost care. We already know that accurate impedance matching is crucial, and that the cable or waveguide connection from the resistor to a measuring receiver will have losses distributed along its length. A further problem is that the cable or waveguide will have a temperature profile along its length, from the termination resistor to the 290 K environment of the calibrated measuring receiver. The thermal noise from the termination is attenuated by this loss, while new noise is added. A set of corrections must be calculated for all of these effects.

World-wide, the national standards labs exchange calibrated sources to compare results, and they steadily improve their systems. The standards labs will initially include a prudent allowance for future drift in their error budgets, and as the noise standard gradually proves to be stable over time, this allowance will be reduced.

The older HP346 noise source family had been calibrated ultimately from the US standards at NIST in Boulder, Colorado, but a problem there had damaged some equipment and caused them to reset their error allowances to larger initial values. It would be years before they got their proven stability back again. (I heard that it was a leaky roof that did the damage.)

Agilent looked around for a national standards lab that could offer tighter uncertainty specifications and chose NPL near London, who offered the best figures. This also offered an improvement for any HP346 sources calibrated after the change-over, until the introduction of the new N4000 range. The standards lab business is a competitive industry, so Agilent's choice of lab may have changed again more recently. It is also quite possible that different labs could be used for different frequency points.

Calibration of noise sources directly against primary standards is obviously very costly, so the company would never use its secondary-standard sources on the production line. Those sources are kept in the corporate standards lab and used to calibrate sub-standards, which in turn are used to calibrate the commercial products. This means that a commercially available noise source is at least three steps removed from its primary standard, and at each step the ENR calibration uncertainty increases.

5 dB, ± *How* much?

We are all aware that modern LNAs should be measured using a noise source with an ENR of about 5 dB. Figures 2 and 3 are calculated using this value, which is about the right magnitude to minimize calculational uncertainties[9] with most low-noise DUTs.

The ENR calibration data can be read from the label on a noise source to the nearest 0.1 dB at each frequency. The same data are available with an extra significant figure if you have access to the latest calibration sheet (the added precision is not physically significant but it avoids unnecessary rounding errors). Take care when copying this

9. The measured numbers in an NF calculation are already 'fuzzed' by the statistical properties of noise. Equations 8–10 show that several subtractions are involved, so it is very important to avoid the further uncertainties that arise when subtracting two numbers that are almost equal. The best situation is usually when all the numbers involved are similar in magnitude, but still distinctly different from each other.

data into your NFA or spreadsheet; and if you have more than one noise source, also take care to select the *correct* set of stored data! The Agilent N4000-series 'Smart Noise Sources' have individual calibration data stored within the source itself, and these data are automatically communicated to the companion NFA-series analyser. The N4000 series also measure their own internal temperature, and communicate that to the analyser as well. As already noted, these features have eliminated many common sources of user error with older analysers.

The N4000-series sources have detailed uncertainty data for each calibration point, but for the purpose of constructing an error budget you can simply assume an ENR uncertainty of ± 0.15 dB. The HP346A and B have no equivalent detailed data but you can reasonably assume ± 0.2 dB.

Unfortunately the trend towards lower-noise devices means that the majority of the commercial noise sources coming onto the surplus market are 15 dB ENR types like the HP346B. It is possible to make a 5 dB source by attaching a high quality 10 dB attenuator to the output of an existing 15 dB source – but you cannot casually subtract 10 dB from the published ENR and expect accurate results! There are many corrections to be applied, requiring detailed measurements with a Vector Network Analyser. As well as a very accurate measurement of the attenuation value in a 50 Ω reference environment, you also need the vector reflection coefficients Γ_{ON} and Γ_{OFF} for the noise source, and Γ_{IN} and Γ_{OUT} for the attenuator. Then you must grind through the calculations… and at the end of it all, the uncertainty in ENR will be worse than that of the bare noise source.

A purpose-made 5 dB noise source like the HP346A will have that attenuator built in – but more importantly, the ENR uncertainty will be the same as for the 15 dB model. Therefore it is better to use a purpose-made 5 dB source if at all possible.

And one more question: even if you do own a factory calibrated noise source, how many years ago was that calibration done?

Total Uncertainty Budget

Fire-up the uncertainty calculator on the Agilent website. We recommend the online Java version[10] because it gives useful guidance about input values that are instrument-specific. Now plug in the numbers for your system… and find out how little you can **really** trust your results.

Here are two examples:

1. **DUT NF 0.3 dB, DUT gain 30 dB**

   A modern LNA design with input return loss **10** dB, output return loss 10 dB

   Measurement uncertainty:

   ± 0.29 dB (HP346A, 8970B)

   ± 0.23 dB (N4000A, N8973A)

2. **DUT NF 0.3 dB, DUT gain 30 dB** – same NF and gain as case 1, but…

   A 1980s-style LNA design with input return loss **1** dB, output return loss 10 dB

   Measurement uncertainty:

   ± 0.60 dB (HP864A, 8970B)

   ± 0.51 dB (N4000A, N8973A)

10.http://sa.tm.agilent.com/noisefigure/NFUcalc.html

We can see three major conclusions:

1. The modern-style LNA has measurement uncertainties comparable with the true NF. Most of this (about 0.2 dB) comes from the ENR calibration. With care, other errors can be held down to about ± 0.1 dB.
2. The old-style tuneable LNA has 2x the measurement uncertainties, and most of that comes from the highly mismatched input circuit. That uncertainty is about 2 times larger than the true NF, so the measurement is not meaningful. Worse still, it isn't even possible to find the optimum input settings.[11]
3. The modern N4000A/N8973A lineup offers a little lower uncertainty than the older HP346A/8970 but it does not cure any of the problems that we have been discussing.

With modern LNAs using devices that have extremely low noise figures, conventional NF measurements are coming close to the end of the road. There are too many sources of uncertainty that simply cannot be avoided.

However, **that should not prevent us from minimizing the <u>avoidable</u> uncertainties**, not only by careful measurement technique but also by avoiding LNA designs that increase those measurement uncertainties even more.

The Roads Less Travelled

Although the conventional noise source and NFA technique has not yet reached the end of the road, that roadblock can clearly be seen ahead of us. It is inherent in the underlying physics.

So perhaps it is time to revisit some of the 'roads less travelled' in NF measurement. This is always a good policy in the Test and Measurement business, because changing technology brings new solutions as well as new challenges. As amateurs we also need to be on the lookout for techniques that we can use but professionals couldn't – for example, we are generally interested only in measurements in specific narrow bands.

CW Techniques

This is the standard technique for HF sensitivity measurements, using a CW test signal and SNR meter. It was superseded by the techniques using noise for the test signal because of the difficulties in computing the equivalent noise bandwidth of the DUT and in measuring the power levels of two very different signals – carrier and noise – with equal accuracy. A particular difficulty with using CW techniques at extremely low signal levels is to prevent direct leakage from the signal generator into the DUT. Although CW techniques could be revisited using DSP to measure bandwidths and power levels, the leakage problem remains. The verdict is that CW techniques are unlikely to offer much new promise when subjected to the same detailed uncertainty analysis as described above.

Hot/cold thermal source

This is a version of the Y-factor method using a 50 Ω resistor as a thermal noise source, and physically transferring the same resistor between a hot and cold environment. In effect, this is emulating the techniques of a national standards lab, twice over. The longer the 'temperature baseline' between T_{HOT} and T_{COLD}, the greater the differences will be between the respective noise power measurements, both during

11. Do not misquote us on this! We did **NOT** say or wish to imply that 1980s-style LNAs "don't work". But we **do** say that with such large uncertainties, measurements on those LNAs using a conventional noise source and NFA cannot be *meaningful*.

calibration and with the DUT, so physically hot and cold thermal sources do offer some potential for more accurate measurements of NF and gain.

The practical difficulties with these techniques are to maintain the respective temperatures and to account for the temperature-dependent noise contributions from the cables and connectors between the source and the DUT. Another very important requirement is that the source impedance offered to the DUT does not change[12] between the two temperatures (exactly the same difficulty that we saw with a conventional diode noise source). Correcting for the uncertainties in this technique will require extensive use of a Vector Network Analyser.

But this hot/cold transfer technique has another fatal flaw. It is very slow, which makes it useless for real-time optimization and also makes measurements very vulnerable to drift while the source temperature is being changed.

Switching between hot and cold thermal sources

The slow changeover in the physical temperature of a single thermal noise source can be avoided by switching between two different sources – but this introduces new uncertainties due to the differences between the two sources, and in the switching mechanism itself. Once again, this is equivalent to emulating a national standards lab, twice over, and a VNA will be a mandatory accessory. And again, unless the source switching is fast and reproducible, this technique will again be too slow for real-time optimization.

Switching with extended T_{HOT}—T_{COLD} baseline

This is an interesting variant on the technique using switched sources, and once again it aims to extend the baseline between the T_{HOT} and T_{COLD} equivalent noise temperatures. This can provide much more favourable numbers in the calculations of noise temperature as illustrated in Figures 2 and 3and NF, and while it cannot eliminate all the other kinds of error and uncertainty identified earlier in this paper, an extended temperature baseline can make them relatively less important.

The avalanche diode noise source is already extending T_{HOT} upward beyond the physical temperature of the diode… but how about extending T_{COLD} downward? If we can move T_{COLD} closer to the origin (0, 0) in Figure 2 or Figure 3, this will reduce the downward extrapolation required to find T_{DUT}, and so will reduce the sensitivity to non-linearity and errors in the slope of the line.

Two methods have come to notice. At this Conference, DF1VH reports a method using the non-thermal properties of Schottky diode in forward conduction at room temperature, to create a ‘cold’ termination with an equivalent noise temperature of about 180 K (multimedia session, full details on the Conference DVD). This reduces the sensitivity to gain errors by a factor of almost 2 compared with a conventional diode source where T_{COLD} is about 300 K. This could be a worthwhile improvement, although it introduces new complications in the need to switch between ‘hot’ and ‘cold’ sources, and the need to characterize two different non-thermal sources in terms of their equivalent noise temperature.

An alternative non-thermal ‘cold’ reference was described by RW3BP at the Dallas EME Conference in 2008. By pointing a low-noise horn straight upward, and avoiding unfavourable times when the horn is pointing towards the Milky Way (Cas A) and other known sources, RW3BP obtained an effective value for T_{COLD} of about 14 K. This re-

12. As with all of the techniques described here, it is not necessary for the noise source impedance to be precisely 50Ω; it is only necessary for it to remain *the same* between the HOT and COLD states of the noise source.

quires a very much shorter extrapolation than any other type of 'cold' load, outside of a cryogenics lab. For the 'hot source' part of the measurement, RW3BP used a high-ENR avalanche diode source with an attenuator and a directional coupler, giving an effective value of about 80 K for T_{HOT}. The design compromise here is between extending the temperature baseline and controlling the small change in source impedance when the noise source is switched on.

RW3BP's technique has several features that would make it completely unacceptable in a commercial Test and Measurement applications: narrowband frequency coverage, dependence on the time of sidereal day, and the need for a considerable amount of installation work – but none of those is necessarily a problem for amateurs.

The techniques by DF1VH and RW3BP are well worth exploring in more detail. A characteristic of these 'roads less travelled' is that access to a Vector Network Analyzer will almost certainly become an essential requirement, in order to evaluate the corrections required.

The only way to evaluate the real improvements over conventional methods is to subject new techniques to the same rigorous uncertainty analysis as the ones we already know.

Conclusion

We already know the disadvantages of conventional NF measurement methods, and we can also see that conventional methods are nearing the end of the road. Achievable values of NF are already lower than the uncertainties in our measurement techniques… especially if we allow ourselves forget that those uncertainties even exist!

Experienced EMEers are already witnessing the next generation repeating mistakes that we already made. We need to take some responsibility for that, and make more efforts to retain this information in the pool of common knowledge.

It is also time to explore other methods of NF measurement, and to subject them to uncertainty analysis using the techniques that we already have.

References

All references are on the Conference DVD (except where noted).

J C Greene, 'Noisemanship'. The art of false NF measurement was documented no less than 50 years ago! The original article was meant as a joke, but the same mistakes are still being made.

'Fundamentals of RF and Microwave Noise Figure Measurements',
Agilent Application Note 57-1.

'Noise Figure Measurement Accuracy – The Y-Factor Method'
Agilent Application Note 57-2.

Rainer Bertelsmeier DJ9BV, 'How to Use a Noise Figure Meter', *DUBUS* 1990/04.

Rainer Bertelsmeier DJ9BV, 'Low Noise GaAsFET Preamps for EME – Construction and Measurements Problems', *DUBUS* 1988/04.

Rainer Bertelsmeier DJ9BV, 'Myths and Facts about Preamp Tuning',
in the *DUBUS TECHNIK V* compilation, available from http://www.dubus.org

Ingo Gaspard DF1VH, 'Solid State Broadband Un-Cooled Noise Generator with Noise Temperature Below Room Temperature'.

Sergey Zhutyaev RW3BP, '1296 MHz Small EME Station with Good Capability,
Part 4: LNA Optimization',
http://www.vhfdx.ru/apparatura/rw3bp_1296mhz_lna_optimization