Accurate Absolute and Relative Power Measurements Using the Agilent N5531S Measuring Receiver System

Application Note

Agilent Technologies

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Introduction

The Agilent N5531S measuring receiver system is the successor to the venerable (and now discontinued) 8902A measuring receiver. The center piece of the N5531S is an optional firmware personality module for the Agilent PSA Series spectrum analyzers (Option 233). The N5531S measuring receiver system matches or outperforms the 8902A in every operating mode, while extending the fundamental operating frequency range from 1.3 GHz (8902A) to as high as 50 GHz.

Accurate power-level measurement was a key capability in the 8902A that has been carried over to the N5531S. This capability is essential for the calibration of signal-generator output level and step-attenuator accuracy.

Among the more important performance parameters for any signal generator are the absolute and relative accuracies of its output power level. The highest-accuracy measurement of the output level of a signal generator requires a power meter and sensor. However, modern power sensors can measure only part of the output range of modern signal generators. Below –70 dBm, some other method must be used. The traditional approach has been to check the relative accuracy of the generator's output level for power levels below the range of power sensors. The power meter is employed to establish a reference power level, and a sensitive, linear measuring receiver is used to verify level accuracy relative to that reference down to –140 dBm.

N5531S
Measuring Receiver System

As shown in Figure 1, the N5531S measuring receiver system consists of an E444xA PSA Series high-performance spectrum analyzer (with measuring receiver personality, Option 233), a P-Series power meter (N1911A or N1912A), and an N5532A sensor module.

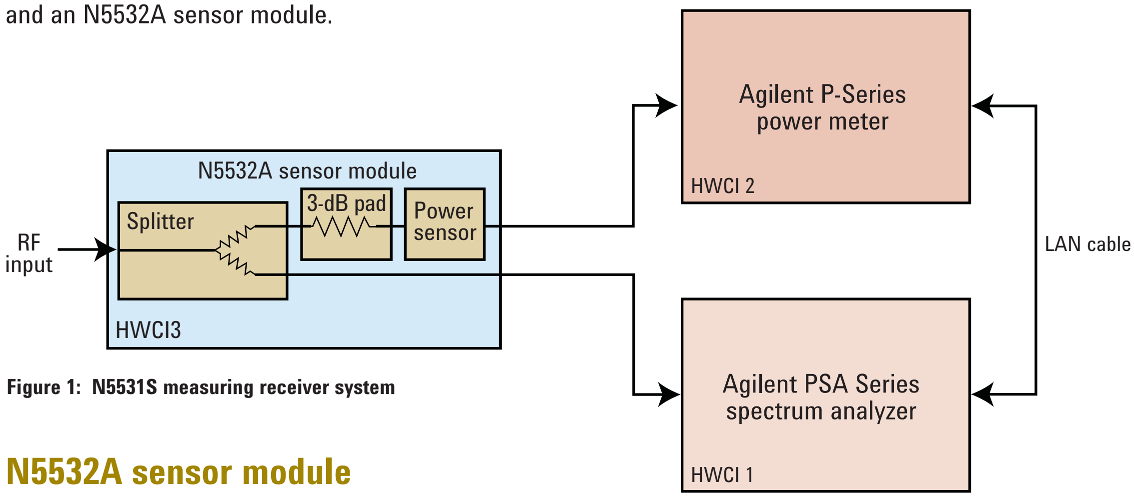


Figure 1: N5531S measuring receiver system

N5532A sensor module

The sensor module includes a low-frequency, multi-signal cable between the power sensor output and the power meter, and a coaxial RF cable between the sensor module and the spectrum analyzer input. Figure 2 shows the internal block diagram of the N5532A sensor module.

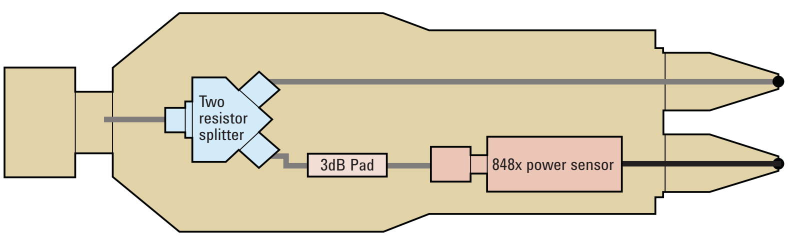


Figure 2: N5532A sensor module

Two-resistor power splitter

The details of the two-resistor power splitter, especially why it was chosen over a three-resistor power splitter, are covered in Appendix C.

3 dB pad

At the expense of 3 dB of signal loss, this pad improves the match between the output of the power splitter and the input of the power sensor. It does this by attenuating the reflected voltage waves moving in either direction (toward or away) from the power sensor.

Power sensor

The N5532A sensor module contains an Agilent 848XA power sensor.

Table 1. N5532A Option map

N5532A Option	Frequency range	Sensor used
504	100 kHz – 4.2 GHz	8482A
518	10 MHz – 18 GHz	8481A
526	30 MHz – 26.5 GHz	8485A
550	30 MHz – 50.0 GHz	8487A

The power measurement range of the 848XA sensor family is +20 dBm to –30 dBm. The (nominal) 7 dB of loss provided by the power splitter and 3 dB pad inside the N5532A offsets the measurement range of the N5532A to +30 dBm to –20 dBm.

The 848XA power sensor family incorporates thermocouple technology. Thermocouple sensors have been the technology of choice for sensing RF and microwave power since their introduction in 1974. This is primarily because of their:

1. higher sensitivity, as compared to thermistor-based sensors
2. higher accuracy, as compared to diode-based sensors
3. inherent square-law detector characteristic (input RF power is proportional to DC voltage out)

Since thermocouples convert RF energy into heat, they are true "averaging detectors." This allows them to accurately measure the power of all types of signals from continuous wave (CW) to complex digital phase-modulated waveforms. In addition, they are more rugged than thermistors, measure power levels as low as 1.0 µW (–30 dBm, full scale), and they have lower reflection coefficients resulting in lower values of mismatch loss.

Although the conversion of RF-input power to DC-output voltage is highly linear over most of a thermocouple's measurement range, the DC-output voltage varies non-linearly with the average temperature of the thermal element. At high power, the average thermocouple temperature increases, increasing the output voltage by a larger factor than predicted by linearly-extrapolating data from lower power levels. At a power level of 30 mW, the DC output is 3% higher than predicted by the low-level value. At 100 mW, the output is about 10% higher. Circuitry inside the power meter compensates for this non-linear temperature-driven power-level effect. Circuitry inside the sensor itself compensates for changes in its ambient temperature.

The outstanding linearity of the 848XA sensors is reflected in their specifications.

Table 2. 848XA Power sensor linearity
Linearity for –30 dBm to +20 dBm

	Sensor model number			
Power range	**8481A**	**8482A**	**8485A**	**8487A**
–30 dBm < P < +10 dBm	~0	~0	~0	~0
+10 dBm < P < +20 dBm	±3.0%	±3.0%	±3.0%	±3.0%

The principles and application of thermocouple power sensors are discussed in: *Fundamentals of RF and Microwave Power Measurements (Part 2): Power Sensors and Instrumentation*, AN 1449-2, literature number 5988-9214EN.

Calibrating (i.e., characterizing) the N5532A sensor module is almost as simple as calibrating an 848XA power sensor. By terminating the spectrum analyzer output of the module with a precision, 50 Ω load, the module becomes a one-port device, behaving essentially as an attenuated power sensor. The path through the series connection of the power splitter/3 dB pad/848XA sensor is calibrated as if it were a regular power sensor. The resulting calibration data includes:

- calibration factor
- calibration factor uncertainty
- reflection coefficient magnitude
- reflection coefficient-magnitude uncertainty
- reflection coefficient phase

This data is provided on a floppy disc so that it can be stored in the PSA memory as a correction file to be used by the N5531S measuring receiver personality. The calibration factor uncertainty is combined with the other power sensor and power meter uncertainties to determine the overall uncertainty of a specific power measurement.

Functional summary

Table 3: N5531S measurement suite

Measurement	N5531S component
Frequency counter	PSA spectrum analyzer, Option 233
AM depth	PSA spectrum analyzer, Option 233
FM deviation	PSA spectrum analyzer, Option 233
PM deviation	PSA spectrum analyzer, Option 233
AM/FM/PM modulation rate	PSA spectrum analyzer, Option 233
AM/FM/PM modulation distortion	PSA spectrum analyzer, Option 233
Tuned-RF level — absolute	N5532A plus PSA spectrum analyzer, Option 233, P-Series power meter
Tuned-RF level — relative	PSA spectrum analyzer, Option 233
Absolute-RF power	N5532A plus PSA spectrum analyzer, Option 233, P-Series power meter
Audio: Frequency, AC level, and distortion	PSA spectrum analyzer with Option 107 and Option 233

RF Power Measurement

In the RF power mode, the N5531S measures absolute power in the range –10 to +30 dBm via the N5532A sensor module and the P-Series power meter. The accuracy of this measurement mode is dependent on power level and frequency as indicated by the specification table below. Please refer to the "Measuring Receiver Personality" chapter of the latest "Specifications Guide" for the Agilent PSA Series spectrum analyzer for the complete and current specifications.

Table 4: N5531S RF power measurement

Power meter range	Power level (dBm)	Frequency range	Absolute RF power accuracy**	N5532A option
1	+20 to +30	100 kHz – 4.2 GHz	±0.356 dB	504
1	+20 to +30	4.2 – 18.0 GHz	±0.400 dB	518
1	+20 to +30	18.0 – 26.5 GHz	±0.387 dB	526
1	+20 to +30	26.5 – 50.0 GHz	±0.420 dB	550
2 – 4	–10 to +20	100 kHz – 4.2 GHz	±0.190 dB	504
2 – 4	–10 to +20	4.2 – 18.0 GHz	±0.267 dB	518
2 – 4	–10 to +20	18.0 – 26.5 GHz	±0.380 dB	526
2 – 4	–10 to +20	26.5 – 50.0 GHz	±0.297 dB	550

** Mismatch effects between the DUT and the N5532A sensor are NOT included in the specifications given in this table, because of the dependence on the DUT SWR. Other mismatches are under control because of the two-resistor splitter (see Appendix C), the effects of mismatch between the spectrum analyzer and the sensor module cause negligible changes in the relationship between the output voltage of the DUT and the power sensor measurement results. Thus, this effect is included in this absolute RF power accuracy specification column. The mismatch presented by the spectrum analyzer does affect the impedance seen at the input of the sensor module. This effect is included in the system "Input SWR" specification.

Example: Verifying the power-level accuracy of a signal generator

E4428C signal generator specification:

For the power output range: +15 to –100 dBm,

Power level accuracy: **±0.5 dB,** for 250 kHz < f ≤ 2.0 GHz
±0.6 dB, for 2.0 GHz < f ≤ 3.0 GHz

Because the power range covered by the N5531S RF power level measurement is limited to a lower level of –10 dBm, the signal generator will be set to an output level of at least –10 dBm.

Measurement procedure

1. Connect and configure the measuring system
2. Calibrate and zero the power meter
3. Select "Measuring Receiver" mode on the PSA
4. Set the signal generator frequency to 2.0 GHz
5. Set the signal generator power level to –10 dBm
6. Press the "RF Power" soft key on the PSA front panel
7. Wait for the N5531S to return a stable reading (e.g., –10 dBm)

Applying the specification

A specification is a quantitative statement about the range of values, centered on the indicated value, within which the true value of some performance parameter lies, with a stated level of statistical confidence (typically 95%). When applied to a measurement, a specification essentially becomes an uncertainty for that measurement.

RF power level accuracy is defined as:

$$ACC_{RF}[\mathrm{dB}] = P_{ACTUAL} - P_{IND}$$

This equation applies both to the E4428C signal generator and to the N5531S measuring receiver.

For the E4428C (assuming: 250 kHz ≤ f ≤ 2.0 GHz), the specified accuracy is:

$$-0.5\ \mathrm{dB} \le [P_{ACTUAL} - P_{IND}]_{SG} \le +0.5\ \mathrm{dB}$$

P_{ACTUAL} = the true power available at the output of the signal generator

P_{IND} = the power level indicated by the signal generator

For the N5531S with the N5532A sensor module option 504, the specified accuracy is:

$$-0.19\ \mathrm{dB} \le [P_{ACTUAL} - P_{IND}]_{MR} \le +0.19\ \mathrm{dB}$$

P_{ACTUAL} = the true power level present at the N5532A input connector

P_{IND} = the power level indicated by the N5531S

Alternatively:

$$P_{IND} - 0.19\ \mathrm{dB} \le [P_{ACTUAL}]_{MR} \le P_{IND} + 0.19\ \mathrm{dB}$$

This says that the actual power present at the input to the N5531S lies within ±0.19 dB of the power level indicated by the N5531S. To understand how this value of ±0.19 dB was derived, see Appendix A.

Note that this specification does not include the effect of mismatch.

Ignoring mismatch for the moment, and equating the actual power out of the signal generator to the power incident on the N5532A sensor module:

$$[P_{ACTUAL}]_{SG} = [P_{ACTUAL}]_{MR}$$

In terms of the equations shown above, this leads to:

$$[ACC_{RF}]_{SG} = [ACC_{RF}]_{MR} + [P_{IND-MR} - P_{IND-SG}],\ \text{in dB}$$

Note that $[ACC_{RF}]_{MR}$ is the uncertainty of the measurement made by the N5531S

Inserting the numerical values listed above:

$$[ACC_{RF}]_{SG} = [\pm 0.19 + (P_{IND-MR} - P_{IND-SG})],\ \text{in dB}$$

Tuned RF Level Measurement

In the N5531S, Tuned RF Level (TRFL) is measured by the PSA spectrum analyzer.

Specifications

Absolute TRFL accuracy is specified relative to an absolute power level reference established by the sensor module and power meter components of the N5531S. The resulting measured value depends on the specifications of the power sensor, power meter, and PSA.

Table 5. N5531S absolute TRFL measurement accuracy

Power level range	Absolute TRFL accuracy [dB]
Pre-amp OFF	
+20 dBm to maximum power	±(Power meter range 1 uncertainty + 0.005 dB/10 dB step)
Residual noise threshold power to +20 dBm	±(Power meter range 2 uncertainty + 0.005 dB/10 dB step)
Minimum power to residual noise threshold	±(Cumulative error[Note1] + 0.0012[Note3]·(Input power – residual noise threshold power)^2)
Pre-amp ON	
Residual noise threshold to +16 dBm	±(Power meter range 2 uncertainty + 0.005 dB/10 dB step)
Minimum power to residual noise threshold	±(Cumulative error[Note2] + 0.0012[Note3]·(Input power – residual noise threshold power)^2)
Definition	
Residual noise threshold	Minimum power + 30 dB

Notes:

1. In absolute TRFL measurements, the "cumulative error" is the error incurred when stepping from a higher power level to the residual noise threshold power level. The formula to calculate the cumulative error is ±(0.190 dB + 0.005 dB/10 dB step). For example, assume the higher level starting power is 0 dBm and the calculated residual noise threshold power is –99 dBm. The cumulative error would be ±(0.190 + [99/10] x 0.005 dB), or ±0.240 dB. The multiplier in brackets is converted to the smallest integer that is not less than the value in brackets.
2. In absolute TRFL measurements, the "cumulative error" is the error incurred when stepping from a higher power level to the residual noise threshold power level. The formula to calculate the cumulative error is ±(0.356 dB + 0.005 dB/10 dB step). For example, assume the higher level starting power is 0 dBm and the calculated residual noise threshold power is –99 dBm. The cumulative error would be ±(0.356 + [99/10] x 0.005 dB), or ±0.406 dB. The multiplier in brackets is converted to the smallest integer that is not less than the value in brackets.
3. The units of absolute TRFL accuracy are decibels. The units of input power and residual noise threshold power are both dBm. The units of their difference is thus decibels. The units of the square of their difference is thus decibels-squared. Therefore, the units of this constant are inverse decibels.

Relative TRFL accuracy is specified relative to an arbitrary power level selected by the spectrum analyzer. This accuracy depends solely on the specifications of the PSA.

Table 6: N5531S relative TRFL measurement accuracy

Power level range	Relative TRFL accuracy [dB]
Residual noise threshold to maximum power	±(0.015 + 0.005 dB/10 dB step)
Minimum power to residual noise threshold	±(Cumulative error [Note 1] + 0.0012 [Note 2]· (Input power – residual noise threshold power)^2)
Range 2 uncertainty	±0.031 dB
Range 3 uncertainty	±0.031 dB
Residual noise threshold power	Minimum power + 30 dB

Notes:
1. In relative TRFL measurements, the cumulative error is the error incurred when stepping from a higher power level to the residual noise threshold power level. The formula to calculate the cumulative error is ±(0.015 dB + 0.005 dB/10 dB step). For example, assume the higher level starting power is 0 dBm and the calculated residual noise threshold power is –99 dBm. The cumulative error would be ±(0.015 + [99/10] x 0.005 dB), or ±0.065 dB. The multiplier in brackets is converted to the smallest integer that is not less than the value in brackets.
2. The units of relative TRFL accuracy are decibels. The units of input power and residual noise threshold power are both dBm. The units of their difference is thus decibels. The units of the square of their difference is thus decibels-squared. Therefore, the units of this constant are inverse decibels.

Supporting information for specifications

In order to use the accuracy specification, the minimum power and residual noise threshold information is needed. The relationship between these levels is given in both Table 5 and Table 6. A subset of information from the specifications guide is given in Table 7.

Table 7: Minimum power

Frequency range	RBW	Preamplifier	
		Uninstalled	Installed
0.01 – 3.05 GHz	10 Hz	–136 dBm	–140 dBm

Range-to-range cal factor correction

The PSA measures the tuned-RF-level in three distinct ranges, as shown in Figure 3 below. This approach allows the PSA to maintain a favorable signal-to-noise ratio as the signal level decreases toward the noise floor of the PSA. The actual range switching involves changing the setting of the internal attenuator and either including or excluding the internal preamp of the PSA. See Table 7.

Table 8: N5531S TRFL measurement ranges

Range	PSA atten (dB)	Pre-amp	Switch point** (dBm)
1	30	OFF	0
2	10	OFF	–58
3	4	ON	–78

** Example values. Actual switch point power level depends on the signal-to-noise ratio.

CalFactor1 (see Figure 3) is simply the ratio of the power measured by the N5531S power meter to the power indicated by the PSA in range 1. Subsequent measurements made in range 1 are automatically corrected by the PSA to include CalFactor1. If a relative TRFL measurement is being made, then CalFactor1 is 0 dB.

CalFactor2 is the ratio of the power indicated by the PSA in range 1 to the power indicated in range 2 at the switch-point between range 1 and range 2. The power level is held constant while the range is switched and the measurements are made. Subsequent measurements made in range 2 are automatically corrected by the PSA to include CaFactor1 plus CalFactor2.

CalFactor3 is the ratio of the power indicated by the PSA in range 2 to the power indicated in range 3 at the switch-point between range 2 and range 3. The power level is held constant while the range is switched and the measurements are made. Subsequent measurements made in range 3 are automatically corrected by the PSA to include CalFactor1, CalFactor 2, and CalFactor 3.

The uncertainty of the values of the three range-to-range cal factors is derived in Appendix B.

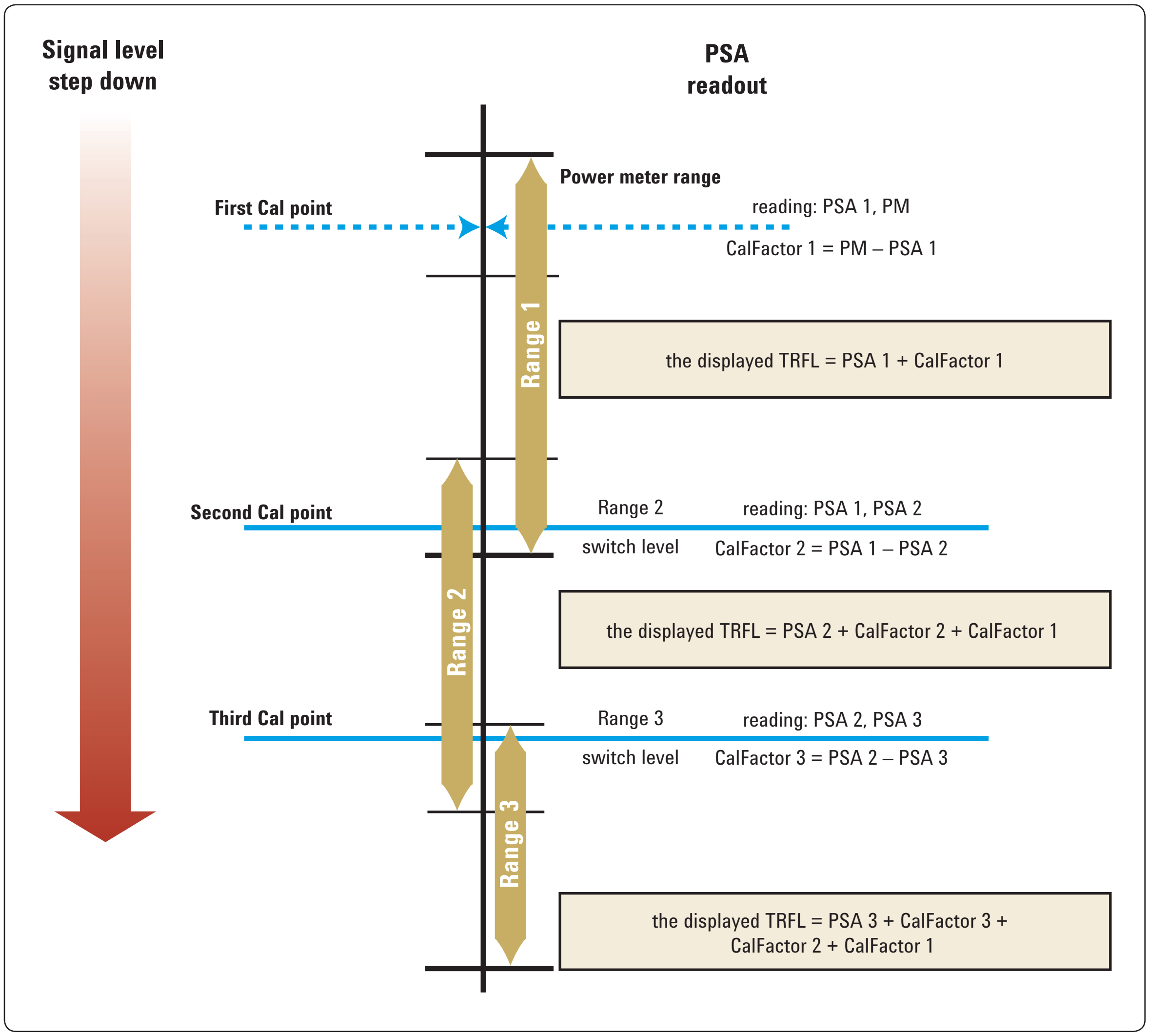


Figure 3. N5531S TRFL measurement ranges

Absolute TRFL measurement

Making an absolute power measurement using TRFL mode requires "calibrating" the PSA with a known, absolute power level via the N5532A sensor module and P-Series power meter. Then, the power level to be measured is presented to the PSA. The PSA will automatically set itself to the correct measurement range and will calculate the appropriate range-to-range cal factors.

Example: Measuring a low power level (–100 dBm) from an Agilent E4428C signal generator.

Measurement procedure

1. Set the signal generator to 2.0 GHz and 0 dBm
2. Select "Measuring Receiver" mode on the PSA
3. Set the PSA center frequency to 2.0 GHz
4. Select "Tuned-RF-Level" mode on the PSA
5, Wait for the PSA to achieve a stable reading near 0.0 dBm
6. Decrease the signal generator output power level in 10 dB steps until –100 dBm is reached
7. The PSA will indicate a value of approximately –100 dBm for the signal level. (This is the value for the signal and noise.)

Applying the specification

Amplitude accuracy in absolute TRFL mode is defined as:

$$[ACC_{TRFL}]_{ABS} = P_{ACTUAL} - P_{IND} = ACC_{PM} + ACC_{TRFL}$$

where:

P_{ACTUAL} = the true power present at the N5532A sensor input in dBm

P_{IND} = the power indicated by the PSA, in dBm

ACC_{PM} = accuracy of the power reading taken by the N5532A and power meter

ACC_{TRFL} = accuracy of the PSA in making the differential measurement between the reference power (established by the power meter) and the power level under test

First, we'll have to look up the minimum power. Table 7 shows that, with the preamp installed and using the 10 Hz resolution bandwidth setting, it is –140 dBm (–136 dBm without the preamp). Next, note that the residual threshold power is defined to be 30 dB higher, or –110 dBm (–106 dBm without the preamp). The power in this example, –100 dBm, exceeds the residual threshold power with or without the preamp. In Table 5, we can choose either case, and the accuracy expression is the same:

±(power meter range 2 uncertainty + 0.005 dB/10 dB step)

Looking at Table 4, we see, in power meter range 2, with the N5532A Option 504, the accuracy is ±0.190 dB. The measurement procedure began with a calibration using a 0 dBm reference. Our measurement is of a power level of –100 dBm, which is 100 dB lower, or 10 steps of 10 dB each. Therefore, the term "0.005 dB per 10 dB step" evaluates to ±0.05 dB. Adding this to the ±0.190 dB discussed above, we find:

$$[ACC_{TRFL}]_{ABS} = \pm 0.240 \text{ dB}$$

Relative TRFL measurement

Example: Accuracy verification of an Agilent 8496G programmable step attenuator

Frequency range:	250 kHz to 4.00 GHz
Attenuation range:	0 to 110 dB

The accuracy of a step-attenuator is usually established by comparison to a laboratory-standard attenuator whose accuracy is traceable to NIST. In the following test, the N5531S will be used in place of the reference attenuator as an attenuation standard.

For this example measurement, the N5531S will employ the relative TRFL function. For optimum signal-to-noise ratio, we'll not use the N5532A.

Measurement procedure

1. Set the signal generator to the frequency required for the attenuator verification measurement
2. Insert a 3 dB pad at the input and output of the step attenuator under test. (This will minimize the amplitude variation due to changes in mismatch as the attenuator is switched from setting to setting.)
3. Set the signal generator power level to 0 dBm
4. Select “Measuring Receiver” mode on the PSA
5. Select FREQ mode on the PSA and manually set the PSA center frequency to the signal generator frequency
6. Select “TRFL” mode on the PSA
7. Wait for the PSA to achieve a stable reading
8. Press the “Set Ref” key to perform a relative TRFL measurement
9. Step the attenuator through its range and record the value indicated by the PSA at each step. The PSA automatically determines the appropriate range change cal factors as the power level passes through the range 2 and range 3 threshold values

Applying the specification

Let’s start by assuming a test frequency of 1 GHz. As in the previous example, we’ll first look up the minimum power. We’ll be using the preamp, so the minimum power is –140 dBm. The residual threshold power is again 30 dB higher at –110 dBm. The N5532A is not in-circuit but the two 3 dB pads on our attenuator-under-test affect the signal level. It varies from –6 dBm with the attenuator set to its reference 0 dB setting to –116 dBm. The final attenuator setting (110 dB) is measured at a power level below the residual threshold power; all other settings are measured at power levels higher than that.

From Table 8, we see that the ranges switch at –58 and –78 dBm at the input of the PSA, which corresponds to attenuations of 52 and 72 dB. Therefore, attenuations of 10, 20, 30, 40 and 50 dB are measured in Range 1, 60 and 70 dB are measured in Range 2, and 80, 90, 100 and 110 dB are in Range 3.

The relative accuracy for attenuations of 10 through 50 dB are given by the first row of Table 6. It is ±(0.015 + 0.005 dB/10 dB step). The number of 10 dB steps is, of course, 1 through 5 respectively. Therefore, the accuracy is:

Attenuation	Accuracy
10 dB	±0.020 dB
20	±0.025
30	±0.030
40	±0.035
50	±0.040

The relative accuracy for attenuations of 60 and 70 dB include the Range 2 uncertainty. Table 6 shows this is ±0.031 dB. This adds to the accuracy in the first row of that table (with 6 and 7 10 dB steps) to give:

Attenuation	Accuracy
60 dB	±0.076 dB
70	±0.081

For attenuations of 80 through 100 dB, we experience the sum of Range 2 and Range 3 uncertainties, at ±0.031 dB each, giving:

Attenuation	Accuracy
80 dB	±0.117 dB
90	±0.122
100	±0.127

Finally, for the 110 dB setting, we have to use the second row of Table 6. We find the cumulative error as we did in the last attenuation: ±0.015 dB, plus 11 steps of ±0.005 dB each for ±0.055 dB, plus Range 2 uncertainty of ±0.031 dB, plus Range 3 uncertainty, another ±0.031 dB, for a subtotal of ±0.132 dB. To this we add the other term. The input power is –116 dBm; the residual threshold power is –110 dBm. We find the difference (6 dB), square it (36 dB^2), and multiply it by 0.0012 dB^{-1}, for another ±0.043 dB error due to noisiness. The total:

Attenuation	Accuracy
110 dB	±0.175 dB

Note that all these accuracy figures are based on worst-case addition of error sources; root-sum-square additions will give lower uncertainty estimates.

Appendix A: Derivation of Absolute RF Power Accuracy

Measurement equation

An RF power measurement with the N5531S is made with the N5532A sensor module and a P-Series power meter. The measurement path consists of one path through the power splitter, the 3 dB pad, the power sensor, and the P-Series power meter. This connection is shown in Figure 2, on page 3.

The measurement made by just a power sensor and power meter is described by:

$$P_{gZ0} = \frac{M_S P}{IKLm}\text{, in Watts}$$

$$P = P_m - P_Z - P_N - P_D$$

$$m = \frac{M_{SC} P_C}{K_C P_{CAL}}$$

$$P_C = P_{mC} - P_{ZC} - P_{NC}$$

where:

P_{gZ0} is the power that would be delivered to a Z_0 (matched) load

M_S is the mismatch factor at the power sensor to 3 dB pad interface

M_{SC} is the mismatch factor at the calibrator output-to-sensor interface

P_m is the power indicated (reported) by the power meter

P_Z is the power offset due to zero error of the power meter

P_N is the power due to noise in the sensor and power meter

P_D is the drift of the power meter after it is zeroed and calibrated

P_{CAL} is the calibrator power level

K is the power sensor calibration factor

L is the linearity of the power sensor/power meter combination

m is the gain factor that forces the power meter to indicate the calibrator power when the sensor is connected to the power meter's calibration port

I represents changes in the average value of m after calibration.

For a detailed derivation of this equation and corresponding uncertainty equation for a standard power meter measurement, follow the "ISO Guide to Uncertainty in Measurement", in the *Fundamentals of RF and Microwave Power Measurements, Application Note 1449-3*, literature number 5988-9215EN.

Adding a 3 dB pad and a power splitter to the power sensor/power meter combination gives the full path through the N5532A sensor module. The measurement equation becomes:

$$P_{IN} = \frac{M_{SS}\, M_{SP}\, M_{PS}}{IK_0 L m_0} \cdot P$$

where: P_{IN} is the power at the input connector of the N5532A sensor module

M_{SS} is the mismatch factor at the input of the power splitter

M_{SP} is the mismatch factor at the power splitter to 3 dB pad interface

M_{PS} is the mismatch factor at the 3 dB pad to power sensor interface

K_0 is the composite calibration factor of the splitter+pad+power sensor

m_0 is the composite gain factor that forces the power meter to indicate the calibrator power when the N5532A module is connected to the P-series power meter's calibrator port.

$$m_0 = \frac{K_{0-C}\, P_{CAL}}{M_{SS-C}\, M_{SP-C}\, M_{PS-C}\, P_C}$$

where: K_C is the composite calibration factor at 50 MHz (the calibrator frequency)

P_{CAL} is the P-Series power meter's calibrator power level (0 dBm)

Uncertainty equation

$$\frac{u^2(P_{IN})}{P_{IN}^2} = \frac{u^2(P_m) + u^2(P_N) + u^2(P_D)}{P^2} + \frac{u^2(I)}{I^2} + \frac{u^2(K_0)}{K_0^2} + \frac{u^2(L)}{L^2} + \frac{u^2(M_{SS})}{M_{SS}^2} + \frac{u^2(M_{SP})}{M_{SP}^2}$$

$$+ \frac{u^2(M_{PS})}{M_{PS}^2} + \frac{u^2(M_{SS-C})}{M_{SS-C}^2} + \frac{u^2(M_{SP-C})}{M_{SP-C}^2} + \frac{u^2(M_{PS-C})}{M_{PS-C}^2}$$

$$+ \frac{u^2(P_{mC}) + u^2(P_{NC})}{P_C^2} + \frac{u^2(K_{0-C})}{K_{0-C}^2} + \frac{u^2(P_{CAL})}{P_{CAL}^2} + \left(\frac{1}{P_C} - \frac{1}{P}\right) \cdot u^2(P_Z)$$

Determining standard uncertainties

- $\frac{u(P_m)}{P}$ = power meter resolution:

For this measurement, the power meter's measurement resolution is set to 0.01 dB. This means that the actual power level will fall somewhere within a 0.01 dB range. The actual value can be anywhere within this window and produce the same reading (to within 0.01 dB) on the power meter. Therefore, all values that fall within the 0.01 dB window are equally likely to produce the quantized readout value. This situation is best represented by a uniform distribution with a total range = ΔP_m = 0.01 dB.

For a uniform distribution, the standard deviation is given by:

$$\sigma_{RES} = \frac{\Delta P_m}{\sqrt{12}} = \frac{0.01 \text{ dB}}{\sqrt{12}} = 0.003 \text{ dB}$$

Since this value is in dB, it has the form

$$\frac{U(P_m)}{P_m} = \sigma_{RES} = \pm 0.003 \text{ dB}$$

Since $P_m >> P_n + P_Z + P_d$, $P \cong P_m$

Converting to a linear power ratio:

$$\frac{U(P_m)}{P} \cong \frac{U(P_m)}{P_m} = \left[10^{\frac{0.003}{10}} - 1\right] = \pm 0.0007 \text{ or } \pm 0.07\%$$

► $\frac{u(P_{mC})}{P_C}$ = Power meter resolution during calibration:

If the resolution is set to 0.01 dB, as in the case of P_m :

$$\sigma_{RES-CAL} = \frac{0.01\text{ dB}}{\sqrt{12}} = 0.003\text{ dB} = \frac{U(P_{mC})}{P_{mC}}$$

Converting to a linear power ratio:

$$\frac{U(P_{mC})}{P_{mC}} = \left(10^{\frac{0.003\text{ dB}}{20}} - 1\right) = 0.0007\ or\ 0.07\%$$

► $\frac{u(P_D)}{P}$ = uncertainty due to power meter drift:

For the N191XA power meter, drift is specified as ±10 nW at 2 GHz.

Example: For this test, the nominal power meter reading is -10 dBm (0.1 mW).

Assuming a uniform distribution for the power meter drift uncertainty:

$$\sigma_{DRIFT} = \frac{u(P_D)}{P} = \frac{\frac{10\text{ nW}}{\sqrt{3}}}{1\text{ mW}} = 5.773 \times 10^{-6}$$

► $\frac{u(P_N)}{P}$ = uncertainty due to power sensor noise (in watts):

For the 8482A power sensor coupled to the N191XA power meter, the measurement noise (free-run) is specified as 50 nW when averaging over a one-minute interval with averaging set to 1.

See the specifications chapter (pages 1-12) in the *N1911A/N1912A Service Guide*, Agilent document number: N1912-90015.

Assuming that the specification of 50 nW represents 95% (2σ) of a Gaussian distribution:

$$u(P_N) = \sigma_N = \frac{50\text{ nW}}{2} = 25\text{ nW}$$

$$\frac{u(P_N)}{P} = \frac{25\text{ nW}}{0.1\text{ mW}} = 250 \times 10^{-6} = 0.025\%$$

► $\frac{u(P_{NC})}{P_C}$ = uncertainty due to power sensor noise (watts) during meter calibration

P_{NC} is identical to P_N .

$$\frac{u(P_{NC})}{P_{NC}} = \frac{25\text{ nW}}{1.0\text{ mW}} = 25 \times 10^{-6} = 0.0025\%$$

► $\left(\frac{1}{P_C} - \frac{1}{P}\right) \cdot u(P_Z)$ = uncertainty due to power meter zero set (in watts):

For the 8482A, power meter zero set is specified as ±50 nW. Assuming a Gaussian distribution for this parameter and assuming that the specified value represents 95% (2σ) of the population of zero set values:

$$\sigma_Z = u(P_Z) = \frac{50\ \text{nW}}{2} = 25\ \text{nW}$$

For this test, the nominal power meter reading is –10 dBm (0.1 mW).

$$\left(\frac{1}{P_C} - \frac{1}{P}\right) \cdot u(P_Z) = \left(\frac{1}{1\ \text{mW}} - \frac{1}{0.1\ \text{mW}}\right) \cdot (25\ \text{nW}) = (-9000) \times (25 \times 10^{-9}) = -0.000225 = -0.023\%$$

where: $P = P_m - P_Z - P_N - P_D = 0.1\ \text{mW} - 50\ \text{nW} - 110\ \text{nW} - 10\ \text{nW} \cong 0.1\ \text{mW}$

$$P_C = P_{mC} - P_Z - P_{NC} = 1.0\ \text{mW} - 50\ \text{nW} - 110\ \text{nW} \cong 1.0\ \text{mW}$$

► $\frac{u(I)}{I}$ = uncertainty of the power meter instrumentation gain:

The uncertainty of the instrumentation gain is specified as ± 0.5%. Assuming a Gaussian distribution for this parameter and assuming that the specified value represents 95% (2σ) of the population of values:

Then: $\frac{u(I)}{I} = \frac{0.5\%}{2} = 0.25\%$

► $\frac{u(K_0)}{K_0}$ = uncertainty of the calibration factor of the composite N5532A sensor/splitter/ 3 dB pad (%):

There are two parts to this uncertainty factor:

1. The uncertainty of the value of the calibration factor, as supplied by the calibration facility.
2. The uncertainty due to the fact that the N5532A sensor module is calibrated with a precision 50 Ω load attached to the PSA output port, but is used with the actual PSA attached to that port. The difference in reflection coefficient between the two conditions causes an uncertainty in the actual power emerging from the power splitter and impinging on the 3 dB pad.

1. $[K_0]_{CAL}$ Specified value (from cal. Lab) = ±1.5%

For the N5532A, assuming a Gaussian distribution for this parameter, and assuming that the specified value represents 95% (2σ) of the population of values:

$$2 \cdot \sigma_{K_{CAL}} = 2 \cdot \frac{u([K_0]_{CAL})}{[K_0]_{CAl}} = 1.5\%$$

$$\frac{u([K_0]_{CAL})}{[K_0]_{CAL}} = \frac{1.5\%}{2} = 0.75\%$$

2. $[K_0]_{PSA}$ A signal flow graph analysis verified with experimental data, has shown that the uncertainty caused by the reflection of power from the PSA input back through the power splitter and through the 3 dB pad into the power sensor is approximately ±3.0%.

Assuming a Gaussian distribution for this parameter, and assuming that the specified value represents 95% (2σ) of the population of values:

$$2 \cdot \sigma_{K_{PSA}} = 2 \cdot \frac{u([K_0]_{PSA})}{[K_0]_{PSA}} = 3.0\%$$

$$\frac{u([K_0]_{PSA})}{[K_0]_{PSA}} = \frac{3.0\%}{2} = 1.5\%$$

► $\frac{u(K_{0-C})}{K_{0-C}}$ = uncertainty of the calibration factor of the composite N5532A sensor/splitter/3 dB pad (%) at the calibrator frequency (50 MHz):

$$2 \cdot \sigma_{K_{CAL}} = 2 \cdot \frac{u(K_{0-C})}{K_{0-C}} = 1.5\%$$

$$\frac{u(K_{0-C})}{K_{0-C}} = \frac{1.5\%}{2} = 0.75\%$$

► $\frac{u(L)}{L}$ = uncertainty due to power sensor linearity:

The linearity of a power sensor depends on the power level being measured.

Power sensor linearity

	8482A with N119XA meter	
	Spec. unc.	Std. unc.
P < +10 dBm	(negligible)	(negligible)
+10 dBm ≤ P ≤ +20 dBm	+2%, –4%	+1%, –2%

For this measurement, P = –10 dBm. Therefore,

$$\frac{U(L)}{L} = 0$$

► $\frac{u(M_{SS})}{M_{SS}}$ – uncertainty due to mismatch between the signal source and the N5532A input (assuming a source VSWR of 1.5:1)

$$\Gamma_{SG} = \frac{VSWR_{SG} - 1}{VSWR_{SG} + 1} = \frac{1.50 - 1}{1.50 + 1} = 0.20$$

$$\Gamma_{SPLITTER} = \frac{VSWR_{SPL} - 1}{VSWR_{SPL} + 1} = \frac{1.10 - 1}{1.10 + 1} = 0.048$$

$$\frac{u(M_{SS})}{M_{SS}} = \frac{\frac{|\Gamma_{SG}| \cdot |\Gamma_{PS}|}{\sqrt{2}}}{|1 - \Gamma_{SG}\Gamma_{PS}|^2} = 0.00692 = 0.69\%$$

► $\frac{u(M_{SS})}{M_{SS}}$ – uncertainty due to mismatch between the power splitter and the 3 dB pad

$$\Gamma_{SG} = \frac{VSWR_{SG} - 1}{VSWR_{SG} + 1} = \frac{1.50 - 1}{1.50 + 1} = 0.20$$

$$\Gamma_{SPLITTER} = \frac{VSWR_{SPL} - 1}{VSWR_{SPL} + 1} = \frac{1.10 - 1}{1.10 + 1} = 0.048$$

$$\frac{u(M_{SS})}{M_{SS}} = \frac{\frac{|\Gamma_{SG}| \cdot |\Gamma_{PS}|}{\sqrt{2}}}{|1 - \Gamma_{SG}\Gamma_{PS}|^2} = 0.00692 = 0.69\%$$

► $\frac{u(M_{PS})}{M_{PS}}$ – uncertainty due to mismatch between the 3 dB pad and the 848XA power sensor

$$\Gamma_{SENSOR} = \frac{VSWR_{SEN} - 1}{VSWR_{SEN} + 1} = \frac{1.05 - 1}{1.05 + 1} = 0.024$$

$$\Gamma_{PAD} = \frac{VSWR_{PS} - 1}{VSWR_{PS} + 1} = \frac{1.06 - 1}{1.06 + 1} = 0.029$$

$$\frac{u(M_{SP})}{M_{SP}} = \frac{\frac{|\Gamma_{SEN}| \cdot |\Gamma_{PAD}|}{\sqrt{2}}}{|1 - \Gamma_{SEN}\Gamma_{PAD}|^2} = 0.0005 = 0.05\%$$

► $\frac{u(M_{SP})}{M_{SP}}$ – uncertainty due to mismatch between the power splitter and the 3 dB pad

$$\Gamma_{SPLITTER} = \frac{VSWR_{SPL} - 1}{VSWR_{SPL} + 1} = \frac{1.10 - 1}{1.10 + 1} = 0.048$$

$$\Gamma_{PAD} = \frac{VSWR_{PS} - 1}{VSWR_{PS} + 1} = \frac{1.06 - 1}{1.06 + 1} = 0.029$$

$$\frac{u(M_{SP})}{M_{SP}} = \frac{\frac{|\Gamma_{SPL}| \cdot |\Gamma_{PAD}|}{\sqrt{2}}}{|1 - \Gamma_{SPL}\Gamma_{PAD}|^2} = 0.001 = 0.10\%$$

► $\frac{u(M_{SS-C})}{M_{SS-C}}$ – uncertainty due to mismatch between the calibrator source and the input of the N5532A during power meter calibration.

$$\Gamma_{CAL\ SOURCE} = \frac{VSWR_{CS} - 1}{VSWR_{CS} + 1} = \frac{1.05 - 1}{1.05 + 1} = 0.024$$

$$\Gamma_{SPLITTER} = \frac{VSWR_{SPL} - 1}{VSWR_{SPL} + 1} = \frac{1.10 - 1}{1.10 + 1} = 0.048$$

$$\frac{u(M_{SS-C})}{M_{SS-C}} = \frac{\frac{|\Gamma_{SPL}| \cdot |\Gamma_{PAD}|}{\sqrt{2}}}{|1 - \Gamma_{SPL}\Gamma_{PAD}|^2} = 0.0008 = 0.08\%$$

- $\frac{u(M_{SP=C})}{M_{SP-C}}$ – uncertainty due to mismatch between the power splitter and the 3 dB pad during power meter calibration.

$$\Gamma_{SPLITTER} = \frac{VSWR_{SPL} - 1}{VSWR_{SPL} + 1} = \frac{1.10 - 1}{1.10 + 1} = 0.048$$

$$\Gamma_{PAD} = \frac{VSWR_{PS} - 1}{VSWR_{PS} + 1} = \frac{1.06 - 1}{1.06 + 1} = 0.039$$

$$\frac{u(M_{SP-C})}{M_{SP-C}} = \frac{\frac{|\Gamma_{SPL}| \cdot |\Gamma_{PAD}|}{\sqrt{2}}}{|1 - \Gamma_{SPL}\Gamma_{PAD}|^2} = 0.001 = 0.10\%$$

- $\frac{u(M_{PS-C})}{M_{PS-C}}$ – uncertainty due to mismatch between the 3 dB pad and the power sensor during the power meter calibration.

$$\Gamma_{SENSOR} = \frac{VSWR_{SEN} - 1}{VSWR_{SEN} + 1} = \frac{1.05 - 1}{1.05 + 1} = 0.024$$

$$\Gamma_{PAD} = \frac{VSWR_{PS} - 1}{VSWR_{PS} + 1} = \frac{1.06 - 1}{1.06 + 1} = 0.029$$

$$\frac{u(M_{PS-C})}{M_{PS-C}} = \frac{\frac{|\Gamma_{SEN}| \cdot |\Gamma_{PAD}|}{\sqrt{2}}}{|1 - \Gamma_{SEN}\Gamma_{PAD}|^2} = 0.0005 = 0.05\%$$

- $\frac{u(P_{CAL})}{P_{CAL}}$ – uncertainty of the calibrator output power (nominally 1 mW)

  Specification: ±0.4 %

Assuming a Gaussian distribution for this parameter:

$$\frac{u(P_{CAL})}{P_{CAL}} = \sigma_{CAL} = \frac{0.4\%}{2} = 0.2\%$$

Combined uncertainty

The combined uncertainty for the reference power sensor reading is:

$$\frac{U^2\left(P_{gZ0}\right)}{P_{gZ0}^2} = \left[\begin{array}{l}(0.07)^2 + (0.07)^2 + (0.0058)^2 + (0.025)^2 + (0.0025)^2 \\ + (-0.023)^2 + (0.25)^2 + (0.75)^2 + (1.5)^2 + (0.75)^2 + (0.0)^2 \\ + (0.69) + (0.10)^2 + (0.05)^2 + (0.08)^2 + (0.10) + (0.05)^2 + (0.2)^2\end{array}\right] = 4.0\ \%^2$$

$$\frac{U_C\left(P_{gZ0}\right)}{P_{gZ0}} = 2.0\%$$

Converting to dB:

$$\frac{U_C\left(P_{gZ0}\right)}{P_{gZ0}} = 10 \cdot \log\left(1 + \frac{2.0}{100}\right) = 0.086\ \text{dB}$$

Expanded uncertainty

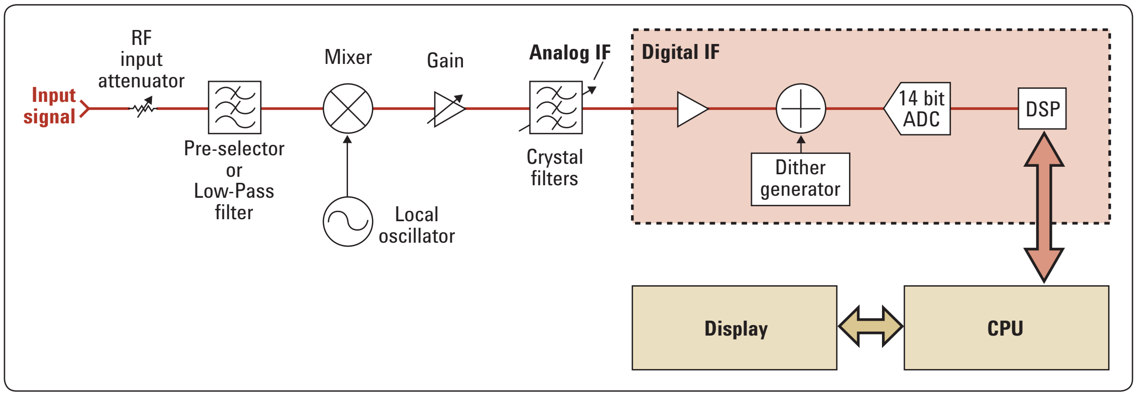
Assuming a Gaussian distribution of combined uncertainty values and assuming a 99% confidence interval, the coverage factor is 2.57.

$$\frac{U_{EXP}\left(P_{gZ0}\right)}{P_{gZ0}} = (2.57) \cdot (0.086\ \text{dB}) = 0.22\ \text{dB}$$

APPENDIX B:
Derivation of Tuned-RF-Level (TRFL) Amplitude Uncertainty

Measurements made in TRFL mode utilize the PSA spectrum analyzer

Measurements made in TRFL mode utilize the PSA spectrum analyzer.



The critical blocks that determine the relative-amplitude (e.g., TRFL) accuracy of the PSA are the Analog IF and the Digital IF. The most important performance characteristic of the PSA for amplitude measurement is "detector linearity". The detector in the PSA is a 14-bit, analog-to-digital converter (ADC) coupled to a digital signal processing (DSP) chip. This detection method is far more accurate than the traditional method employing analog logarithmic amplifiers and analog detectors.

Detector linearity applies throughout the entire TRFL amplitude measurement range. When the signal level drops below about –70 dBm, residual noise and range changing in the PSA become the dominant contributors to the uncertainty of the TRFL measurement.

Absolute power measurement by the N5531S below the range covered by the power meter (below –10 dBm) requires the TRFL measurement mode utilizing the PSA. The PSA's absolute power reference is established by the power meter. Consequently, in TRFL mode, there are two contributors to the value of the absolute RF power accuracy specification:

- The absolute reference measurement made by the power sensor/power meter combination;
- The relative TRFL measurement made by the PSA.

The combined uncertainty of these two separate contributing measurements becomes the specified performance limit (i.e., the specification) for absolute TRFL power accuracy.

$$u^2\left(P_{ABS-TRFL}\right) = u^2\left(P_{REF}\right) + u^2\left(P_{TRFL}\right)$$

When operating as a measuring receiver in the TRFL mode, the PSA amplitude accuracy (uncertainty) is dominated either by the PSA linearity or by the residual noise level, as shown in **Figure B-1**. These two regions are separated by the line labeled "Residual Noise Threshold", which marks the power level below which the signal-to-noise ratio (SNR) becomes the dominant contributor to the accuracy of the TRFL measurement. In the Measuring Receiver Personality chapter of the PSA Specifications Guide (hereinafter called the "N5531S data sheet"):

Residual noise threshold = minimum power + 30 dB

"Minimum Power" is specified in the N5531S data sheet for various frequency bands, resolution-bandwidth settings and pre-amp settings.

Throughout the power ranges shown in Figure B-1, the power indicated by the PSA, P_{TRFL}, is subject to the following sources of measurement uncertainty:

- $u_{LIN}(P_{TRFL})$ Uncertainty due to the detector linearity
- $u_{SNR}(P_{TRFL})$ Uncertainty due to the residual noise
- $u_{RANGE}(P_{TRFL})$ Uncertainty of the range-to-range cal factor

Measured uncertainty vs. input power relationship

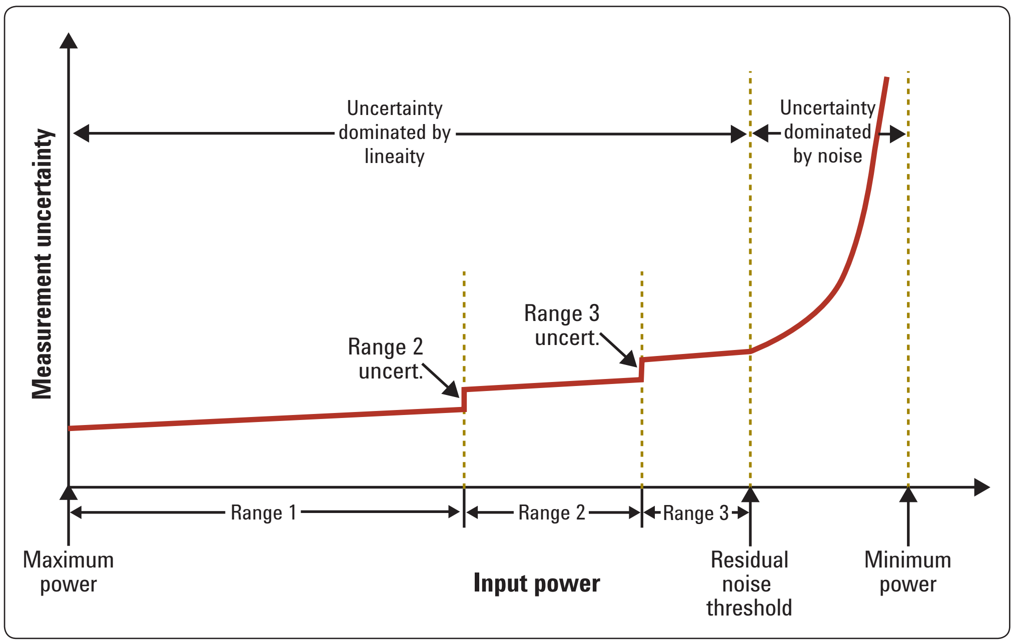


Figure B-1: Tuned RF level measurement uncertainty

Detector linearity

For the PSA, there are no factory or field adjustments for detector linearity. Its value depends directly on the accuracy of the analog-to-digital converter (ADC) in the PSA.

The uncertainty due to the detector "nonlinearity" arises from five distinct mechanisms in the PSA:

a. Single-tone compression in the conversion chain

The nonlinear behavior of the front-end of the spectrum analyzer can be characterized as compression. Compression begins when the plot of output level vs. input level begins to deviate from a linear relationship. As the input power level increases, the output level begins to flatten out and it takes larger and larger increases in input level to produce steady increments in output level. Compression in the PSA is caused primarily by the conversion chain (mixers and amplifiers). If the power level at the first mixer is limited to ≤ –28 dBm, the uncertainty caused by compression has been determined to be less than 0.002 dB.

$$u_{COMPR} = 0.002 \text{ dB}$$

The N5531S measurement algorithm adjusts the amount of PSA input attenuation to maintain input mixer level at or below –28 dBm.

b. Crystal filter hysteresis

The PSA utilizes a single-pole crystal filter in the final analog IF stage. Crystal filters suffer from a "hysteresis effect", wherein their loss depends on their signal-level exposure history. This problem is well known and designers take precautions to keep the drive level to the crystal filter low enough to minimize this effect. This hysteresis effect decreases as the bandwidth of the crystal filter is increased. In the worst case, a crystal filter can have hysteresis as large as 0.04 dB. This level is assured by factory and field-calibration testing of the scale fidelity.

The effect of hysteresis can be modeled as "ΔESR with drive", the change in the crystal filter's equivalent series resistance with drive level. When the bandwidth of the analyzer is increased from the narrowest settings, the effect of the ΔESR decreases proportionately. The linearity of the PSA is tested with the narrowest crystal-filter bandwidths.

In the PSA, the crystal-filter bandwidth is a function of resolution-bandwidth setting when in swept mode, or of the FFT frequency span in FFT mode. The crystal-filter bandwidth is never less than 5 kHz. All PSAs are tested in production to ensure that the nonlinearities in the minimum bandwidth case never exceed 0.04 dB. For bandwidths of 50 kHz or greater, the hysteresis effect will, therefore, never exceed 0.004 dB. Crystal bandwidths are set to 2.5 times the RBW for swept operation or 1.25 times the FFT-width for FFT operation. The N5531S system uses zero-span mode for TRFL measurements. The hardware in the analyzer is custom-controlled to apply the 50 kHz crystal filter to the measurement. The resulting uncertainty due to that filter is 0.004 dB.

$$u_{HYST} = 0.004 \text{ dB}$$

c. Processing resolution

Trace processing in the PSA generates an error due to quantization that can be as large as 0.001 dB.

$$u_{TRACE} = 0.001 \text{ dB}$$

d. ADC-range gain alignment

The ADC auto-ranges its gain from nominally zero to as much as +18 dB, based on the input signal level. This ranging is subject to alignment errors that have never been seen to exceed 0.01 dB.

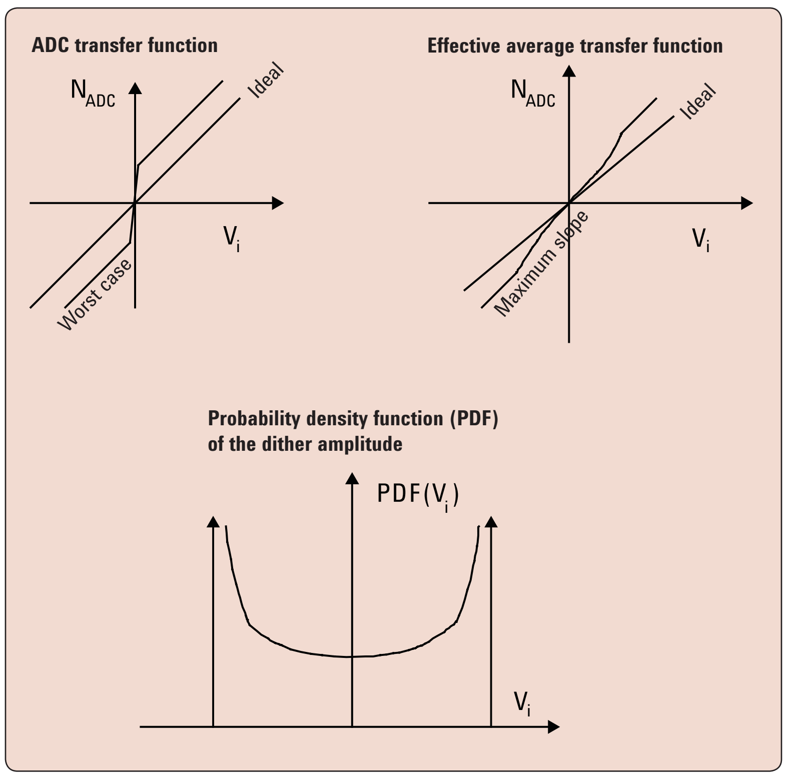
However, if the power level at the first mixer of the PSA is ≤ –28 dBm, then the ADC-range gain is always set to maximum. Consequently, for an N5531S TRFL measurement, there will be no error due to switching the gain in front of the ADC.

$$u_{ADC\,GAIN} = 0.000 \text{ dB}$$

e. ADC linearity

The ADC is guaranteed by its manufacturer to have an integral linearity error of less than one part in 214. The PSA employs a "dithering signal" technique to enhance the ADC linearity. The dither signal is a sine wave with pseudo-random frequency modulation at 20% of full-scale level. The effect of this dither signal can be modeled as a transfer function that is the convolution of the ADC transfer function with the probability density function of the dither signal. In the worst case, the ADC transfer function has a maximum linearity error of one polarity for a positive signal just above zero and an opposite linearity error for a signal just below zero. Even with this nonlinearity, the slope variation of the effective transfer function does not exceed **0.002 dB**.
See figures below.

$$u_{ADC\,LIN} = 0.002 \text{ dB}$$



f. Combined uncertainty for detector linearity

As a conservative estimate of the total uncertainty due to the detector linearity, the worst case linear sum of the contributors listed above is:

$$u_{LIN}\left(P_{TRFL}\right) = u_{COMPR} + u_{HYST} + u_{TRACE} + u_{ADC\ GAIN} + u_{ADC\ LIN}$$

$$u_{LIN}\left(P_{TRFL}\right) = 0.002\text{ dB} + 0.004\text{ dB} + 0.001\text{ dB} + 0.000\text{ dB} + 0.002\text{ dB} = 0.009\text{ dB}$$

When the uncertainty is computed, as shown, instead of demonstrated, there is always the risk that additional error contributors exist but have not been discovered or included in the sum. To minimize that risk, many analyzers were tested against calibrated reference attenuators. These reference attenuators had errors of their own that increased with the amount of attenuation. This verification technique still allowed risk that unknown errors might exist in the PSA that are not included in this analysis. To cover that risk, the PSA is specified to have another component of error. That component is ±0.003 dB/10 dB. To match customer needs to replace the HP 8902, and to be extra conservative in specifications, the specification is widened to:

$$u_{LIN}(P_{TRFL}) = 0.015\text{ dB} \pm 0.005\text{ dB/10 dB}$$

Residual noise

The indicated TRFL power is due to the sum of the signal power at the frequency of interest and the noise power within the selected resolution bandwidth. For high values of signal-to-noise ratio (≥ 40 dB), the noise has a negligible effect on the displayed amplitude of the signal. When the signal-to-noise ratio (SNR) drops below about 20 dB, the noise becomes a significant component of the indicated value of the signal.

At low SNR, the instantaneous, indicated value of a steady CW signal is continually changing due to the additive/subtractive effect of the noise. The PSA actually indicates the combined power of the signal plus noise: P_{S+N}. In order to make a precise estimate of just the signal power, P_S, we need to eliminate the mean error due to noise and minimize the fluctuation due to the noise.

The PSA acts to minimize the fluctuation due to the residual noise by *averaging* the combined $P_S + P_N$ and to eliminate the mean error due to the noise by performing a subtraction:

$$P_S = P_{S+N} - P_N \,[\mathrm{dB}]$$

where: P_{S+N} is the power in the signal plus noise

P_N is the power in the residual noise alone

After the mean error is subtracted, the fluctuations remain. The mean error subtraction process has errors due to imperfect estimation of the noise, but these errors are much smaller than the errors due to noise fluctuations even with large amounts of averaging. Therefore, the rest of this section will deal with the noise fluctuations only.

The analyzer is set up in varying ways according to the signal-to-noise ratio. Our interest here is in the most difficult, low signal-to-noise ratio cases. In these cases, the measurement is made by averaging a zero span sweep with a sweep time computed as 10/RBW, and with the maximum number of such traces averaged automatically selected by the software (to keep the measurement time from growing without bound), which is 900.

The signal-to-noise ratio can be used to compute the standard deviation of the averaged traces. First, we must compute the standard deviation of each individual trace. To compute that, let us start with the standard deviation of the instantaneous measurement of a CW signal with noise.

The noise added to a CW signal is random. Being random, its phase relative to the CW component is random. We can describe this noise in more than one way. One way would be in terms of its magnitude and phase distribution. Another way would be in terms of its real and imaginary components. The most useful way is a rotated version of its real and imaginary components: we can decompose it into two random components—one component is in-phase with the CW signal, and one is in quadrature. Each of these components has the same average power. Therefore, each one has 3 dB less power than the total noise. When the signal-to-noise ratio is large, the quadrature component adds negligibly to the variation in measured level. So the relevant noise is the in-phase component with half the total noise power.

The standard deviation of the envelope voltage of the in-phase component is equal to the voltage computed from its total power. With this information, the standard deviation of the instantaneous envelope voltage can be computed.

Let's start with a noise voltage sinusoid, of level v, that is expressed in decibels relative to the signal power, that may either add to or subtract from the carrier power. The error associated with v is:

$$Error\_dB = 20 \cdot \log_{10}(1 \pm 10^{v/20})$$

For small values of v, the error is linearly related to v. We can do a Taylor-series expansion on the error equation:

$$Error\_dB \approx \Delta v \cdot \frac{d}{dv}(20 \cdot \log_{10}(1 + 10^{v/20}))$$

This works out to:

$$Error\,dB \approx \Delta v \cdot 20 \cdot \log_{10}(e)$$

The latter term, 20log(e), is 8.69 dB, also known as 1 neper.

Substituting in the voltage error (from the signal-to-noise ratio) for v, we can compute the standard deviation of our measurement:

$$\sigma_{CW} = (8.69\text{ dB}) \cdot 10^{-\frac{SNRatio + .301}{20}}$$

This expression is actually slightly conservative compared to a full statistical treatment using Rician distributions, so it is useful for our purposes.

This standard deviation is reduced by filtering. The filtering comes from averaging the result for a duration of 10/RBW. How much filtering does that give us?

The distribution of the noise of the detected signal can be described as having a noise bandwidth of half of the predetected signal. The noise bandwidth of the predetected signal is 1.056 times the RBW for our very close to Gaussian RBW filters. The noise bandwidth of the averaging process is $1/(2 \cdot t_{INT})$, where t_{INT} is the integration time.

The standard deviation of the filtered envelope will be reduced by the square root of the ratio of the noise bandwidth of the signal to the noise bandwidth of the averaging process. Therefore:

$$\sigma_{TRACE} = \frac{8.69\text{ dB}}{\sqrt{t_{INT} \cdot NBW_{RBW}}} \cdot 10^{-\frac{SNRatio + 3.01}{20}}$$

Given that t_{INT} = 10/RBW, and that NBW_{RBW} = 1.056 · RBW, we get:

$$\sigma_{TRACE} = 2.67\text{ dB} \cdot 10^{-\frac{SNRatio + 3.01}{20}}$$

When we average this 900 times, the standard deviation improves by the square root of 900, which is a factor of 30, giving:

$$\sigma_{AVG} = 0.0891\text{ dB} \cdot 10^{-\frac{SNRatio + 3.01}{20}}$$

We can express the total error in this format instead:

$$Error = 0.0012 \cdot (SNRatio - 30)^2$$

SNRatio – 30 dB is the same as input power minus residual noise threshold power. For SNRatio values in the 0 to 30 dB range, the 3σ result is either well under 0.01 dB, or the Error expression is conservative relative to three standard deviations.

Here is a table that demonstrates the numerical relationship:

Signal-to-noise ratio	3 x standard deviation	Error expression
30 dB	0.006 dB	0 dB
25	0.011	0.030
20	0.019	0.120
15	0.034	0.270
10	0.060	0.480
5	0.106	0.750
0	0.189	1.080

The Error expression is used as a component of both the absolute and relative TRFL accuracy expressions shown in Table 5 and Table 6.

Range-to-range cal factor

TRFL measurement with the PSA spans three amplitude ranges. This section will discuss the errors in each of those three ranges. Here is how those ranges are arranged:

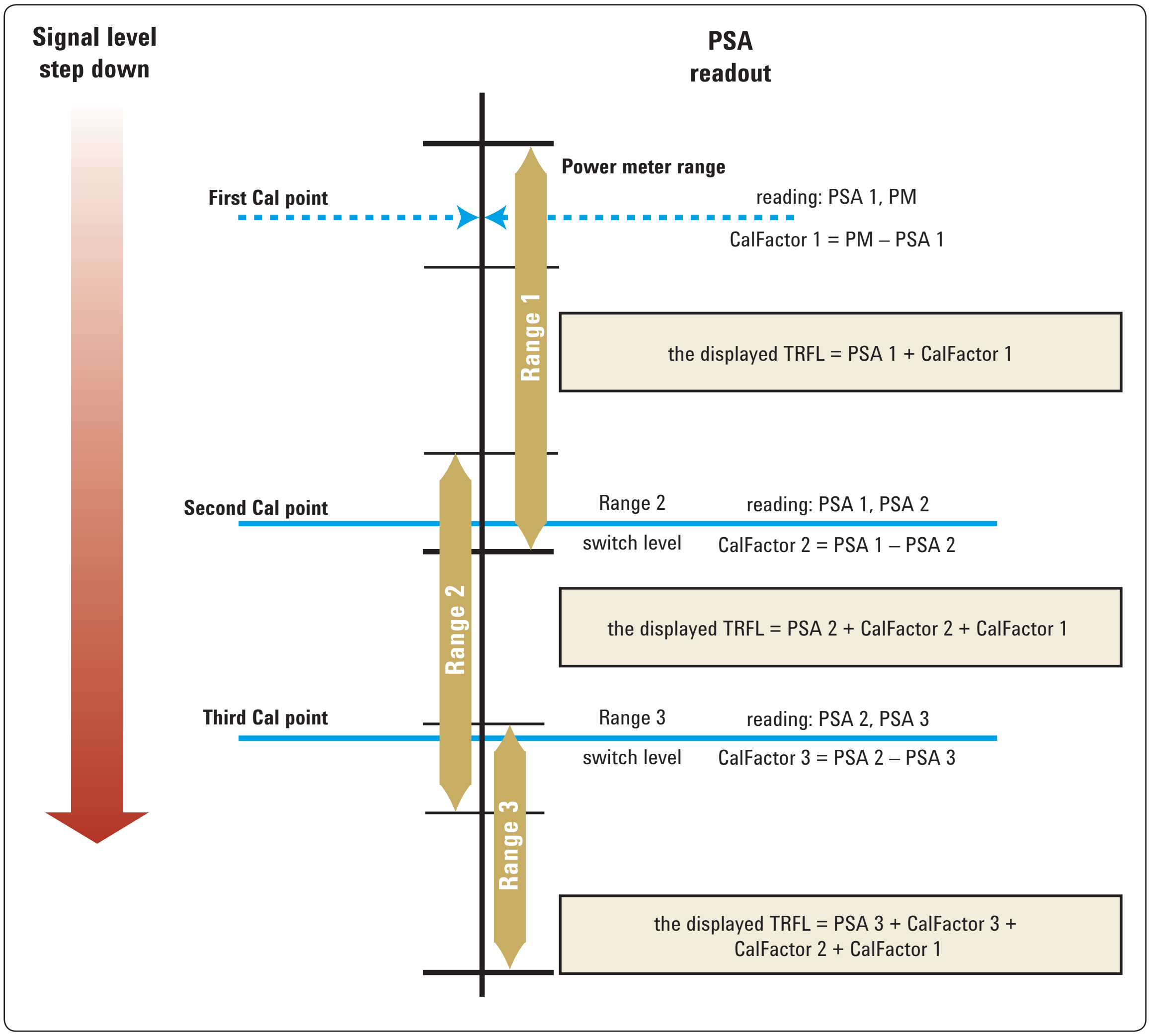


Figure B-2: Tuned RF level measurement ranges

To avoid error due to signal compression, the power level at the first mixer of the PSA is maintained at ≤ –28 dBm by manipulating the input attenuator of the PSA and internal preamplifier settings.

Example: If the signal power level at the input of the N5532A sensor module is 0 dBm, then the nominal power level at the PSA input will be –7 dBm (due to the power splitter loss). The internal attenuator of the PSA must be set to 21 dB or more to ensure that the power level never exceeds –28 dBm at the mixer. By maintaining this attenuator setting, the PSA mixer always identifies a level that is 28 dB lower than the level at the sensor module input.

Range-to-range cal factor uncertainty

Cal factor 1

Cal factor 1 applies in Range 1, Range 2, and Range 3. It accounts for the difference in reading between the power meter and the PSA for a fixed power level measured in Range 1. Cal factor 1 is determined as follows:

1. The power level of the signal is measured in Range 1 by the N5532A sensor module and the P-Series power meter. The value is stored as P_{PM1} (in dBm);

2. The PSA is set to Range 1 (the internal attenuator is set to 30 dB) and the same power level is measured by the PSA, via the N5532A connection. The value is stored as P_{PSA1}.

3. $$CalFactor\,1 = P_{PM1} - P_{PSA1}\,[dB]$$

The uncertainty of Cal factor 1 is:

$$u^2(CF1) = C_1^2 \cdot u^2(P_{PM1}) + C_2^2 \cdot u^2(P_{PSA1})$$

where C_1 and C_2 are sensitivity coefficients given by:

$$C_1 = \frac{\partial(CF_1)}{\partial P_{PM1}} = 1 \quad \text{and} \quad C_1 = \frac{\partial(CF_1)}{\partial P_{PSA1}} = -1$$

Therefore: $u^2(CF1) = u^2(P_{PM1}) + u^2(P_{PSA1})$

The uncertainty value $u(P_{PM1})$ is derived in Appendix A and is 0.086 dB.

$$u(P_{PM1}) = 0.086\text{ dB}$$

The uncertainty value $u(P_{PSA1})$ is dependent only on the marker readout resolution. All other PSA uncertainties are eliminated by the comparison of the power meter reading to the PSA reading.

So, $u(P_{PSA1})$ can be expressed as:

$$u(P_{PSA1}) = u(MKR)$$

From the PSA data sheet: u(MKR) = 0.001 dB, with averaging ON.

Because this parameter is specified in dB, we will assume that it was determined from data taken in dB. The marker readout value is a quantized value that is assigned to the power level within a distinct vertical-scale "bucket." Any value that falls within the boundaries of this "bucket" will be assigned the same amplitude value. Conversely, the probability of any value within the bucket boundaries having caused the reading is the same. This situation is best described by a uniform probability distribution (i.e., every value in the "bucket" interval is equally likely and will be assigned the same value by the PSA).

For a uniform distribution, the standard deviation is

$$\sigma_{UNIFORM} = \frac{w}{\sqrt{12}}$$, where w is the width of the uniform distribution.

So, the standard uncertainty of a true value that falls within a particular amplitude bucket is equal to:

$$u(MKR) = \sigma_{MKR} = \frac{0.001\text{ dB}}{\sqrt{12}} = 0.0003\text{ dB}$$

The uncertainty of the Range 1 Cal factor is then:

$$u^2(CF1) = u^2(MKR) + u^2(P_{PM1})$$

$$U_C^2(CF1) = (0.0003\text{ dB})^2 + (0.086\text{ dB})^2$$

$$U_C(CF1) = 0.086\text{ dB}$$

Cal factor 2

Cal factor 2 applies in Range 2, and then contributes in Range 3 as well. It accounts for the change from Range 1 to Range 2. Cal factor 2 is determined as follows:

1. The power level of the signal is measured in Range 1 and the value is stored as P_{R1}(in dBm);
2. The PSA is switched to Range 2:
   a. The preamp (if present) remains OFF.
   b. The internal attenuator is changed from 30 dB to 10 dB.
3. The power level of the signal (unchanged) is measured again, now in Range 2, and is stored as P_{R2} (in dBm);

$$CalFactor2 = P_{R1} - P_{R2} \left[dB\right]$$

The uncertainty in Cal factor 2 is due entirely to the noisiness of the pair of measurements that compare Range 2 and Range 1.

The transition between ranges is constrained to occur at a signal-to-noise ratio of 22 dB or more, therefore, 22 dB is the worst case. The computation of uncertainty due to this noise will be brief because the computation process has been more thoroughly explained in the "Residual Noise" section of this appendix. With 22 dB signal-to-noise ratio, the in-phase noise is 25 dB below the CW signal, giving a standard deviation error of:

$$\sigma_{single} = 8.69\ \text{dB} \cdot 10^{-\frac{25}{20}}$$

This standard deviation is reduced by the square root of the number of averages (900) to give a standard deviation of 0.0158 dB. The signal is measured again in Range 2, but the signal-to-noise ratio is 100 times better, so when the standard deviation of that measurement is combined with the standard deviation of this measurement, the effect is negligible.

To compute the 95% confidence interval due to this noise, we multiply the standard deviation by 1.96 and get an interval of ±0.031 dB. Therefore, ±0.031 dB can be said to be the 95% confidence result for the worst case signal-to-noise ratio. The signal-to-noise ratio is likely to be any value between 22 and 32 dB. When we compute the probability of the noise causing an error outside the range of ±0.031 dB, we find the result is within this error 99.3% of the time. Unlike other specifications for this product which are worst-case specifications, this one is 99.3% confidence specification. This seems suitable because this specification is never used by itself, only in combination with other terms which are themselves pretty large, so that statistical combining leaves very little risk of failing to meet specifications.

Cal factor 3

Cal factor 3 applies in Range 3 only and accounts for the change from Range 2 to Range 3. Cal factor 3 is determined as follows:

1. The power level of the signal is measured in Range 2 and the value is stored as P_{R2} (in dBm);
2. The PSA is switched to Range 3:
3. The preamp (if present) is turned ON;
4. The internal attenuator is changed from 10 dB to 4 dB.
5. The power level of the signal (unchanged) is measured again, now in Range 3, and is stored as P_{R3} (in dBm);
6. $CalFactor\,3 = P_{R2} - P_{R3}\,[dB]$

The derivation of Cal factor 3 is the same as that of Cal factor 2. The only difference is that the noisiness of the measurement in Range 3 is nominally 40 times lower than the noisiness of Range 2, compared with a factor of 100 in the former derivation. Both these noise ratios give negligible contribution to the uncertainty.

A measurement made in Range 3 will involve CF1, CF2 and CF3. The uncertainty of these combined cal factors will be:

$$u_{RANGE}(P_{TRFL}) = \sqrt{u^2(CF1) + u^2(CF2) + u^2(CF3)}$$

Measurements made in Range 2 or Range 1 will have smaller values of overall range-to-range uncertainty.

Combined uncertainty for P_{TRFL} measurement

Since the effects of linearity, range offset, and signal-to-noise ratio are independent of one another, the combined uncertainty can be expressed as:

$$u_C^2(P_{TRFL}) = u_{LIN}^2(P_{TRFL}) + u_{SNR}^2(P_{TRFL}) + u_{RANGE}^2(P_{TRFL})$$

From this equation, it is clear that the overall uncertainty of an indicated TRFL measurement made by the N5531S is dominated by the PSA linearity and range offsets until the signal level nears the level of the noise.

Appendix C:
Two-Resistor Versus Three-Resistor Power Splitter Choice

The N5532A sensor module figure is repeated here:

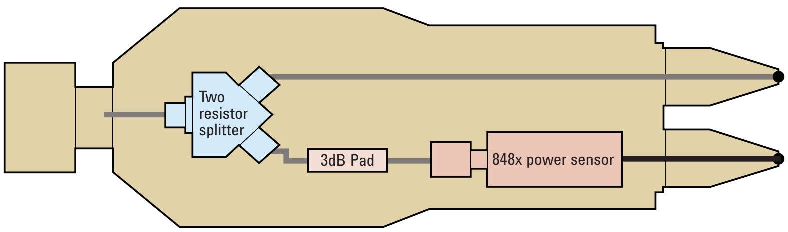


Figure 2: N5532A sensor module

Inside the N5532A, a two-resistor power splitter supplies equal signals to the spectrum analyzer output port and to the 3 dB pad/power sensor combination. A two-resistor power splitter is constructed with 50 Ω resistors from the input port to each output port. The nominal power loss through the splitter to each output port is 7 dB (6 dB, ideally).

Why not use a three-resistor power divider instead? A three-resistor power splitter is constructed with 16.7 Ω resistors from each of the three input ports to a central node. At first glance, this seems preferable for this application. If an ideal three-resistor power divider is terminated at each of its ports by the characteristic impedance of the transmission line (50 ohms), then each port of the power divider presents a 50 Ω input impedance. This insures that optimizing the match presented by the spectrum analyzer will also optimize the match seen by the signal generator and power meter. The same concept applies to all three ports, however, power reflected from the spectrum analyzer input would be attenuated by approximately 6 dB before appearing at the input to the 3 dB pad. This undesired reflection from the spectrum analyzer would change the total power seen by the sensor, causing an error in the power indicated by the external power meter.

The effect of reflections with a two-resistor power splitter is qualitatively similar, though the reflection attenuation is about 14 dB instead of about 6 dB. But let's look at it another way.

Our goals in designing the sensor module are these: We want the optimum match at the input port. We want to optimize the accuracy of both the spectrum analyzer and the power sensor.

Both the two- and three-resistor power splitters show ideally 50 Ω input impedance when the spectrum analyzer and the power sensor are both 50 Ω input impedance. The two-resistor splitter has the advantage, with its larger resistor values, the input port match depends less on the matches at the splitter outputs. Thus, the "optimum match" goal is best served by the two-resistor choice.

Regarding the "optimize the accuracy" goal is where the two-resistor choice is most valuable. Both of the measuring devices are optimized and calibrated for accuracy when driven from a 50 Ω impedance. Regardless of the impedances of the other instruments in the circuit, both are configured to be reporting one-half of the voltage (the situation is easier to understand from a voltage perspective than from a power perspective) at the node between the two 50 Ω resistors. This situation is equivalent to the situation under which they are calibrated. Thus, they are both, simultaneously, optimally accurate as configured with a two-resistor splitter. This is why two-resistor splitters are known to be the most accurate devices for leveling and ratio measurements. Another discussion of this topic is available in the Agilent Application Note *Differences in Application Between Power Dividers and Power Splitters*, literature number 5989-6699EN.

It is true that the instantaneous impedance seen by the two measuring devices is 83.3 Ω (VSWR of 1.67:1) in the two-resistor case, while being 50 Ω in the three-resistor case. But it is not the instantaneous impedance that determines the accuracy, so the two-resistor splitter is the better choice.

Related Literature

- "*Fundamentals of RF and Microwave Power Measurements (Part 2)":*
  Power Sensors and Instrumentation, AN 1449-2,
  literature number 5988-9214EN.

- "*ISO Guide to Uncertainty in Measurement",*
  in the *Fundamentals of RF and Microwave Power Measurements*, Application Note 1449-3,
  literature number 5988-9215EN.

- *N1911A/N1912A Service Guide*,
  Agilent document number: N1912-90015.

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