



NIST Technical Note
NIST TN 2391

**Noise Based Power Calibration
Techniques for Mid-band Radio Frequency
Sensors**

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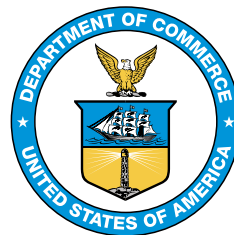
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National Advanced Spectrum and Communications Test Network (NASCTN)

Mission Statement: *The National Advanced Spectrum and Communications Test Network (NASCTN) provides, through its members, a network for robust test processes and validated measurement data necessary to develop, evaluate and deploy spectrum sharing technologies that can improve access to the spectrum by both federal agencies and non-federal spectrum users.*

NASCTN originated in 2015 as a multi-agency chartered partnership and as of 2022, is a congressionally mandated function of the National Institute of Standards and Technology (NIST). NASCTN is coordinated with the National Telecommunications and Information Administration (NTIA) and other Federal agencies, as appropriate. NIST hosts the NASCTN capability at the Department of Commerce Boulder Laboratories in Boulder, Colorado.

The NASCTN program office facilitates and coordinates the use of intellectual capacity, modeling and simulation, and laboratory and test facilities to meet national spectrum interest and challenges.

NASCTN test projects follow a transparent framework including public dissemination of test plans, adjudicated comments to the test plans, formal reports containing test methods, data, and software to the maximum extent permissible.

The NASCTN provides an impartial, scientifically rigorous forum for addressing spectrum-sharing challenges in an effort to accelerate the deployment of wireless technologies among commercial and federal users and to measure the impacts of spectrum dependent systems deployments. NASCTN provides robust test processes, validated measurements data, and statistical analysis necessary to develop, evaluate, and deploy spectrum sharing technologies that can improve access to the spectrum by both federal agencies and non-federal spectrum users.



NASCTN CBRS SEA

This work is a result of the [NASCTN](#) Citizens Broadband Radio Service ([CBRS](#)) Sharing Ecosystem Assessment ([SEA](#)) program. The [NASCTN CBRS SEA](#) program was requested and sponsored by Defense Information Systems Agency ([DISA](#)) to measure emissions in the [CBRS](#) band. These emissions would be analyzed to ascertain the effectiveness of the sharing ecosystem between [CBRS](#) systems as managed by Spectrum Access Systems ([SASs](#)), and Department of War ([DoW](#)) systems as monitored by Environmental Sensing Capabilities ([ESCs](#)). The collected data will be used by [DoW](#) to provide insight into the sharing ecosystem's effectiveness, and track changes in the spectrum environment over time. This document explains the calibration process of sensors that recorded that data. Test members are drawn from government, academia, and/or industry and are the foundation to [NASCTN's](#) coordinated, national testing and evaluation capability to enable the improvement and expeditious addressing of national spectrum test interest and challenges.

- Provide expertise and capabilities to cooperatively enable [NASCTN](#) studies;
- Work as a partnership to address the interests and equities of all spectrum stakeholders in a fair, equitable, and non-preferential manner; and
- Through sharing of technical resources, with consideration for cost, provide for, liaison, or coordinate access to testing and evaluation capabilities.

Test Members to the [CBRS SEA](#) Program are (in alphabetical order):

- Department of Defense (DoW), Chief Information Officer (CIO)
- MITRE Corporation
- National Aeronautics and Space Administration (NASA), Langley Research Center
- National Institute of Standards and Technology (NIST), Communications Technology Laboratory
- National Telecommunications and Information Administration (NTIA), Institute for Telecommunication Sciences (ITS)

Other collaborative partners to [CBRS SEA](#) are:

- Hampton University, Norfolk VA.
- Naval Surface Warfare Center, Dahlgren Division
- University of Colorado, Boulder CO.
- United States Coast Guard, C5ISR Program, Rescue 21
- United States Marine Corps, IC4
- United States Navy, Naval Base Pt Loma
- Virginia Department of Transportation

Abstract

This technical report details approaches to the amplitude calibration and noise characterization of frequency-dependent power sensors. The frequency range of sensing addressed is between 1 and 6 GHz and calibrations are performed via the Y-factor technique. The Y-factor technique relies on noise with a known power spectral density to extract the sensors' effective gain and noise figure. This technique, including its limitations, is reviewed. The construction and characterization of a programmable noise source or calibration probe is also documented. In addition, methods to transfer the known power of a commercial noise diode to the calibration probe are discussed. Measurements of commercial noise diodes and the calibration probe in by the NIST radiometer noise service are detailed. Finally, results from the [NASCTN CBRS SEA](#) project are presented. This program involved an extended large-scale sensor deployment monitoring the electromagnetic environment in the 3530 - 3710 MHz range. These deployed sensors were calibrated hourly over multiple years. The calibration was traceable to NIST through a radiometer measurement of the installed noise diode. The uncertainty of the gain and noise figure for multi-sensor field calibrations is developed under multiple assumptions.

Keywords

Advanced Communications;CBRS;CBSD;DPA;ESC;GAA;NASCTN;PALS;RADAR;SAS;Sharing Ecosystem Assessment

Acronyms

ADC	analog-to-digital converter	ND	noise diode
CBRS	Citizens Broadband Radio Service	NEMA	National Electrical Manufacturer's Association
COTS	commercial off-the-shelf	NIST	National Institute of Standards and Technology
CONUS	contiguous United States	NTIA	National Telecommunications and Information Administration
dB	decibel	P1dB	1 dB gain compression point
DISA	Defense Information Systems Agency	PAPR	Peak-to-Average Power Ratio
DoW	Department of War	PMPR	Peak-to-Median Power Ratio
DPA	Dynamic Protection Area	AMPR	Average-to-Median Power Ratio
DSP	digital signal processing	RF	radio frequency
ENR	excess noise ratio	RMS	root mean square
ESC	Environmental Sensing Capability	SAS	Spectrum Access System
IQ	in-phase and quadrature	SDR	software defined radio
ISM	industrial, scientific and medical	SEA	Sharing Ecosystem Assessment
ITS	Institute for Telecommunication Sciences	SigMF	signal metadata format
LTE	long-term evolution	SPU	signal processing unit
MITRE	The MITRE Corporation	TEC	thermo electric heater/cooler
NASCTN	National Advanced Spectrum and Communications Test Network	VBA	verification bench testing and acceptance
		VST	vector signal transceiver

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Xifeng Lu	NIST	Radiometer Measurements
Aric W. Sanders	NIST	Analysis of Calibration Data and Primary Writer
Todd Schumann	NTIA ITS	Calibration Analysis, Sensor Setup

Executive Summary

This technical note serves as a collection of practical techniques and results for noise based power calibrations. In particular, this note focuses on sensors that are designed to detect power in the 1-6 GHz frequency range. It simultaneously acts as documentation of methodology and as a specific supplement to a larger body of work, driven by the National Advanced Spectrum and Communications Test Network (NASCTN) Citizens Broadband Radio Service (CBRS) Sharing Ecosystem Assessment (SEA) project. The purpose of the NASCTN CBRS SEA project is to provide data-driven insight into the CBRS sharing ecosystem's effectiveness between commercial and Department of War (DoW) incumbent systems, and to track changes in the spectrum environment over time. CBRS is the "first of a kind" nationwide shared spectrum ecosystem in the 3550-3700 MHz band. The implementation of this project included a sensor network that required calibration.

1. **Introduction:** A description of the Y-factor method, sensor definition and the NASCTN CBRS SEA program.
2. **General Discussion:** An extended discussion of the statistics and common issues with the Y-factor calibration.
3. **Calibration Probe:** An example of the design and characterization of an amplified noise source for calibrations.
4. **Transfer System:** The setup and use of a system to transfer calibrations to other systems.
5. **Noise Source Radiometer Characterization** The use of the National Institute of Standards and Technology (NIST) radiometer measurement to characterize noise diodes and calibration probes.
6. **SEA Sensor Results** The results of laboratory and field calibrations of sensors used in the NASCTN CBRS SEA program.
7. **Conclusion:** A summary of the findings from previous chapters including an estimate of calibration times and uncertainties.

This calibration report is not designed to be a comprehensive document about the NASCTN CBRS SEA project. Detailed information about the project's goals, sensor construction, data manual, and ecosystem analysis can be found at <https://www.nist.gov/programs-projects/cbrs-sharing-ecosystem-assessment>. Access instructions for data, documentation and code produced by this project can be found at <https://pages.nist.gov/SEA-DATA>.

This calibration report is targeted at commercial, military, and civilian stakeholders who would benefit from calibration of sensors in midband frequencies. To access the data and detailed analyses produced by this program, email nasctn@nist.gov.

1. Introduction

The understanding of the electromagnetic environment surrounding us is an endeavor as natural as using our eyes to appreciate a sunrise. However, understanding the electromagnetic environment that we are not naturally sensitive to requires some sort of enhancement to our senses. Therefore, it is important to understand how to link that enhancement to physical units, allowing us to share our observations with others. Here we discuss how to transform measurements made by a radio-frequency sensitive transducer to physical power units. There are many ways to accomplish this feat, however in this technical note we limit the discussion to a noise-based procedure. This procedure relies on the basic physical principal that at thermal equilibrium, energy distributes itself equally among the avenues available to it. This is known as the equipartition theorem, and it implies that all things at a finite temperature will have fluctuations in all degrees of freedom. When the system is electrical in nature, these fluctuations manifest as small random changes in voltage. These random changes are used to link measured voltages from sensors to equivalent temperatures. This technical note is dedicated to the specific use of the results from this to determine the value of power for a mid-band radio-frequency sensor. For the purpose of this technical note, a mid-band radio-frequency sensor is a device that is sensitive to electromagnetic radiation with frequencies between 1 GHz to 6 GHz. This note will pay special attention to the frequency range around 3550 - 3700 MHz known as the [CBRS](#) band. Including a detailed analysis of several different configurations of sensor deployed across the continental United States. Using a noise diode to characterize a sensor or measurement system is a well understood process and a plethora of technical notes, documents, and tutorials exist describing both techniques and uncertainties[1–6]. Although the techniques in this note build upon well established procedures, several key aspects required enhanced documentation:

1. The construction and use of a programmable amplified noise source.
2. The use and application of a transfer system to characterize unknown noise sources.
3. The use of a primary noise source and radiometer to characterize a large number of commercially available noise sources and realizations of amplified programmable noise sources.
4. A longitudinal study of field integrated noise sources in multiple sensors, installed in three distinct geographic situations.

1.1. Basics of The Y-Factor Technique

Since the calibrations in this technical note are variations on the well known Y-factor technique, it is instructive to reiterate the general workflow. The workflow can be summarized as:

1. Connect a noise diode with a known excess noise ratio ([ENR](#)) to the port of interest

2. Measure the noise power with the noise diode off (unplugged or de-energized)
3. Measure the noise power with the noise diode on (energized)
4. Calculate the ratio of powers or Y-factor $y = \frac{P_{on}}{P_{off}}$
5. Calculate the noise factor using $NF = \frac{ENR}{(y-1)}$
6. Calculate the effective gain using $G_{eff} = \frac{P_{on}}{k_B T B (ENR + NF)}$
7. Convert to logarithmic units decibel (dB) and report the results

This experimental procedure is based on the premise that at a physical temperature a random voltage in a sensor will have a total power of

$$P = G k_B T B \quad (1)$$

where k_b is Boltzman's constant $1.380649 \times 10^{-23} \frac{J}{K}$, B is the effective bandwidth of the observation in Hz, G is the effective gain of the detector and T is the temperature in Kelvin. This relationship is an approximation of the one-dimensional Planck law for thermal radiation [7] in the classical or long-wavelength limit. In this limit, the characteristic energy of single photons, or hf is small compared to the thermally available energy or $k_b T$. Specifically, the fluctuation of charge carriers in resistive devices, known as Johnson-Nyquist noise, states that for a resistor of resistance R we expect [8].

$$V_{rms} = \sqrt{4 k_B T R B} \quad (2)$$

If we devise a circuit with maximum energy transfer (matched resistance), we find that the total noise power available is the power flowing to a resistor from a source with open-circuited voltage given by equation 2, and internal resistance R [9].

$$P = \frac{1}{2} \frac{V_{rms}^2}{(2R)} = k_B T B \quad (3)$$

If an ideal sensor that has a power gain (G_p) is used to amplify the signal across the matched resistor we now have

$$P = G_p k_B T B \quad (4)$$

Of course, an ideal sensor is an approximation of a real sensor. This approximation is further improved by introducing an excess noise term from the sensor. This excess noise is equivalent to a temperature (T_{sensor}) so the amplified power from an external, matched-load after amplification is:

$$P = G_p k_B B (T_{sensor} + T_{thermal}) \quad (5)$$

T_{sensor} is commonly identified as the noise temperature of the sensor. Another way to frame this, is by quantifying the degradation of signal-to-noise ratio caused by the sensor

$F = \frac{S_{input}/N_{input}}{S_{output}/N_{output}}$, where F is known as the noise factor (or noise figure in linear units), and its value in **dB** is known as the noise figure. It can be shown that if the input noise is due to a $290K$ source ($T_{thermal} = 290K$) that the noise factor is

$$F = \frac{G_p k_B B T_{sensor} + G_p k_B B T_{thermal}}{G_p k_B B T_{thermal}} \quad (6)$$

and the noise figure (NF) is:

$$NF = 10 \log_{10}(F) \quad (7)$$

for more complete derivations see [2, 10]. To characterize the noise temperature, noise factor or noise figure of a sensor, the temperature of the matched resistor network can be changed from T_{cold} to T_{hot} and then the results compared. This is the essential premise of the Y-Factor method, and the hot and cold loads can be realized in a number of ways. For example, in the **NIST** radiometer, it is a precisely matched impedance held at a constant temperature (see section 5). In a system calibrated by a noise diode, T_{hot} is the energized state of the noise diode and T_{cold} is the off state, assumed to be at ambient temperature ($\approx 290K$). For the noise diode, the hot state is not at the actual temperature, but is electronic noise with the same properties as thermal noise with a very high equivalent temperature (normally $> 8,000K$). Specifically, shot noise from a reverse-biased Zener diode creates a source of noise equivalent to resistors at thousands of degrees Kelvin [11]. This substitution of shot noise for thermal noise is possible because the statistics of the random voltage produced by both are similar in time and frequency.

It is important to note the limitations of the approximations used in this section. First, the assumption that we are monitoring a pair of perfectly matched resistors is an approximation that can easily be violated. If the source of noise comes from a device with impedance that changes with frequency, some of the power will be reflected away. A common way to approach this problem in cabled systems, where most devices are well matched, is to characterize any impedance mismatch and use it in an uncertainty analysis. If the discrepancy is large, the previous equations can be modified to general complex impedance [10] or multi-port formulation even for nonreciprocal networks [12]. In addition, as indicated earlier, the basic equations that link noise power with temperature are derived from the one dimensional Planck's law. Planck's law requires that quantization of energy from each mode is taken into account at high energies or low temperatures. In particular equation 2 more generally reads [8]:

$$S_{v_{m}v_n}(f) = 4k_B T |Re(Z)| B \left[\frac{hf/k_B T}{e^{hf/k_B T} - 1} + \frac{hf}{2k_B T} \right] \quad (8)$$

where $S_{v_{m}v_n}(f)$ is the power spectral density of the series noise voltage, $|Re(Z)|$ is the magnitude of the real part of the complex impedance, and $\frac{hf/k_B T}{e^{hf/k_B T} - 1} + \frac{hf}{2k_B T}$ is the occupation of the electromagnetic mode with frequency f at a temperature T . This is very close

to 1 if $hf \ll k_B T$, and so does not normally appear in the room temperature analysis of frequencies lower than one terahertz. The equivalent rms voltage from equation 8 is:

$$v_{rms} = \sqrt{\int_{f_1}^{f_2} S_{v_m v_n}(f) df} \quad (9)$$

In conclusion, the standard Y-factor procedure is a powerful technique to determine the noise performance and gain of sensors, with limitations based on the assumptions of specific approximations.

1.2. Sensor Definition

The word sensor is an ambiguous term. In this technical note, we define the term sensor to mean a one-port device that measures power in a frequency selective manner. In practical terms, we can subdivide the sensor into two main components, the signal processing unit (SPU) and the preselector. The SPU is frequently a commercial off-the-shelf (COTS) spectrum analyzer, software defined radio (SDR), or fast analog-to-digital converter (ADC). The preselector is a set of attenuators, filters and amplifiers arranged to create the gain, noise performance and isolation desired. This type of sensor acquires in-phase and quadrature (IQ) information that can be further processed into power statistics with varying time and frequency characteristics. In Figure 1, the basic design is illustrated. In addition to the assorted radio frequency (RF) electronics, monitoring external signals requires an antenna. The common approach is to attach an antenna with known properties to the preselector. In order to accommodate the practical limitations of field installations, a coaxial cable of site-specific length and purpose-selected composition will normally be used to connect the antenna to the preselector. It is notable that the antenna will have an impedance that matches the preselector impedance (50Ω) to that of free space (377Ω) over a certain frequency range. The coaxial cable will have a loss (and phase that is ignored in this note) based on its construction. These characteristics of the antenna and its "feed" need to be accounted for. There are many ways to accomplish this accounting task, one being to measure the impedance (or scattering parameters) of the antenna directly with a vector network analyzer or equivalent device. The cable loss can be characterized in a similar manner, or by using a noise calibration from the end of the cable.

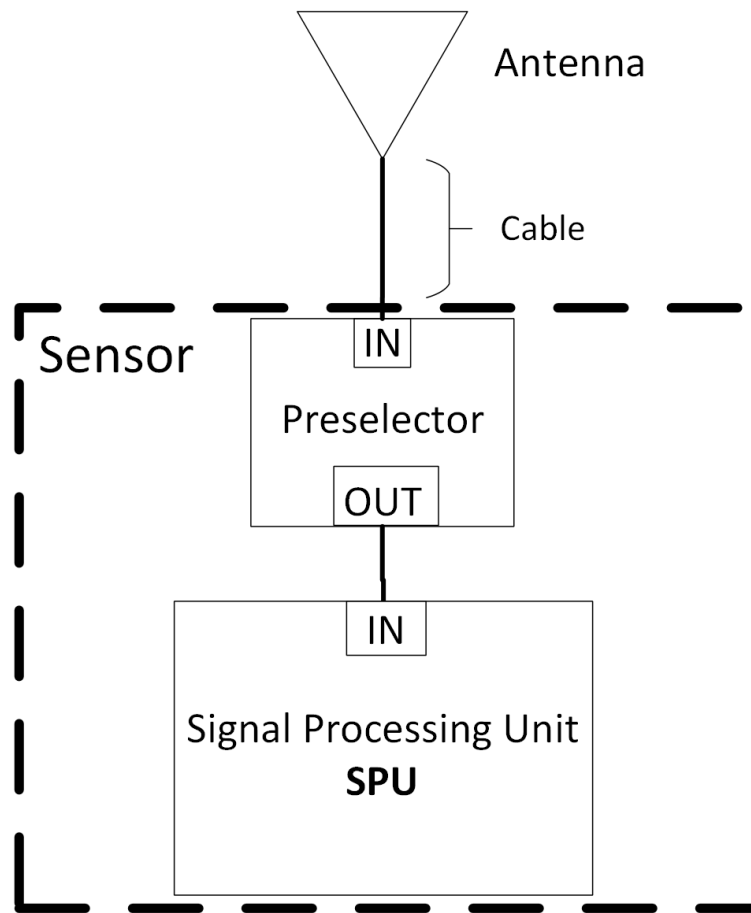


Fig. 1

1.3. NASCTN CBRS SEA Project

This technical note presents a synopsis of techniques employed in and results derived from a multi-year project of [NASCTN](#) to monitor an existing ecosystem, known as the [NASCTN CBRS SEA](#) project. The purpose of the [NASCTN CBRS SEA](#) project is to provide data-driven insight into the [CBRS](#) sharing ecosystem's effectiveness between commercial and [DoW](#) incumbent systems, and to track changes in the spectrum environment over time. [CBRS](#) is the "first of a kind" nationwide shared spectrum ecosystem in the 3550-3700 MHz band. The implementation of this project included the installation of sensor systems at selected sites on the east and the west coast of the United States, along with a control and prototyping system in Boulder, Colorado. In Figure 2, the sensor's locations are visualized. The sensor in Boulder, Colorado, provides a non-Environmental Sensing Capability ([ESC](#)) controlled reference point, monitoring an area not covered by a Dynamic Protection Area ([DPA](#)).

This document is not designed to be a comprehensive document about the [NASCTN CBRS SEA](#) project. Several documents are available in draft form [13, 14], published [15, 16] or

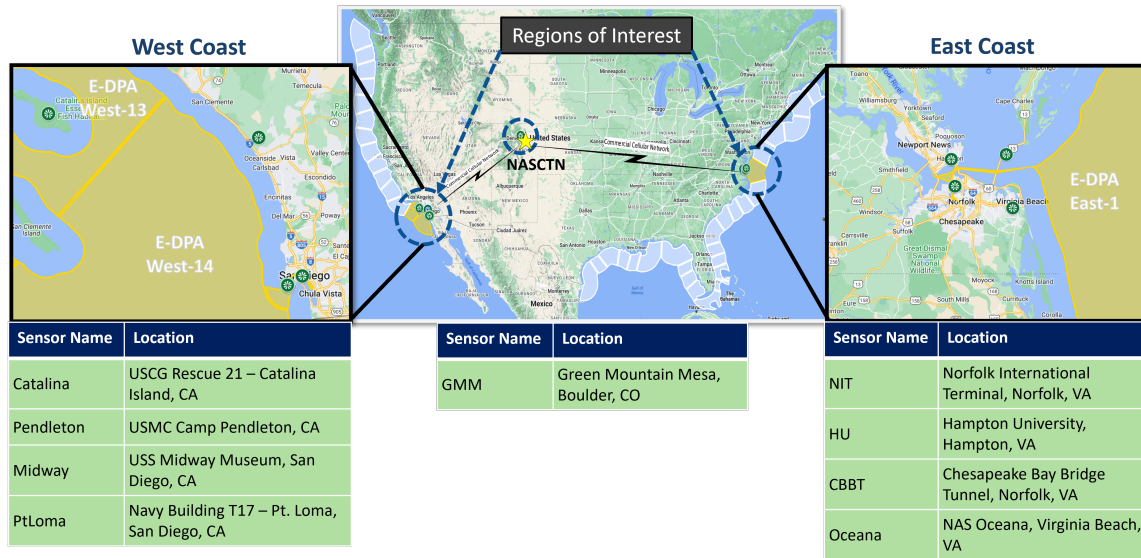


Fig. 2. Map of Deployed Sensors

are in preparation. In Figure 3, the theme, publication status at the time of this publication, and responsible agency are shown. This document supplements the series of previous and pending publications. The currently published material is summarized below.

1. **NASCTN CBRS SEA Website:** [NASCTN Project Website Hosted by NIST](https://www.nist.gov/programs-projects/cbrs-sharing-ecosystem-assessment), <https://www.nist.gov/programs-projects/cbrs-sharing-ecosystem-assessment>.
2. **NASCTN CBRS SEA Data Manual:** [NIST](https://nvlpubs.nist.gov/nistpubs/TechnicalNotes/NIST.TN.2359.pdf) technical note that describes data storage, format and calculation in detail, <https://nvlpubs.nist.gov/nistpubs/TechnicalNotes/NIST.TN.2359.pdf>
3. **NASCTN CBRS SEA Sensor Development:** The MITRE Corporation ([MITRE](https://www.mitre.org)) report that describes the sensor construction and characterization in detail. <https://www.mitre.org/sites/default/files/2026-01/PR-25-3159-spectrum-monitoring-sensor-radio-service-assessment.pdf>
4. **NASCTN CBRS SEA Data:** Web portal with data access instructions. <https://pages.nist.gov/SEA-DATA/>

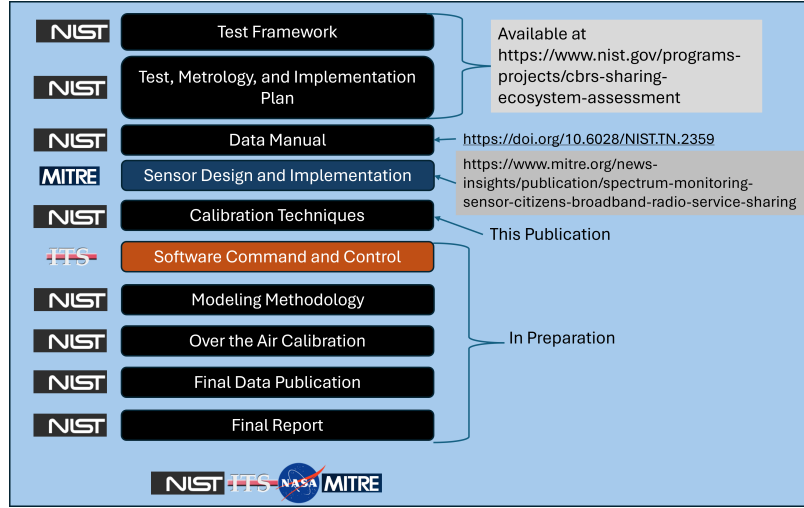


Fig. 3. NASCTN CBRS SEA Publications

2. General Discussion

2.1. Statistics of Noise Calibration

The derivation reiterated in section 1 implies for small excursions in frequencies each mode of the electromagnetic spectrum is uniformly occupied. This leads to a summation of modes each with a random phase, each with a uniform power spectral density versus frequency characteristic. This "flat" power spectral density versus frequency is colloquially known as white noise. Although the power spectral density of perfect white noise is flat everywhere, real sensor systems have different frequency characteristics, see figure 4

In the time domain, white noise has a voltage that is a random variable with a normal or Gaussian distribution with a zero mean $\eta(0, \sigma)$, see figure 5.

$$\text{prob}(v) = \frac{1}{\sqrt{2\pi}\sigma} e^{-\frac{v^2}{2\sigma^2}} \quad (10)$$

The rms voltage is then proportional to the standard deviation σ of the distribution. The power from two channels with a 90 phase difference, IQ, that are not correlated, is proportional to the sum of the variances σ^2 . This power distribution is itself a random variable with an exponential distribution, see equation 11.

$$\text{prob}(P) = \lambda e^{-\lambda P} \quad (11)$$

It should be noted that in this case the amplitude is a Rayleigh distributed random variable [17].

The procedure of calibration using a noise source is then one of specifying the variance of the voltage in units related to the macroscopic temperature. Specifically, if the variance of

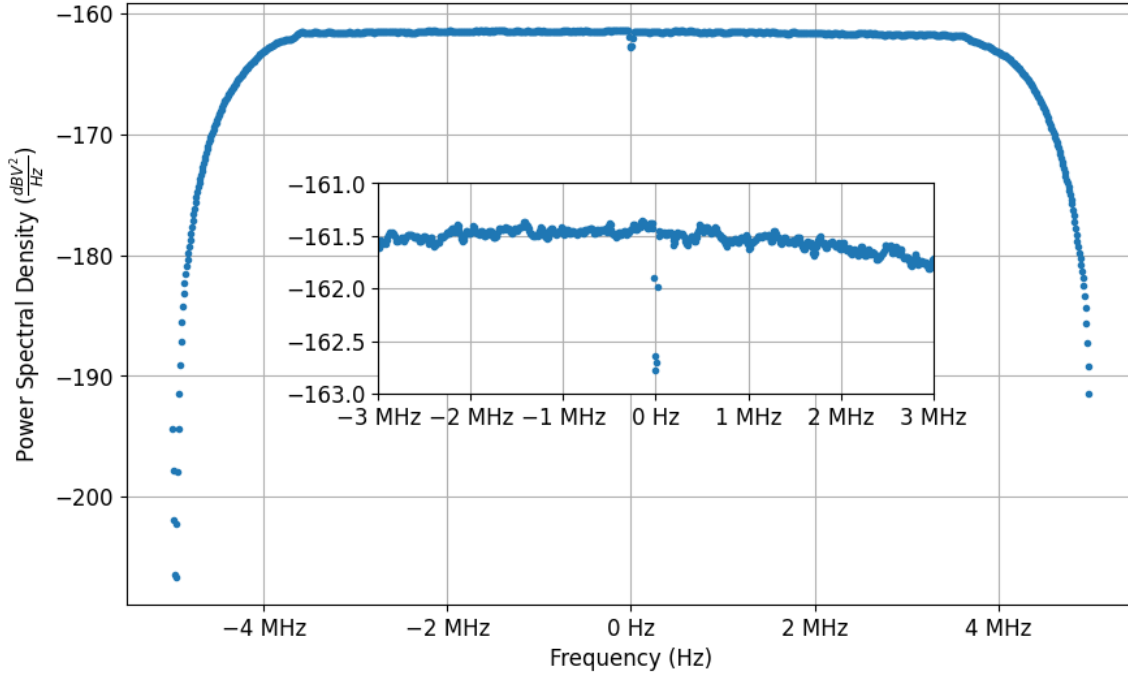


Fig. 4. Power Spectral Density of Noise Signal. The IQ used was an acquisition of a terminated receiver, demonstrating an instrumentation based roll-off at the edges of the acquisition frequency and local oscillator correction at 0 Hz.

I and the variance of Q are equal $\sigma_I = \sigma_Q = \sigma$ and the impedance is 50Ω then the power is

$$P = \frac{I^2 + Q^2}{2 \times 50} = \frac{\sigma_I^2 + \sigma_Q^2}{100} = \frac{\sigma^2}{50} = Gk_B T$$

$$\Rightarrow \sigma = \sqrt{50Gk_B T}$$

Furthermore, equation 11 can be rewritten in terms of the standard deviation of the voltage distribution

$$prob(P) = \frac{100}{2\sigma^2} e^{-\frac{100P}{2\sigma^2}} \quad (12)$$

This leads to an interesting property of the power histogram when shown with a y-axis $\log(P)$ and an x-axis of linear power. The histogram will be a line with slope $-\frac{100}{2\sigma^2}$, see figure 6.

Additionally, a few ways to estimate the value σ present themselves. We could take the standard deviation or variance of the I or Q values. We could fit the linear slope of the power histogram. Or we could estimate it from the amplitude ($|IQ|$), but that requires

In-phase (real) component

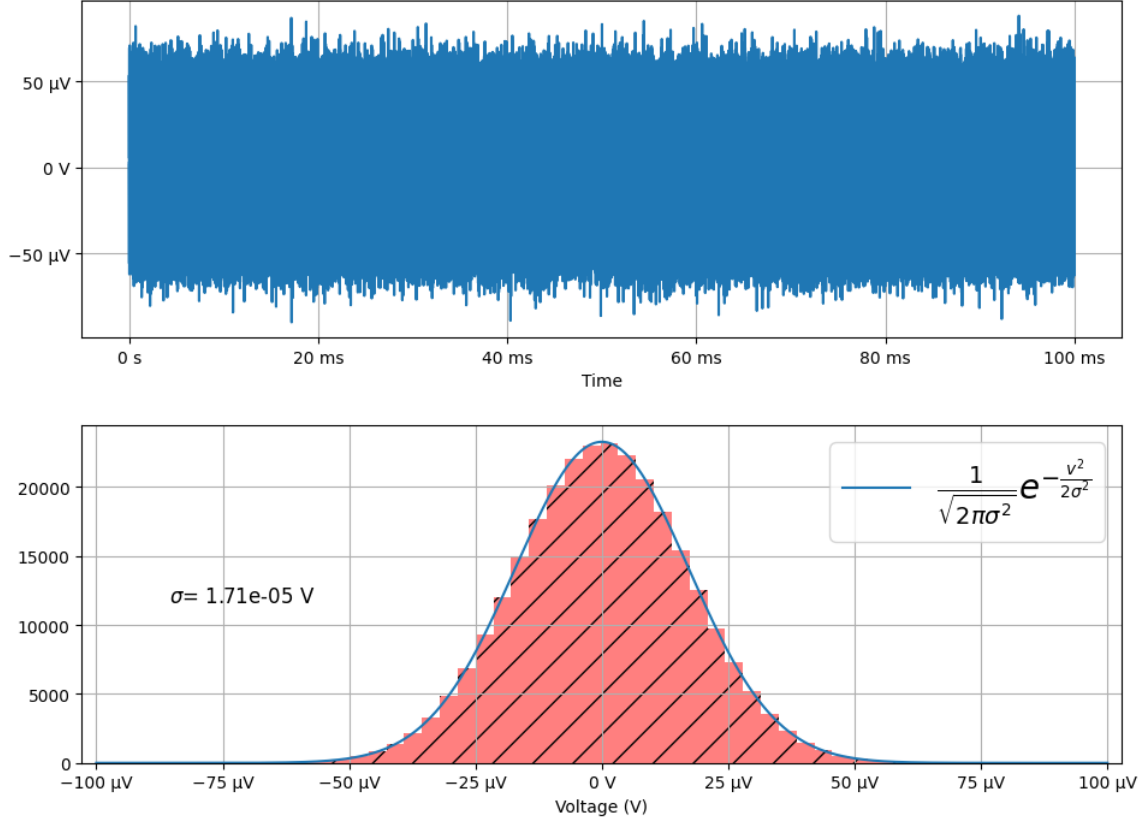


Fig. 5. In-phase voltage as a function of time (top) and as a histogram in voltage (bottom). The voltage is a random variable with a Gaussian distribution. Data is measured while the receiver is terminated.

knowledge of the Rayleigh distribution $\sigma = \sqrt{\text{variance}(|IQ|) \frac{2}{4-\pi}}$ [18, Chapter 33, pg 618]. In Figure 7, the results of those calculations are presented graphically.

The exponential power distribution has several properties that are sometimes overlooked. First, the expected value of the Peak-to-Average Power Ratio (**PAPR**) depends only on the number of samples measured [17].

$$PAPR \approx \ln(n) + \gamma \quad (13)$$

where $\ln(n)$ is the natural log of the number of samples and γ is Euler's constant 0.577... . A large deviation from this curve indicates an issue in the acquisition of **IQ**. In particular, it indicates a distortion in the shape of the probability distribution function of the power. Although the **PAPR** does not indicate the cause of this distortion, the most common causes are the presence of an interference signal, the sensor being too sensitive (overload), the

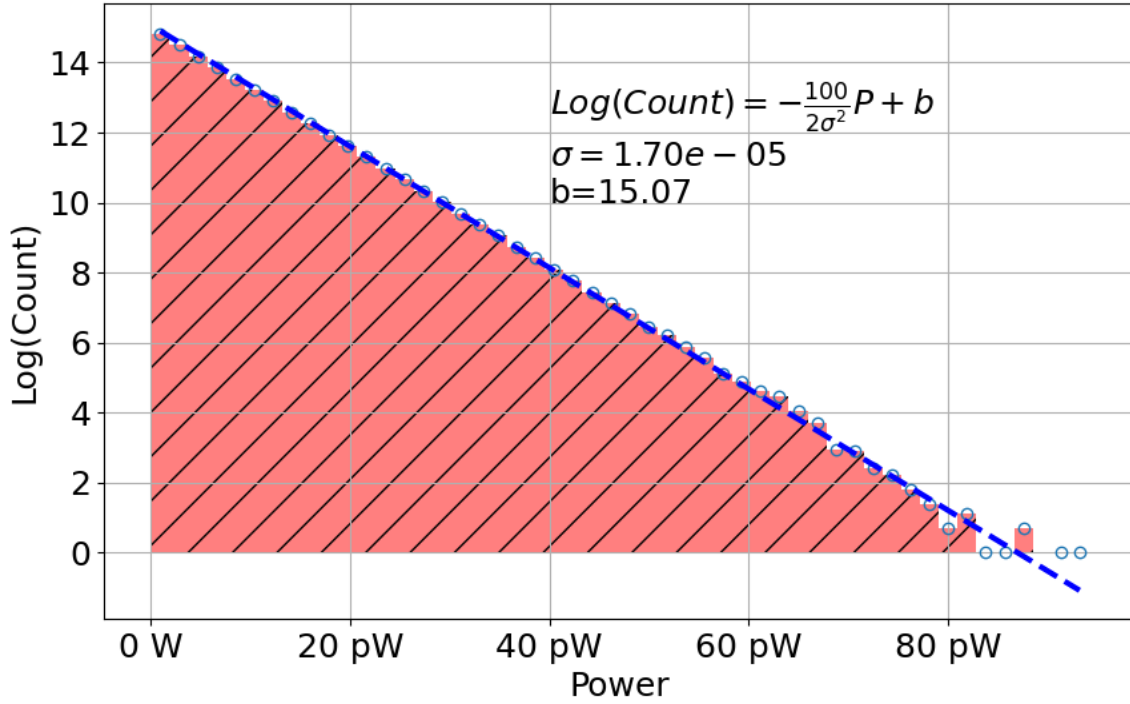


Fig. 6. Histogram of Power for Noise. The exponential probability distribution in power leads to a log-linear characteristic with slope $-\frac{100}{2\sigma^2}$, where sigma is the standard deviation of the voltage of I or Q, data is measured with receiver terminated.

sensor not being sensitive enough (quantization error) or distortion from digital signal processing. This is a powerful indication if non-exponentially distributed samples are present. The median and mean (average) of the exponential distribution are closely related, and are within a constant factor. This means that if the Average-to-Median Power Ratio (**AMPR**) is sampled over a period being used to calibrate the ratio should approach the constant value $\frac{1}{\ln(2)} \approx 1.443$, or in decibels $10 \text{Log}(\frac{1}{\ln(2)}) \approx 1.592$. To illustrate this point, a simulation of power calibration results is presented in Figure 8. We first assume that the **IQ** captured for calibration has the following properties:

- A bandwidth of 10 MHz
- A length of capture of 1 second
- A total number of 10^7 **IQ** points
- A standard deviation of $10 \mu V$, or equivalently a gain of 40 (16 dB), and a $T_{\text{sensor}} + T_{\text{thermal}}$ of $362^\circ K$

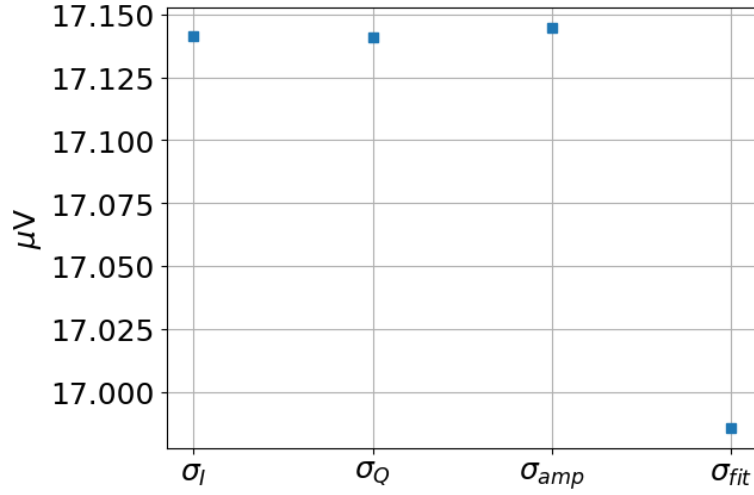


Fig. 7. Comparison for Estimating the σ Parameter. Each estimate is derived from a different source, σ_I is the standard deviation of I, σ_Q is the standard deviation of Q, σ_{amp} is $\sqrt{\text{variance}(|IQ|)\frac{2}{4-\pi}}$ and σ_{fit} is derived from the slope of the log-linear fit of the power probability distribution.

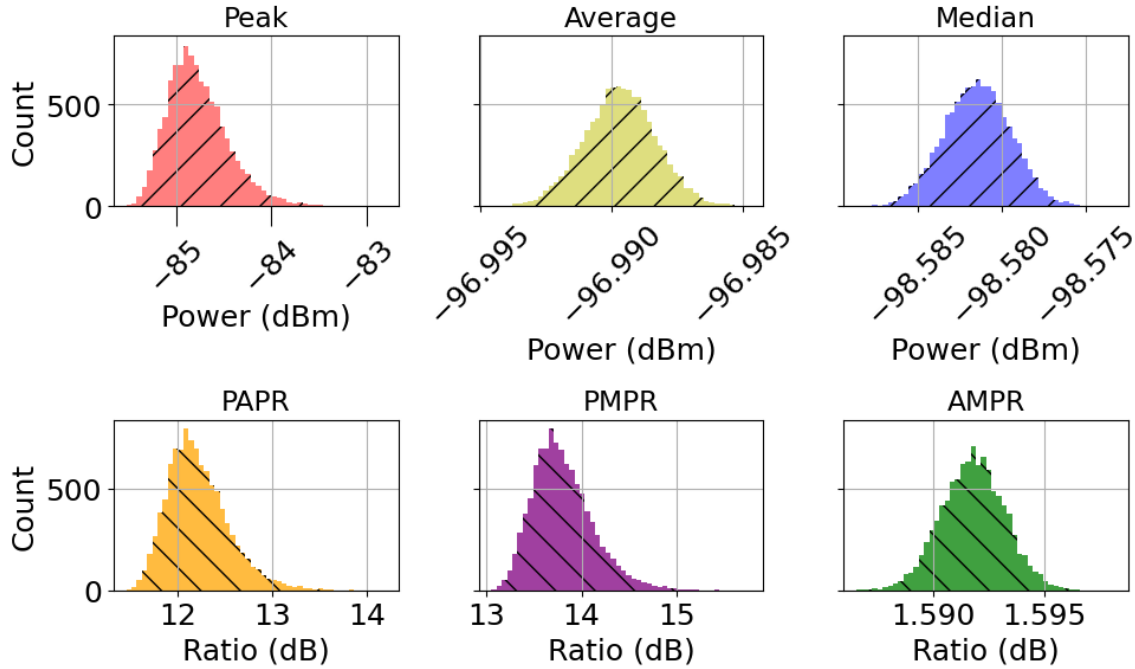


Fig. 8. Distribution of Simulated Power and Ratios. Each histogram is the result of ten thousand Monte Carlo simulated measurements assuming 10^7 points in each measurement.

2.2. Presence of Interfering Signals

Whenever a calibration is performed, it is important that the only signal received by the sensor system is the noise from the thermal source or the calibration noise diode. In the modern world, other signals abound, especially in the mid-band region. The frequency ranges around Wi-Fi systems (2.4 GHz and 5 GHz) are a particular issue (these are known as the industrial, scientific and medical (ISM) bands). Even in the most isolated laboratory environments, the sensor itself can create sources of interference. These sources of interference can cause a spurious signal to overtake the desired source of noise and lead to a poor calibration. Fortunately, most signals due to communication systems, the most prolific culprit, are not exponentially distributed in power, and hence violate the expected values of PAPR, Peak-to-Median Power Ratio (PMPR) or AMPR. For example, data with 10^6 points was taken from several different IQ sources and plotted in figure 9. The exponential simulation shows the expected PMPR and AMPR for an exponentially distributed power (the calibration signals). The conducted long-term evolution (LTE) are calculated from the experiment documented in [19], and are of a cell phone and base station emulator coaxially connected, and represent different configurations documented in the technical report. The radiated LTE represents 25 randomized samples of a single configuration documented in [20]. The experimental noise data represents a terminated measurement using the apparatus documented in [21]. The practical observation is that the average to median ratio should always be within ± 0.5 dB of 1.592. A more stringent criteria would put a limit on the either the PAPR or PMPR.

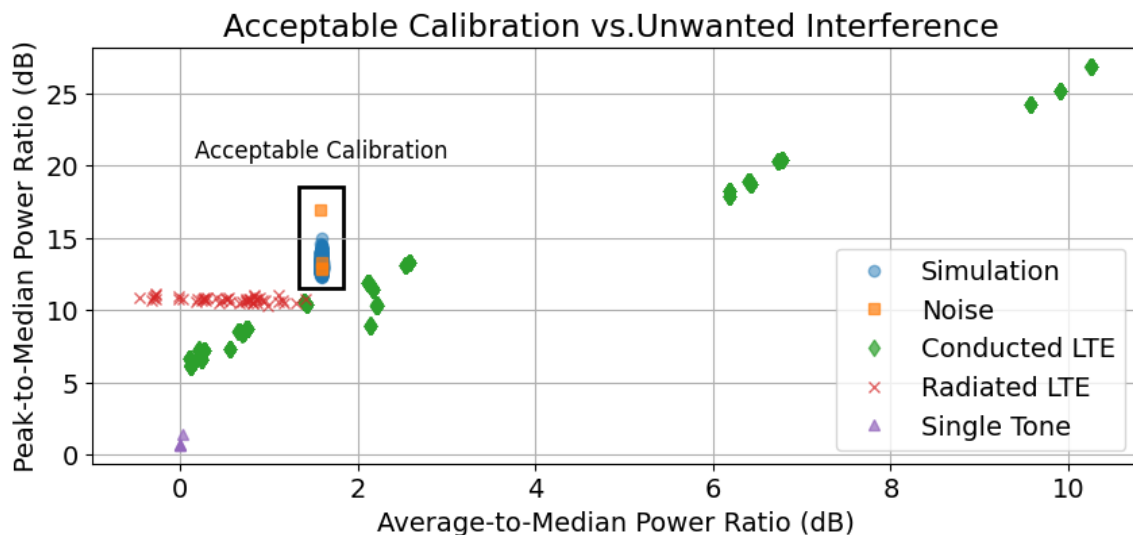


Fig. 9. Using PMPR and AMPR to Determine Acceptable Calibrations. A region around 1.592 in AMPR and between 11.5-18.5 dB in PMPR for 10^6 points can be used to reject interference corrupted IQ measurements

2.3. Effects of Saturation

When a signal has sufficiently high amplitude, it can cause errors to occur in the sensor. The incident power at which this occurs can be determined by introducing a progressively more intense signal at the input and recording the measured output power derived from the sensor. The measured output power should be a linear function of the power of the input signal. If the input signal does not exceed the intrinsic noise of the system there will be no response, which is not an error. However, if the signal exceeds the linear amplification region of the preselector, the calculated power will no longer conform to the equation 4. In Figure 10 each of these cases are visualized. The determination of P_{sat} is subjective, but is frequently chosen to be 1 dB gain compression point (P1dB) or the point that the measured power is 1 dB lower than expected from the equation 1. Additionally, if the voltage of the signal arriving at the ADC in the SPU exceeds the reference voltage of the ADC, the ADC will either return the maximum allowed voltage or enter into an error state, see for example [16, Figure 13.]. To record accurate power values, the sensor must always be in a linear region, or extreme caution and extra characterization must be preformed. Techniques that correct for non-linear behavior, while possible, are not best practice and are not addressed in this technical note.

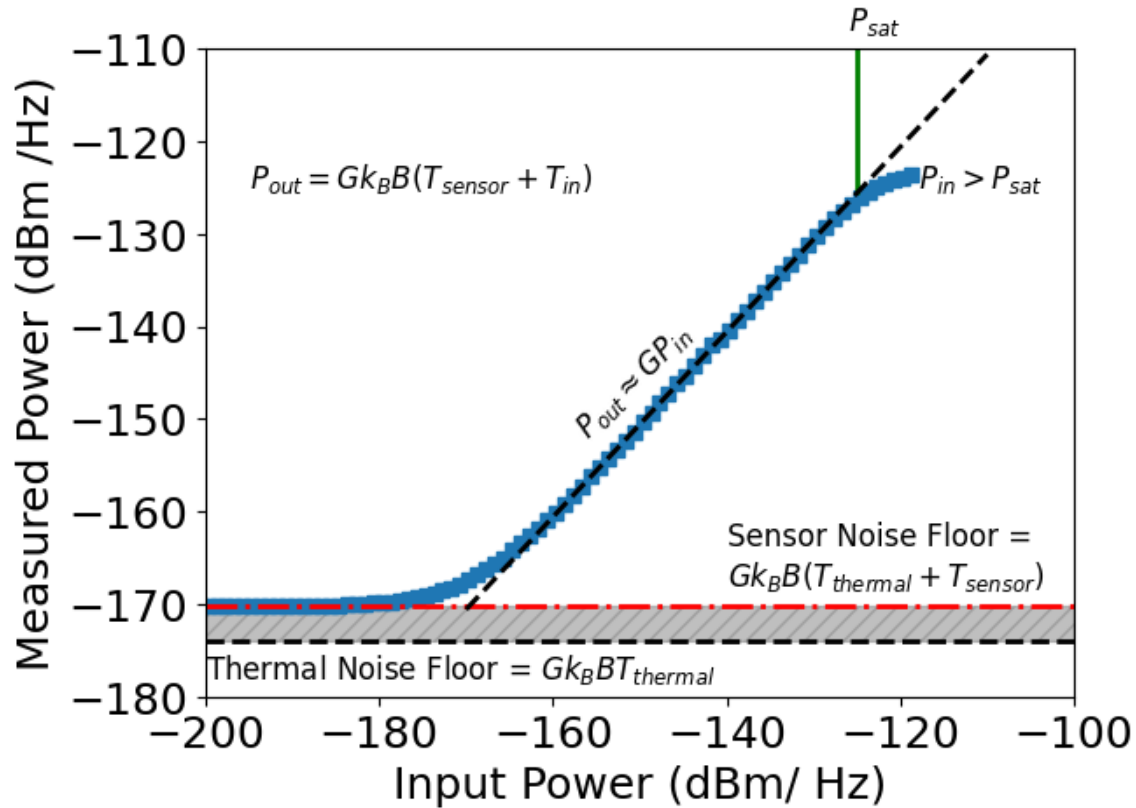


Fig. 10. Sensor Measured Power Showing Saturation

2.4. Quantization Errors

In general, the digitization of the [IQ](#) voltage is a process in which a reference voltage is subdivided by the number of bits in the [ADC](#). If this reference voltage is very large compared to the noise signal, effects of the digitization process become apparent. In figure 11, an example of the in-phase voltage for a terminated input of a vector signal transceiver ([VST](#)) are shown. In the left panel of Figure 11, the reference power (which determines the reference voltage) is chosen to be 20 dBm. This value is very large compared to the thermal noise power leading to discrete values of the voltage. In the right panel, a reference power of -50 dBm is chosen and the quantization error disappears. This type of error typically results in a reduced peak value, and thus reduced [PAPR](#) and [PMPR](#), a detailed effects are covered in [17, Section IV. A].

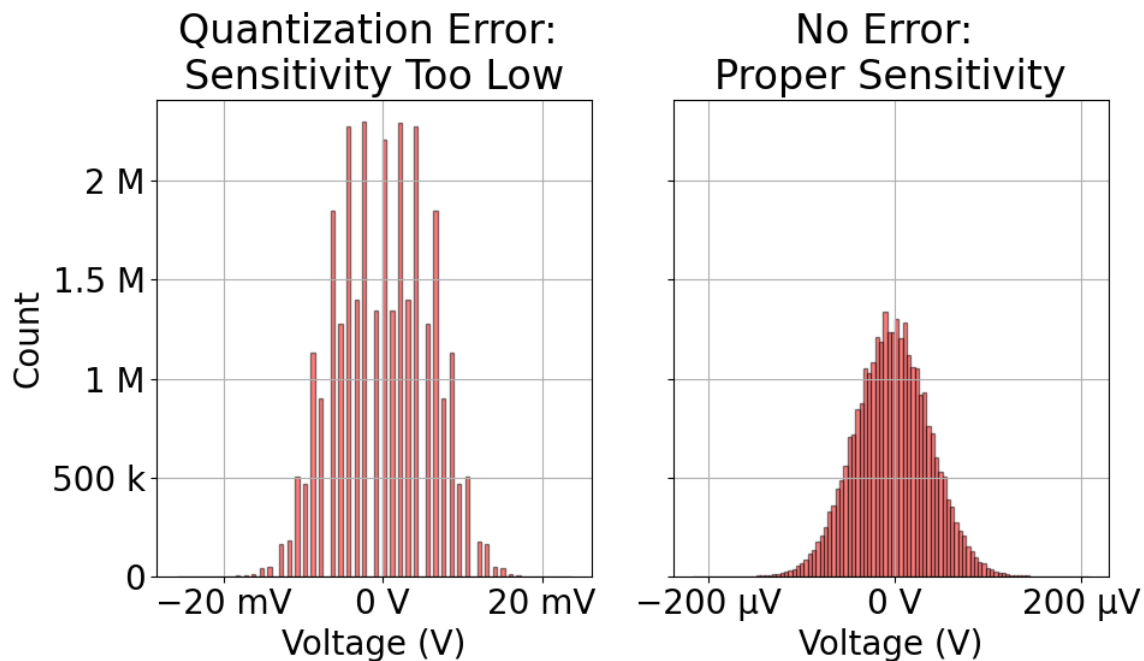


Fig. 11. Comparison of Signal with and Without Quantization Error

2.5. Distortion Due to Signal Processing

In our definition of a sensor, see section 1.2, we defined the [SPU](#) to be some sort of device that records [IQ](#) data. We did not, however specify that it be free of post-processing math (digital signal processing ([DSP](#))), or pre-processing from elements contained in the [SPU](#). In rare occasions, this processing can cause issues with calibration. If, for example, the processing changes the measured [IQ](#) in a way that affects the root mean square ([RMS](#)) voltage, the calibration can be erroneous. Most [DSP](#) techniques are focused on averaging or frequency selection and these techniques may linearly affect the measured signal, but

will be compensated for by the calibration process. However, if one is using the techniques described in section 2.2, the signal processing can cause spurious results.

3. Calibration Probe

A noise source with a fixed noise temperature has several limitations. This section is devoted to the construction and characterization of a calibration probe, or amplified programmable noise source, that addresses many of these limitations. For example, one clear limitation is the range over which one can use that diode to characterize the noise floor of a device like the one shown in Figure 1. The Y-factor measurement requires a differential change in the sensor reading for an active noise diode and a room temperature load. This difference is used with the absolute ratio of the noise temperature of the diode to the ambient temperature to characterize the noise floor (noise figure, noise temperature, etc), hence if the noise temperature of the system is significantly higher than the noise temperature of the diode while active, the characterization starts to falter. In the case that the system to be characterized has a much higher noise temperature there can be a null response and the method fails all together. This means for noisy systems, like non-amplified broadband signal analyzers or software defined radios in low-gain states, the typical noise diode that has an ENR of 15 dB might not be intense enough to measure the noise and gain. The common solution is to either move to a tone-based approach or use a much more intense noise source. To optimize for both low and high noise figure, the calibration probe uses a programmable attenuator that is added to an amplified source. In addition, the source impedance of some noise diodes can be non-ideal matches to the network to be characterized [22], to improve this an inline attenuator or isolator can be added. Finally, some noise diodes and their respective power supplies can be slow to respond to power cycling. This can cause a reading of the energized state of the diode to be extremely slow for all the frequencies of interest, or erroneous when done too fast. To avoid this, the calibration probe's noise source can be always on and for ambient noise measurements a large attenuation can be set, causing the thermal noise to be dominant. In figure 12, a simplified block diagram of the calibration probe is shown. Two designs of calibration probe have been constructed and characterized with the transfer system outlined in Section 4. The components used in the first design are detailed in Table 1, three devices of this design were realized (see Figure 13), referenced as probe 1, probe 2 and probe 3. In Figure 15, the ENR of these probes at a programmable attenuator setting of 30 dB are shown. Similarly, the components used in the second design are detailed in Table 2, two devices of this design were realized (see Figure 14, referenced as probe 4 and probe 5. In Figure 17, the ENR of these probes at a programmable attenuator setting of 30 dB are shown.

The calibration probe generates a known noise temperature for a given attenuation level at a given frequency. This allows for an extended power out versus input temperature characterization for instance a curve like the one in Figure 10 can be generated with a single connection.

Table 1. Probe 1,2,3 Components

Designator	Description	Manufacturer	Part Number
AMP	Low noise amplifier, 3300 MHz to 3800 MHz	Minicircuits	ZX60-3800LN-S+
Isolator	Isolator, 2-4 GHz	CentricRF	CI2040
Noise Diode	Noise source, 10 MHz to 18 GHz	Keysight	346B
Programmable Attenuator	Programmable Attenuator, 110 dB, 1 MHz - 6 GHz	Minicircuits	RUDAT6000-110

Table 2. Probe 4,5 Components

Designator	Description	Manufacturer	Part Number
AMP	Low noise amplifier, 1800 MHz to 6000 MHz	Minicircuits	ZX60-63GLN
Isolator	Isolator, 2-6 GHz	Fairview	SFI0206S
Noise Diode	Noise source, 10 MHz to 18 GHz	Keysight	346B
Programmable Attenuator	Programmable Attenuator, 110 dB, 1 MHz - 6 GHz	Minicircuits	RUDAT6000-110

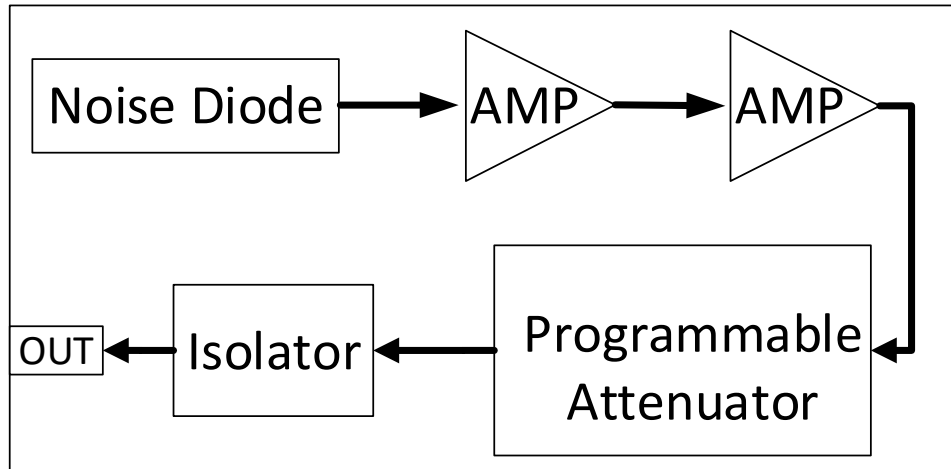


Fig. 12. Block Diagram of Amplified Programmable Noise Source

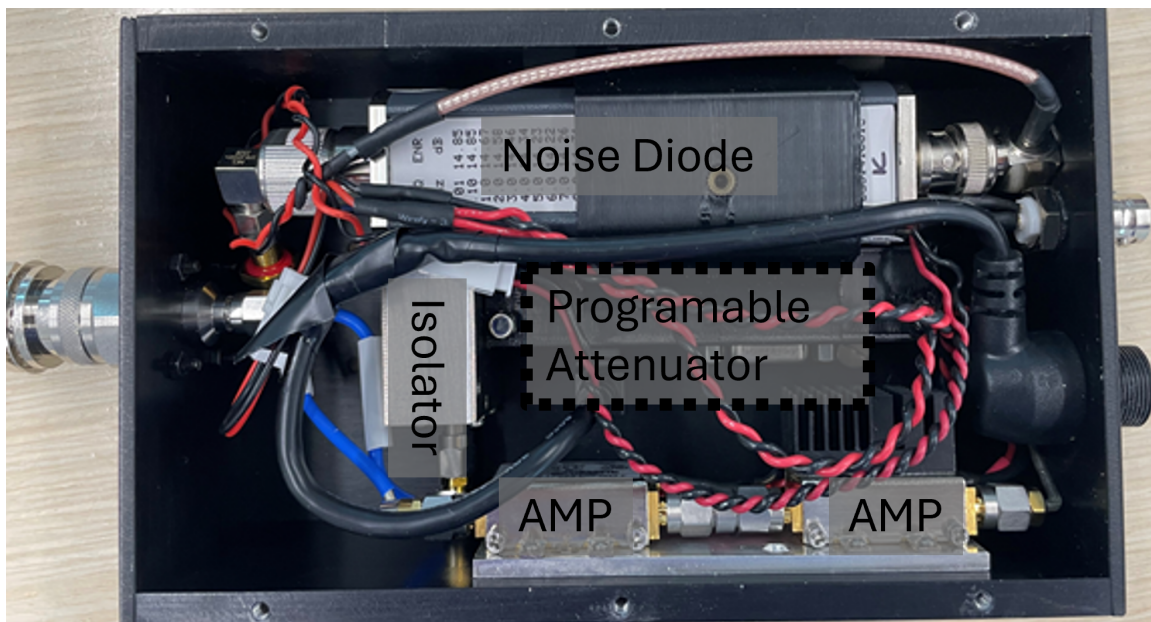


Fig. 13. Image of Realized Calibration Probe for 3-4 GHz

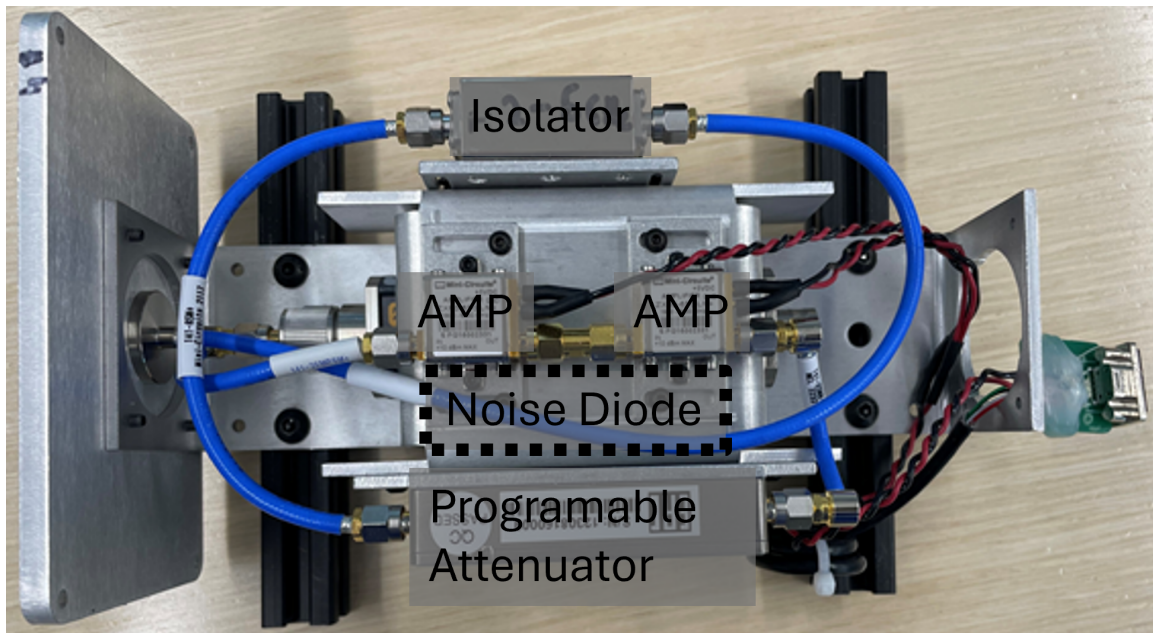


Fig. 14. Image of Realized Calibration Probe for 2-4 GHz

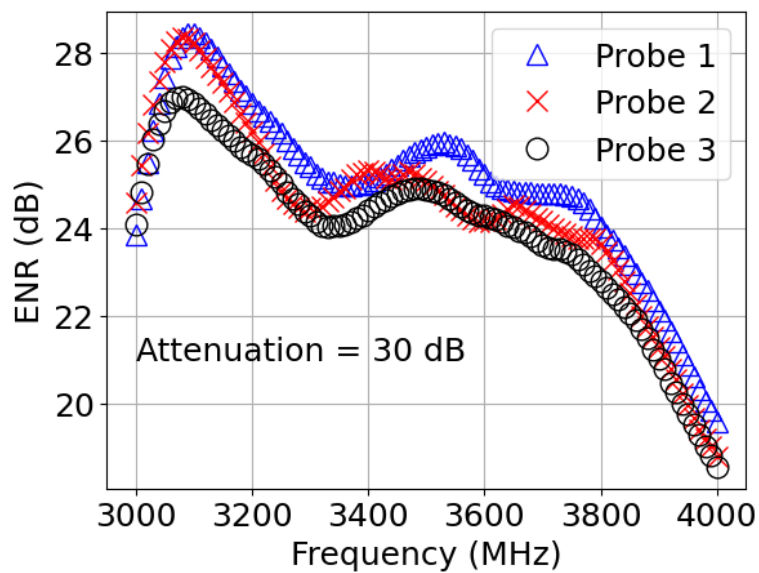


Fig. 15. ENR of Calibration Probes 1,2,3 at 30 dB attenuation

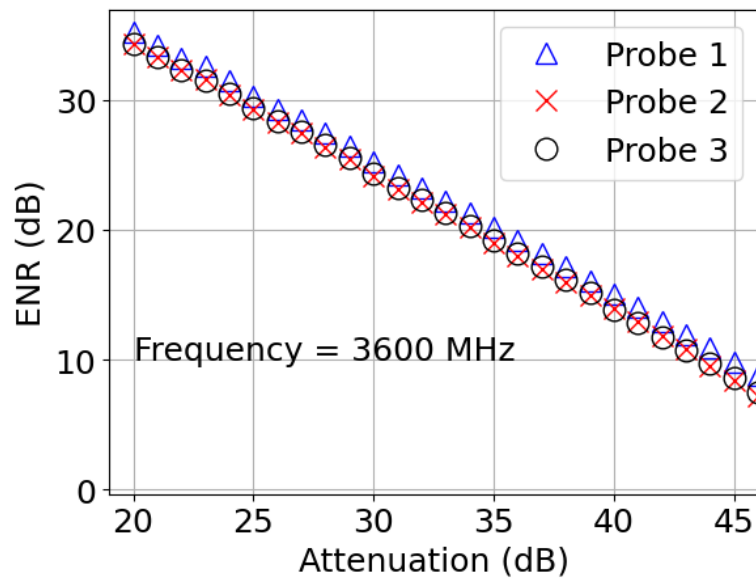


Fig. 16. ENR of Calibration Probes 1,2,3 at 3600 MHz for Attenuations 20 to 46 dB

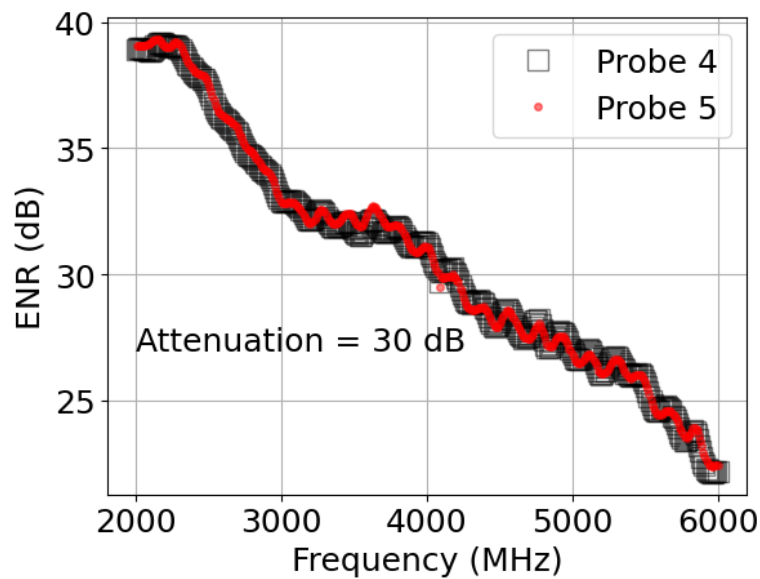


Fig. 17. ENR of Calibration Probes 4,5 at 30 dB attenuation

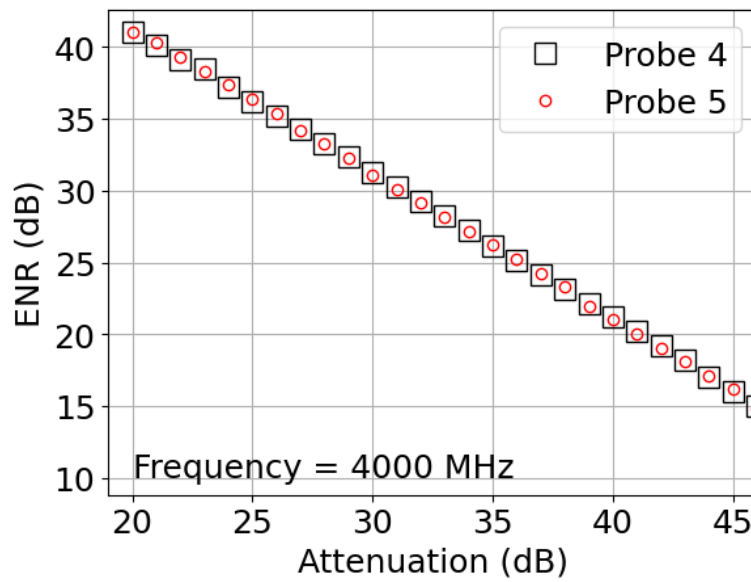


Fig. 18. ENR of Calibration Probes 4,5 at 4000 MHz for Attenuations 20 to 46 dB

4. Transfer System

A transfer system is any system capable of relating a measurement of a standard to an unknown. Here we explain how to use a single artifact (a noise diode) to then measure the noise temperature of another device, such as the amplified programmable noise source. A viable realization of a transfer system is a combination of an amplifier and a software defined radio or spectrum analyzer. For example in Figure 19, a block diagram of a transfer system is shown. The definition of sensor presented in section 1.2, is also an example of a transfer system.

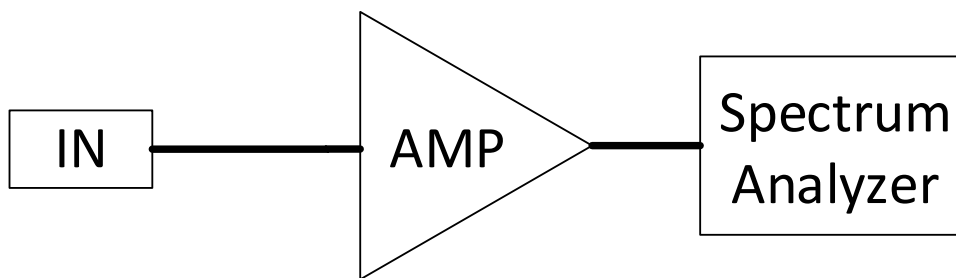


Fig. 19. Block Diagram of Transfer System

The system can first be calibrated using the Y-factor method. In order to accomplish this, a mechanism for energizing and de-energizing the noise diode is required. In this example, a programmable power supply is used for automation, but a manually operated switch would also be suitable. The frequencies of interest are selected, along with the operational bandwidth¹. Here we performed measurements of four noise diodes to both calibrate and form an uncertainty estimate. The manufacturer calibration data for each noise diode is shown in Figure 20. The four noise diodes were measured as outlined in section 1 for each frequency and at a sample rate of $10^7 \frac{\text{samples}}{\text{s}}$ for a period of 100 ms. The power for the on and off state of the noise diode was recorded, and is shown in Figure 21. Using the formulas in section 1 the gain and noise figure was calculated, the result of these calculations are shown in Figure 22. The uncertainty of this process is dictated by the transfer system configuration and the device used to calibrate it. In section 6.7, a full accounting of uncertainties for deployed sensors is outlined, including frequency dependence. In most situations that require a transfer system, multiple years of calibrating every hour is not practical, so the uncertainty needs a different approach. In order to estimate the total uncertainty as a single number for all frequencies, multiple measurements of different artifacts (noise diodes and calibration probes), the frequency variation of the gain, and the reported manufacturer's uncertainty for the calibration device are all considered. Here any

¹Note the sample rate of the IQ collection might not be the effective noise bandwidth due to rf front-end and processing in the system

gain values failing an outlier rejection step were removed. Outliers are gains falling outside of the 25th percentile minus 1.5 times the interquartile range and the 75th percentile plus 1.5 times the interquartile range, following the Tukey fence method [23]. Most outliers were caused by the noise diode not turning on due to command and control issues with the programmable power supply. The estimated uncertainty was calculated as follows:

1. the maximum of the maximum of the difference of gain for neighboring frequencies was calculated for each noise diode (0.44 dB)
2. twice the maximum standard deviation of the difference of gain for different noise diode calibrations was calculated (0.08 dB)
3. the manufacturers estimate for all frequency uncertainty (0.2 dB)

The total uncertainty is determined by Monte Carlo simulation [24], with 10,000 trials. Errors are assumed to be multiplicative so that they are additive under log transformations. The simulation used a uniform probability distribution for elements 1 and 3 with endpoints equal to $\pm$ Error, and a normal distribution for 2 with zero mean and standard deviation of 0.08 dB. The resultant distribution is normal with a standard deviation of 0.29 dB. We therefore estimate the total error as twice the standard deviation or **0.6 dB**. This estimate is of maximum total error for all frequencies.

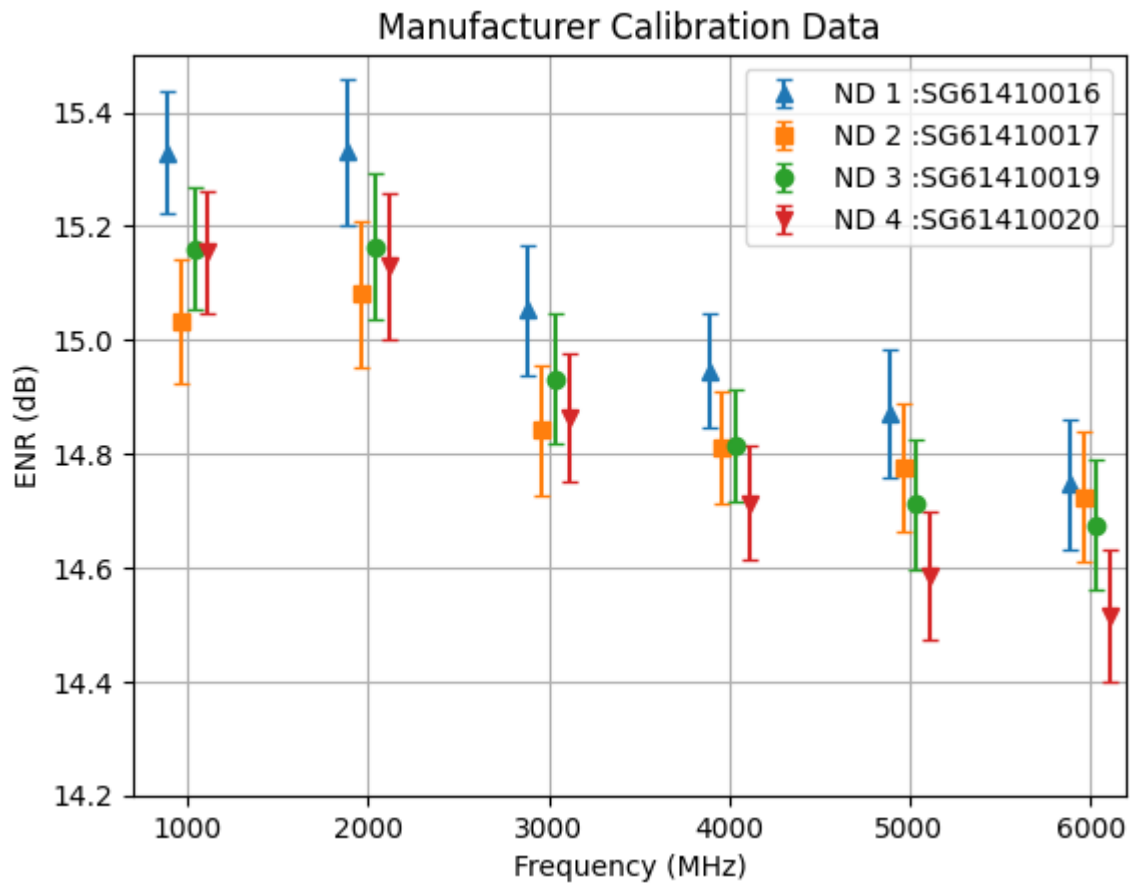


Fig. 20. Noise Diode Manufacturer Data. Markers have been horizontally offset for clarity, reference temperature is $T=290\text{K}$.

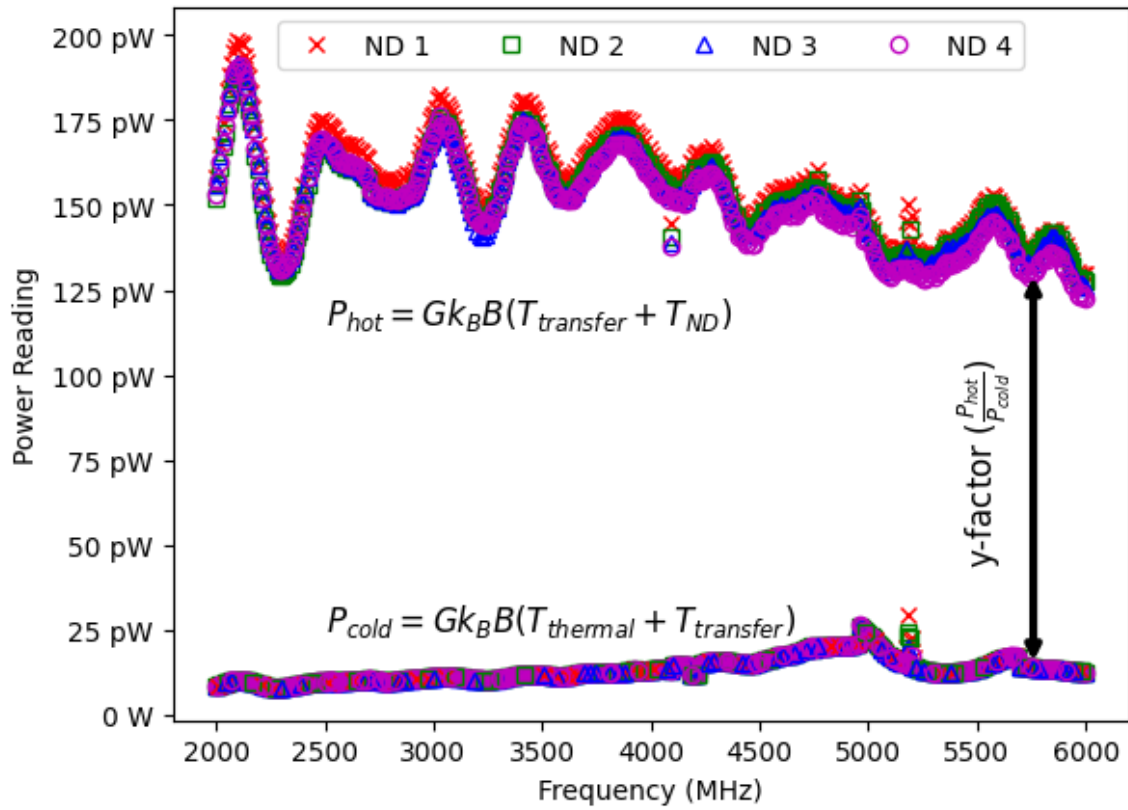


Fig. 21. Y-Factor of Transfer System from Multiple Noise Diodes . Data is acquired at 10 MHz Bandwidth, $T_{thermal} = 290^{\circ}K$

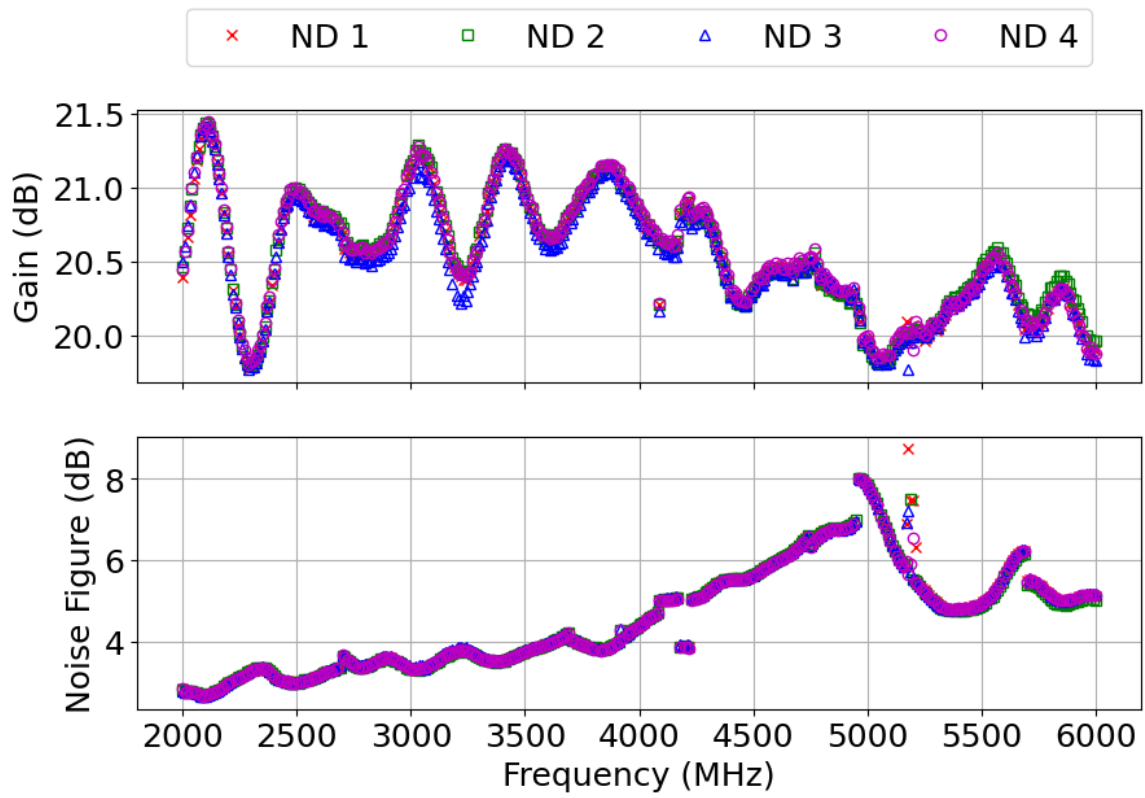


Fig. 22. Gain and Noise Figure of Transfer System

5. Noise Source Radiometer Characterization

Each source of noise used to calibrate a device must be measured to determine its effective noise temperature. In the previous section, section 4, methodology to create a transfer system from COTS components was discussed. This type of transfer system, while effective, has limited temporal stability and needs frequent calibration. In addition, the system is dependent on the calibration artifacts selected, which are themselves transfer standards from the manufacturer of choice's calibration laboratory. An important goal of the NASCTN CBRS SEA project was to minimize the uncertainty of the project's sensors and provide traceability. In order to accomplish this, the calibration artifacts used in the sensors were independently characterized by the NIST noise calibration service. This provides two benefits, a reduction of uncertainty, and the traceability of the artifacts to a temperature standard. For the NASCTN CBRS SEA project, 15 commercial noise diodes and 2 calibration probes were characterized. 10 of the commercial diodes were installed into the preselectors of sensors deployed, see Figure 2. The NIST radiometers are double-sideband, total-power radiometers. The IF frequency is 0 (i.e., the LO frequency is set to the measurement frequency), and the IF bandwidth (BIF) ranges from 5 MHz to 20 MHz, depending on the particular radiometer. Thus the reported noise temperature represents an average over a frequency range of 2BIF centered at the measurement frequency. The noise-temperature measurements are performed using two thermal noise standards [25], one of which is at ambient temperature and one of which is at cryogenic (liquid nitrogen) temperature. These national standards served as primary standards traceable to the SI unit kelvin (through calibrated thermistors and the boiling temperature of liquid nitrogen) before its redefinition in May 2019 [26]. Any adapter required in the measurement is characterized and corrected for using the methods in [27, 28]. The two characterized calibration probes were measured at a set of attenuation values for frequencies of interest, In Figures 23 and 24 the result for calibration probes 1 and 3 respectively, see Section 3, are reported. The noise diodes characterized for deployed sensors are summarized in Table 3 along with the manufacturers ENRs at 3000 and 4000 MHz. These noise diodes were characterized in three separate measurement batches. Batch 1, measured in December, 2022, are summarized in Table 4. Batch 2 measured in October, 2023 are summarized in Table 5. Batch 3 measured in August, 2024 are summarized in Table 6.

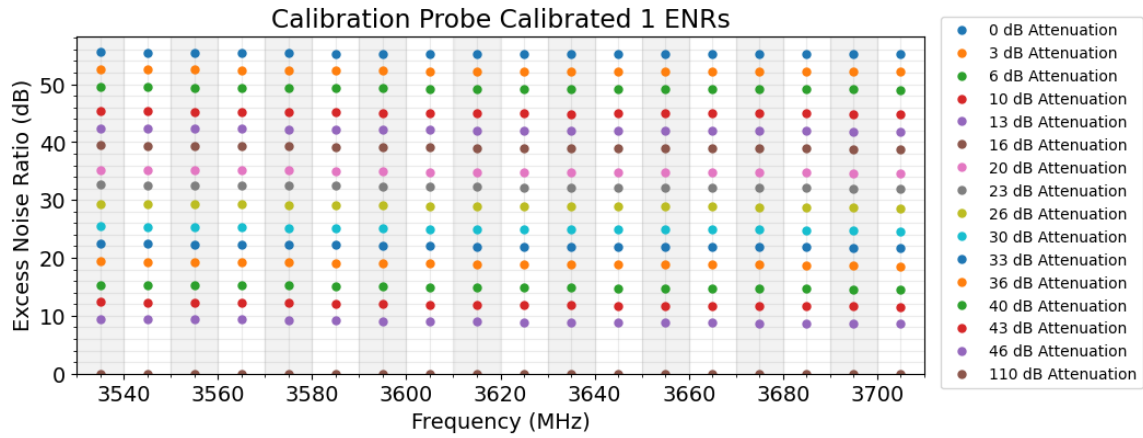


Fig. 23. Measured ENR for Calibration Probe 1 from NIST Radiometer

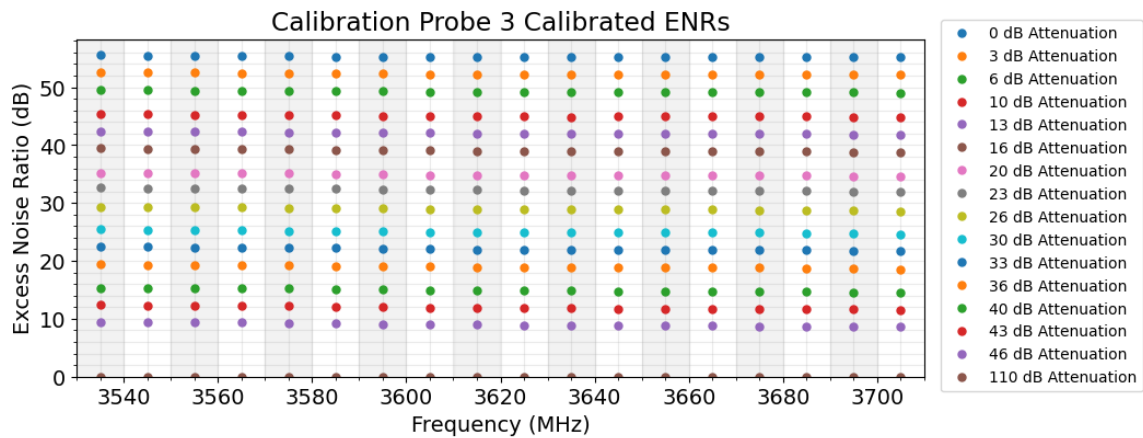


Fig. 24. Measured ENR for Calibration Probe 3 from NIST Radiometer

Table 3. Noise Diode Manufacturers ENR for Comparison

Sensor Name	Noise Diode SN	3000 MHz ENR (dB) ± 0.20 dB	4000 MHz ENR (dB) ± 0.20 dB
Midway	MY61410090	14.87	14.78
NA	MY61410089	14.99	14.89
NA	MY61410080	14.84	14.71
NIT	MY58230806	14.74	14.64
HU	MY61410041	14.70	14.67
GMM (pre-production)	MY61410032	14.64	14.58
Catalina	SG61410026	15.04	14.93
CBBT	SG61410024	14.91	14.81
Oceana	SG61410023	15.18	15.13
Pendleton	SG61410021	14.96	14.85
GMM (production)	SG61410011	14.77	14.72
NA	SG61410012	14.85	14.78
NA	SG61410016	15.05	14.95
NA	SG61410017	14.84	14.81
NA	SG61410019	14.93	14.81

Table 4. Batch 1 Radiometer Noise Diode Characterization

Frequency	MY61410090	MY61410089	MY61410080	MY58230806	MY61410041	MY61410032
MHz	ENR (dB) ± 0.05 dB					
3500	14.77	14.90	14.74	14.64	14.66	14.54
3550	14.81	14.94	14.78	14.68	14.70	14.58
3600	14.78	14.91	14.75	14.65	14.68	14.55
3650	14.78	14.91	14.74	14.65	14.68	14.55
3700	14.76	14.90	14.73	14.63	14.67	14.54

Table 5. Batch 2 Radiometer Noise Diode Characterization

Frequency	SG61410026	SG61410024	SG61410023	SG61410021	SG61410011
MHz	ENR (dB) ± 0.05 dB				
3535	15.00	14.89	15.17	14.92	14.77
3545	15.01	14.89	15.17	14.92	14.77
3555	15.00	14.89	15.17	14.92	14.77
3565	15.00	14.89	15.17	14.92	14.77
3575	14.99	14.88	15.16	14.91	14.76
3585	14.99	14.88	15.15	14.90	14.75
3595	14.98	14.86	15.14	14.91	14.75
3605	14.98	14.86	15.14	14.89	14.74
3615	14.96	14.85	15.13	14.89	14.74
3625	14.97	14.86	15.13	14.89	14.73
3635	14.97	14.85	15.13	14.89	14.73
3645	14.98	14.87	15.14	14.90	14.75
3655	14.97	14.86	15.14	14.90	14.75
3665	14.99	14.88	15.15	14.90	14.75
3675	14.98	14.87	15.15	14.90	14.75
3685	14.97	14.87	15.14	14.89	14.74
3695	14.96	14.85	15.13	14.88	14.73
3705	14.95	14.84	15.12	14.87	14.72

Table 6. Batch 3 Radiometer Noise Diode Characterization

Frequency	SG61410009	SG61410012	SG61410016	SG61410017	SG61410018	SG61410019
MHz	ENR (dB) ± 0.05 dB					
3535	14.87	14.83	15.02	14.85	15.15	14.88
3545	14.87	14.83	15.02	14.85	15.16	14.88
3555	14.87	14.83	15.02	14.85	15.15	14.88
3565	14.86	14.82	15.01	14.85	15.14	14.87
3575	14.85	14.81	15.00	14.84	15.13	14.87
3585	14.84	14.81	15.00	14.84	15.13	14.86
3595	14.84	14.81	15.00	14.82	15.12	14.85
3605	14.83	14.80	14.99	14.82	15.12	14.85
3615	14.82	14.79	14.98	14.81	15.12	14.84
3625	14.83	14.79	14.98	14.82	15.11	14.84
3635	14.83	14.80	14.99	14.81	15.11	14.84
3645	14.84	14.80	14.99	14.83	15.12	14.85
3655	14.84	14.80	14.99	14.82	15.12	14.85
3665	14.85	14.81	15.00	14.84	15.13	14.86
3675	14.85	14.81	15.00	14.83	15.13	14.85
3685	14.84	14.80	14.99	14.83	15.13	14.85
3695	14.83	14.79	14.97	14.81	15.11	14.83
3705	14.82	14.78	14.96	14.81	15.10	14.82

6. SEA Sensor Results

As noted in the section 1.3, the [NASCTN CBRS SEA](#) program deployed a series of sensor nodes in three broad geographical locations in contiguous United States ([CONUS](#)). The locations were chosen to be aligned with [DPAs](#) on the east coast and west coast. In addition, Boulder Colorado was chosen as a prototyping and control location, for detailed location information see [15, Table 1.]. Each sensor node was equipped with a switch and a radiometer characterized noise diode (see section 5) located in the preselector, see [16, Table 16.]. Every sensor was set to calibrate on an hourly basis by switching to the path with the radiometer characterized noise diode and performing the standard Y-factor measurement outlined in section 1. The manufacturer specification for temperature variations of the noise diode [ENR](#) is $0.01 \frac{dB}{^{\circ}C}$, and variations in the gain greater than this are assumed to be real changes in system gain and not a change in power emitted from the noise diode. In addition, at the installation site, the antenna was removed and a calibration probe was used to preform a Y-factor measurement that included any extra cabling from the antenna to the preselector of the sensor, see Section 6.1. This value translates the reference plane of the noise diode to the attachment site of the antenna and is included in the values reported, approximating the results of Section 6.1. The hourly on-sensor calibration leads to approximately 8,760 Y-factor measurements per 10 MHz channel per sensor per year. These Y-factor measurements are further separated by the state of the sensor. For the sensor nodes in this report, the states are described by the attenuation and reference level of the spectrum analyzer found in the [SPU](#), see [16, Section 2.1.7] for specifics. Additionally, the antenna the system was using has been used to logically divide data for the three sensors with dual antenna configurations. All calibration data is available in the data repository detailed by [15]. Specifically, the calibration information is saved in the metadata [15, Section 3.3] of each signal metadata format ([SigMF](#)) that are found as zip archives in the Raw Data folder [15, Section 2.2.1]. For reference, the calibration field is contained in captures located with the key `ntia-sensor:sensor_calibration`, or `metadata/captures/capture x/ntia-sensor:sensor_calibration` where x is a number between 0 and 17. A summary of all calibration results from sensors with a spectrum analyzer reference level of -19 dBm and internal attenuation of 6 dB are shown in Appendix C, refer to [16, Section 2.1.7] for specifics. It should be noted that not all the calibrations are done with the same reference level and attenuation, therefore for a full accounting of every calibration it is necessary to access the full data set [29].

6.1. Dual Calibration Using an Onboard Noise Diode and The Calibration Probe

This section describes the parameters of importance to a dual calibration. For simplicity, these parameters are grouped into the form of equation 5, and in practice we calculate the gain from the onboard noise diode and calibration probe separately, applying a constant difference to the gain values to account for the path difference. In Figure 25, a simplified block diagram of the sensor is shown. The antenna port is labeled as reference plane

1, the output of the onboard noise diode is labeled as reference plane 2. To accurately measure power arriving at the calibration plane labeled C from the plane marked 1, we need to calculate the transducer gain ratio $G_T^{2 \rightarrow C} / G_T^{1 \rightarrow C}$. Both the probe and the noise diodes are calibrated with noise temperatures, T_p and T_{ND} respectively. Note that the noise diode (ND) has two known states; hot and cold (ambient). The power measured by the signal analyzer is given by the equations [30]

$$P_p = k_B B \left[G_T^{1 \rightarrow C} T_p + M_{C-p} T_s + (M_{C-p} - G_T^{1 \rightarrow C}) T_a \right] G \quad (14a)$$

$$P_{NDh} = k_B B \left[G_T^{2 \rightarrow C} T_{NDh} + M_{C-NDh} T_s + (M_{C-NDh} - G_T^{2 \rightarrow C}) T_a \right] G \quad (14b)$$

$$P_{NDc} = k_B B M_{C-NDc} (T_s + T_a) G \quad (14c)$$

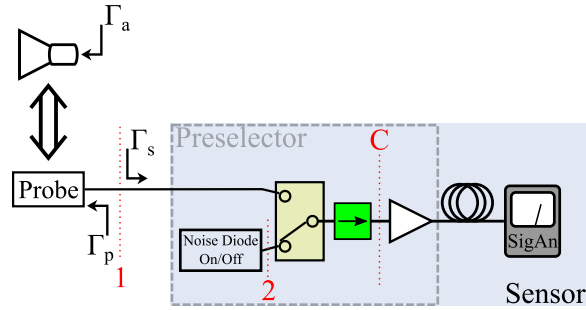


Fig. 25. The calibration block diagram, identifying the reference plane of antenna input (1), the reference plane of the calibration diode (2) and the calibration plane (C). The probe is inserted into the antenna port with reflection Γ_p , and then is replaced by the antenna in the field with reflection Γ_a , the reflection of the sensor input is Γ_s

Here k_B is the Boltzmann constant. B is the bandwidth set by the signal analyzer or the post-processing DSP (usually the frequency bin size). T_s is the equivalent noise temperature of the sensor at plane 'C'. G is the deliverable gain from 'C' to the signal analyzer. The inclusion of the isolator (the green component in Figure 25) allows us to approximate the mismatch factors at plane 'C' for dissimilar input terminations $M_{C-p} \approx M_{C-NDh} \approx M_{C-NDc} \equiv M$. As a result, (14) can be approximated as

$$P_p = k_B B G_T^{1 \rightarrow C} (T_p - T_a) G + k_B B M (T_s + T_a) G \quad (15a)$$

$$P_{NDh} = k_B B G_T^{2 \rightarrow C} (T_{NDh} - T_a) G + k_B B M (T_s + T_a) G \quad (15b)$$

$$P_{NDc} = k_B B M (T_s + T_a) G \quad (15c)$$

The three linear equations in (15) provide a solution of the transducer gain ratio $\frac{G_T^{2 \rightarrow C}}{G_T^{1 \rightarrow C}}$. We label it as β .

$$\beta = \frac{(P_{NDh} - P_{NDc})(T_p - T_a)}{(P_p - P_{NDc})(T_{NDh} - T_a)} \quad (16)$$

In the field, the calibration probe is replaced with the antenna to monitor CBRS emissions. The parameter of interest is T_x , the field radiation temperature measured at the antenna port. When the antenna is connected to the sensor, the transducer gain from '1' to 'C' is $G_T^{1a \rightarrow C}$. Similar to (15a), the power measured at signal analyzer in the field is

$$P_x = k_B B G_T^{1a \rightarrow C} (T_x - T_a) G + k_B B M (T_s + T_a) G \quad (17)$$

In combination with the measurements of the noise diode in the field, we obtain:

$$T_x = \frac{G_T^{2 \rightarrow C}}{G_T^{1a \rightarrow C}} \frac{P_x - P_{NDc}}{P_{NDh} - P_{NDc}} (T_{NDh} - T_a) + T_a \quad (18a)$$

$$= \beta \frac{G_T^{1p \rightarrow C}}{G_T^{1a \rightarrow C}} \frac{P_x - P_{NDc}}{P_{NDh} - P_{NDc}} (T_{NDh} - T_a) + T_a \quad (18b)$$

Note that the T_a 's in (16) and the T_a 's in (20) may be drastically different. T_a 's in (16) are measured in the laboratory. β is determined in the lab experiments prior to the sensor deployment. T_a 's in (20) are frequently updated by temperature monitoring devices in the field. In addition, P_{NDh} and P_{NDc} in (20) are calibrations performed in the field.

We determine the ratio $\frac{G_T^{1p \rightarrow C}}{G_T^{1a \rightarrow C}}$. From network analysis, the transducer gain relates to the reflection coefficients of the input termination and the receiver. We can express the ratio by [31]

$$\frac{G_T^{1p \rightarrow C}}{G_T^{1a \rightarrow C}} = \frac{(1 - |\Gamma_p|^2) / |1 - \Gamma_p \Gamma_s|^2}{(1 - |\Gamma_a|^2) / |1 - \Gamma_a \Gamma_s|^2} \quad (19)$$

In virtue of (19), we represent (20) by

$$T_x = \beta \frac{(1 - |\Gamma_p|^2) |1 - \Gamma_a \Gamma_s|^2}{(1 - |\Gamma_a|^2) |1 - \Gamma_p \Gamma_s|^2} \frac{(P_x - P_{NDc})}{(P_{NDh} - P_{NDc})} (T_{NDh} - T_a) + T_a \quad (20)$$

To estimate the variation due to mismatch between the antenna when compared with the variable noise source the term

$$\frac{(1 - |\Gamma_p|^2) |1 - \Gamma_a \Gamma_s|^2}{(1 - |\Gamma_a|^2) |1 - \Gamma_p \Gamma_s|^2}$$

is calculated using multiple measurements of the antenna in different laboratory locations. These locations were chosen at random to reflect an potential installation in an uncontrolled environment. In Figure 26, the result of this calculation can be seen. This ratio varies from 0.98 to 1.07. This term has a frequency dependence that varies between sensor deployments and configurations. To account for this, we introduce a scalar error term $\Delta_\Gamma = 0.2 \text{ dB}$. Although the algebra can obscure it, the power reading on the sensor is

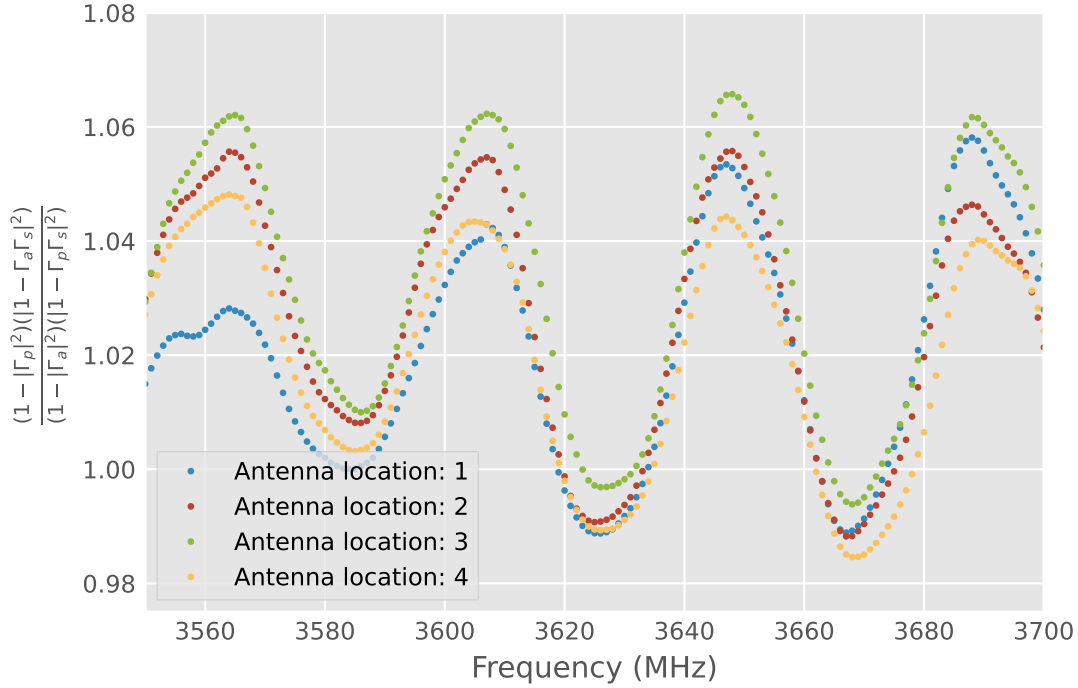


Fig. 26. Comparison of Antenna Mismatches. The calibrated noise temperature reading requires adjusting for the antenna γ_a and that of the probe γ_p , every installation potential has a slightly different antenna mismatch. To assess the impact of this difference an antenna was measured in four locations.

linearly related to the impinging signal temperature. To facilitate understanding, we may combine and regroup Equation 20, as a simple linear equation 21. We begin by defining the term α equal to the $\alpha = \beta \frac{(1-|\Gamma_p|^2)(1-|\Gamma_a\Gamma_s|^2)}{(1-|\Gamma_a|^2)(1-|\Gamma_p\Gamma_s|^2)} \frac{(T_{NDh}-T_a)}{(P_{NDh}-P_{NDc})}$ after algebraic manipulation we are left with

$$P_x = \frac{T_x}{\beta\alpha} + P_{NDc} - \frac{T_a}{\beta\alpha} \quad (21)$$

or more succinctly

$$P_x = AT_x + B \quad (22)$$

Where A is related to the effective gain and B is a gain scaled constant that represents the ambient noise temperature and the noise temperature of the sensor. The calibration procedure quantifies A and B at a given moment in time at a specific temperature. It is assumed that both of these coefficients can change with time (due to ageing and finite system stability) and with temperature (changes to system gain). The noise diode integrated into the sensor allows us to quantify these changes. The more common linear form is

$$P_x = G_{eff}Bk_B T_x + G_{eff}Bk_B T_{sensor} \quad (23)$$

Where P_x is the reading of the power sensor, G_{eff} is the effective gain in linear units, B is the bandwidth, k_B is Boltzmann's constant, T_x is the signal noise temperature and T_{sensor} is the noise temperature added by the sensor. This is the same as Equation 5 with a signal term instead of thermal noise.

6.2. Overview of Field Measurements

Each Y-factor measurement of a sensor node returns the gain and noise figure at that moment in time. The gain and noise figure are dominated by the temperature response of the amplifiers located in the preselector and SPU. There are two primary configurations of sensor nodes deployed. One configuration has a preselector located outside in an National Electrical Manufacturer's Association (NEMA) rated box and the SPU in a climate controlled location. In the second configuration both the preselector and SPU are located outside without building climate controls, however the SPU has a thermo electric heater/cooler (TEC) that prevents temperature over-ranging. In the case that only the pre-selector is allowed to vary in temperature we expect the gain to change exponentially with temperature due to the semiconductor components or in logarithmic units to follow the equation 24.

$$G(T) = \alpha T + G_0 + \varepsilon_r \quad (24)$$

Where α is the change in log gain versus temperature, G_0 is the gain at 0° C, and ε_r is a normally distributed variable with zero mean. Similarly, the noise figure is modeled by

$$NF(T) = \alpha^{NF} T + NF_0 + \varepsilon_r^{NF} \quad (25)$$

Where α^{NF} is the change in noise figure versus temperature, NF_0 is the noise figure at 0° C, and ε_r^{NF} is a normally distributed variable with zero mean.

6.3. Variation of Calibration from Input Port

The calibration probe detailed in section 3, whose programmable noise temperature values were determined by a radiometer measurement outlined in section 5, was connected to a production sensor in a laboratory environment. Multiple data points were taken at each calibration probe attenuation of 20 dB, 30 dB, 40 dB, and 110 dB to give the linear trend as well as the matched "0 dB ENR". To confirm each of these measurements was free of other signals, the process suggested in 2.2 using AMPR, was employed, in Figure 27 the results for each frequency channel and all measurements are shown. The above process was then carried out for all permutations of these data points to indicate the variability in the verification bench testing and acceptance (VBA) calibration process. In Fig. 28, the results of fitting each permutation are shown.

With nearly 45,000 iterations, the total standard deviation of offset gains around the mean value was 0.025 dB, or an estimated 95% confidence interval of 0.05 dB. The total standard deviation of noise figures around the mean value was 0.034 dB, or an estimated 95% confidence interval of 0.07 dB. These values represent the smallest random uncertainty due

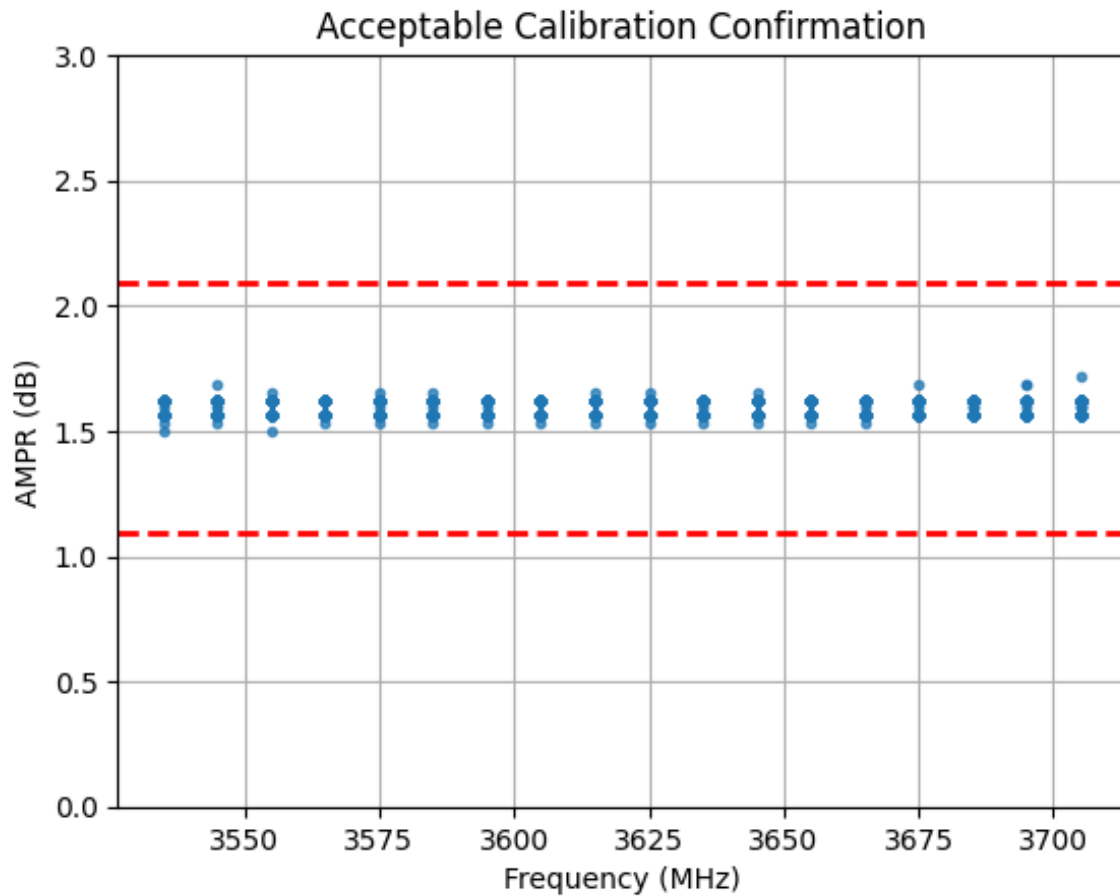


Fig. 27. Confirmation of Calibration Probe Measurements using AMPR

to multiple laboratory based Y-factor calibrations one could expect in these type of sensors. It should be noted that for the channels centered at 3695 MHz and 3705 MHz, the filter in the preselector causes a large frequency dependent loss, violating the criteria for accurate noise calibrations. These channels were included in the variational analysis, but have much larger uncertainties than the variational analysis would indicate. In Figure 29, the linear fits for each channel are shown. For a detailed accounting of the filter affect on gain and noise figure as a function frequency consult [16, Figure 5].

6.4. Pre-production Field Testing

The [NASCTN CBRs SEA](#) project was phased in its approach to sensor deployment. Initial sites were chosen for testing based on [RF](#) field of regard, location and expediency. The first sensors deployed were located at GMM, HU and NIT [15, Table 1.]. These early deployments have a slightly different configuration from the production sensors that are addressed in [14] with updates being noted in [13]. The most important difference for our

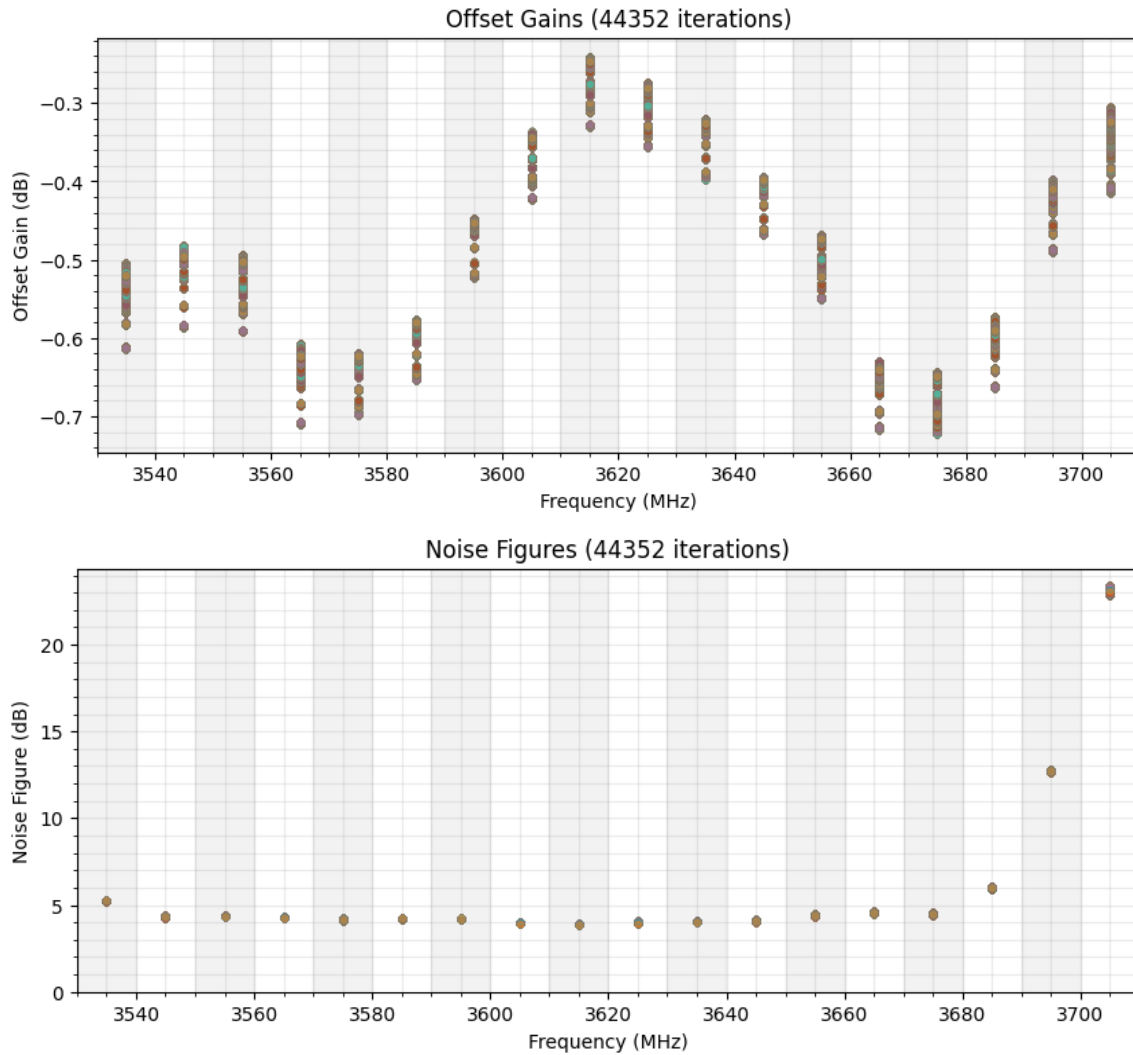


Fig. 28. Variability of gain and noise figures for 45352 iterations of four point fits

purposes is that the frequency range is 150 MHz, and a different filter in the preselector is utilized. The inclusion of this pre-production sensor calibration responses are meant as a guide to demonstrate methodology, and do not represent the final uncertainty estimates. The temperature variation of the gain for a coastal region (east coast of CONUS) can be assessed from field data from the sensor deployed at Hampton University. Data from the time period 10/06/2022 - 10/26/2022 is shown in Figure 30 for a single 10 MHz channel, identified as channel 7 with the center frequency of 3615 MHz. Each gain and noise figure value is accessed at every measurement [15], but the calibration procedure happens on a longer time interval, causing values to be repeated. It can be seen that the gain follows equation 24, a fit using this model displays a slope of $0.06 \frac{dB}{^{\circ}C}$. When the linear trend of gain values for temperature are subtracted from the data, an estimate of the residual gain

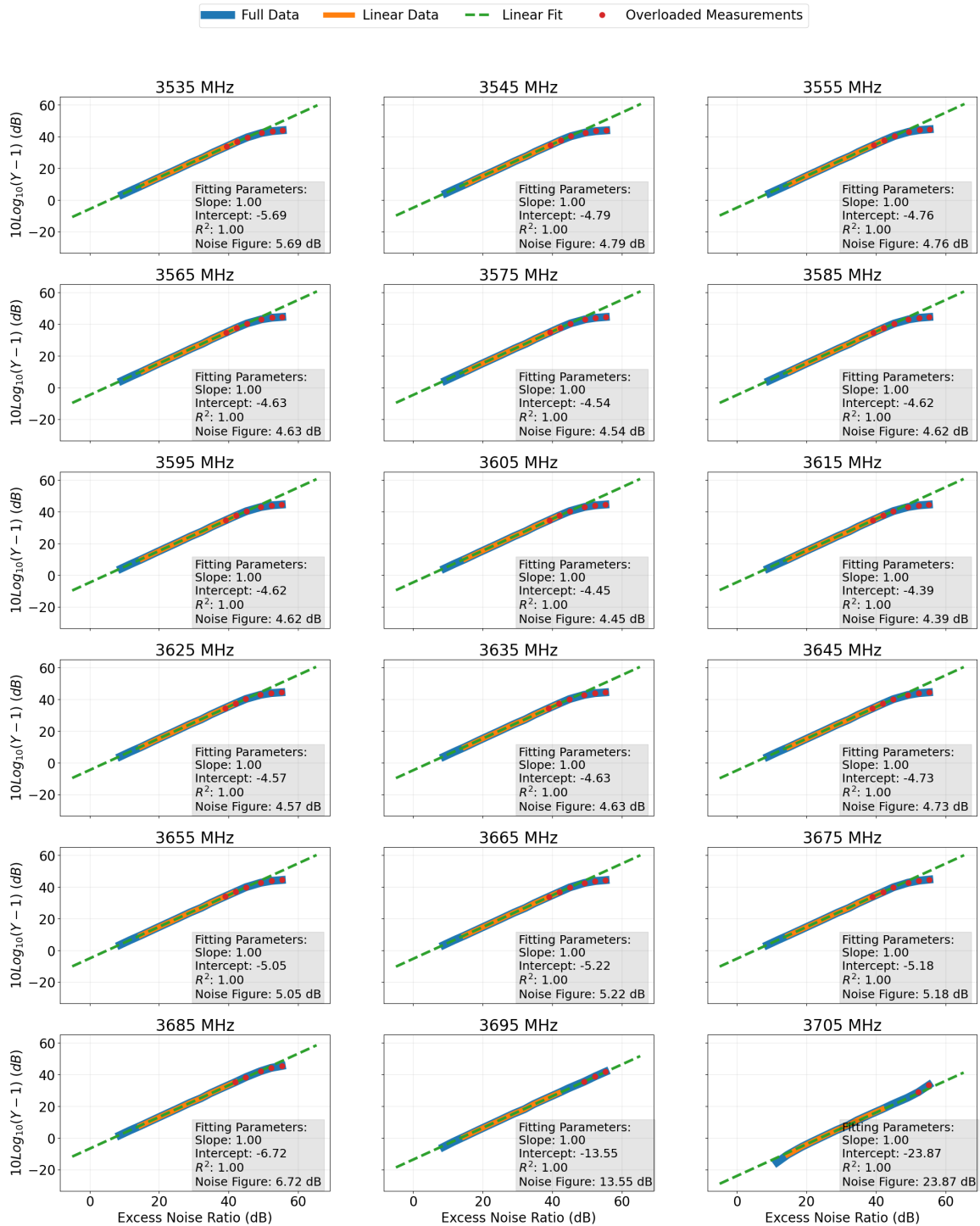


Fig. 29. Programmable Noise Source Calibration with Linear Fit for Each Channel

ϵ_r distribution remains. The residual gain is shown in Figure 31. To better represent the to-

tal of gain variations for the sensor that are not corrected for, this process was repeated for all 15, 10 MHz channels, and aggregated. By aggregating all residual distributions across all channels a maximum channel random uncertainty is estimated. In Figure 32, the results of this process can be seen. The residuals for 597540 measurements shows a standard deviation of 0.15 dB, which leads to an estimate of the expanded uncertainty due to variability of the sensor to measure a known, fixed power and any time and temperature variations unaccounted for as ± 0.30 dB.

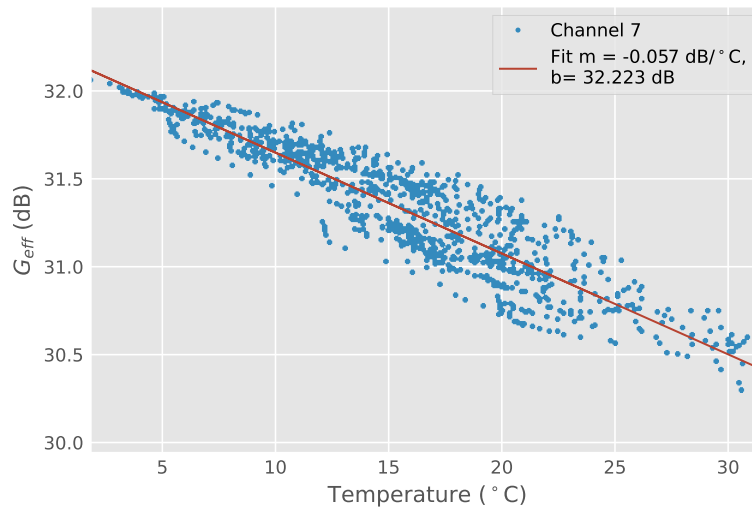


Fig. 30. Gain (logarithmic units) versus temperature of channel 7 (center $f = 3615$ MHz) shows a clear linear trend, this is repeated for all channels

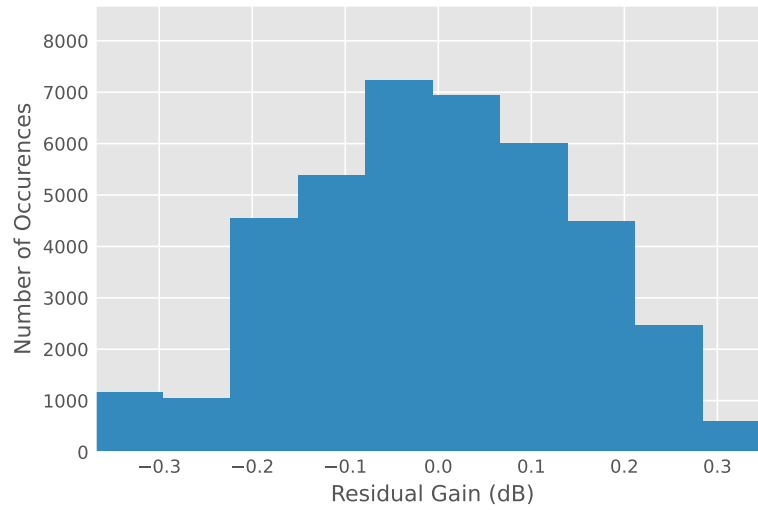


Fig. 31. Residual Gain After Detrending of Channel 7, Hampton University Sensor.

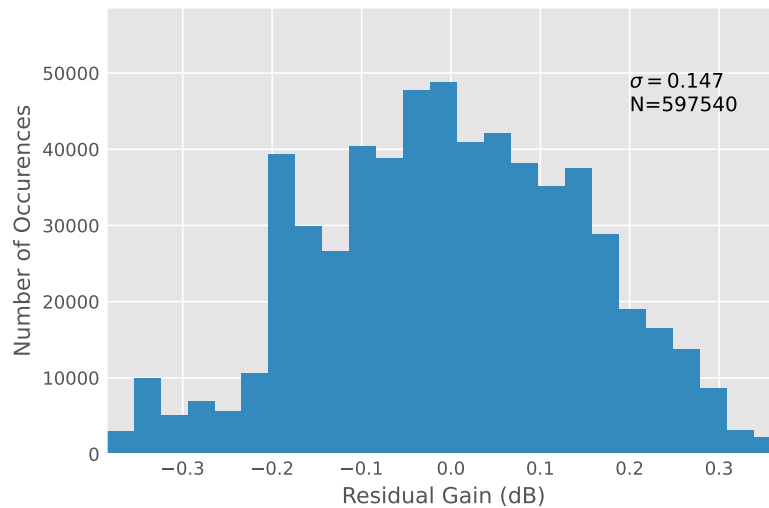


Fig. 32. Residual Gain after Detrending Temperature Dependence of All Channels, Hampton University Sensor. Histogram of the residual gain after detrending temperature dependence.

6.5. Limited Time Fielded Sensor Calibration Results

After the pre-production period discussed in section 6.4, sensors at the remaining sites were deployed, see Figure 2 for location and naming. These sensors represent a statistical sample of similar, but not identical, realizations of the general sensor design discussed in

section 1.2. The official operational period for production sensors began mid-2024 and ended March, 2026. Results of periodic calibrations over the an extended period are covered in section 6.6. In this section, we analyze a limited amount of data from the time period between October 6th and October 12th 2025 assuming equation 24 is valid. For every sensor, antenna and each 10 MHz frequency span from 3530 MHz to 3710 MHz, the gain versus temperature was least squares fit to a line and the residuals found by taking the difference between the linear fit and the data. In Figure 33, an example of the fit process is shown for HU at 3635 MHz. As stated in 6.2 the residuals are approximated as a normal distribution with zero mean and a standard deviation of σ_{ϵ_r} . As with the pre-production analysis in section 6.4, the slope is the change of gain with temperature in logarithmic units and the intercept is the gain at zero degrees C. We expect similar results to the pre-production, however since this analysis encompasses all sensors we can use it to develop an understanding of the variation between different realizations and installation locations of sensors. In Figure 34, the results for all sensor-antenna combinations and for all monitored frequencies are shown. The large reduction of G_0 for frequencies greater than 3685 MHz is due to the filter in the preselector. Higher than this frequency, the assumptions made in equation 24 are affected by the presence of the filter, and the temperature slope of the gain and the standard deviation of the residual distribution all change. For a reduced frequency range from 3555 to 3645 MHz, the frequency dependent difference in α and σ_{ϵ_r} is very small compared to the value measured.

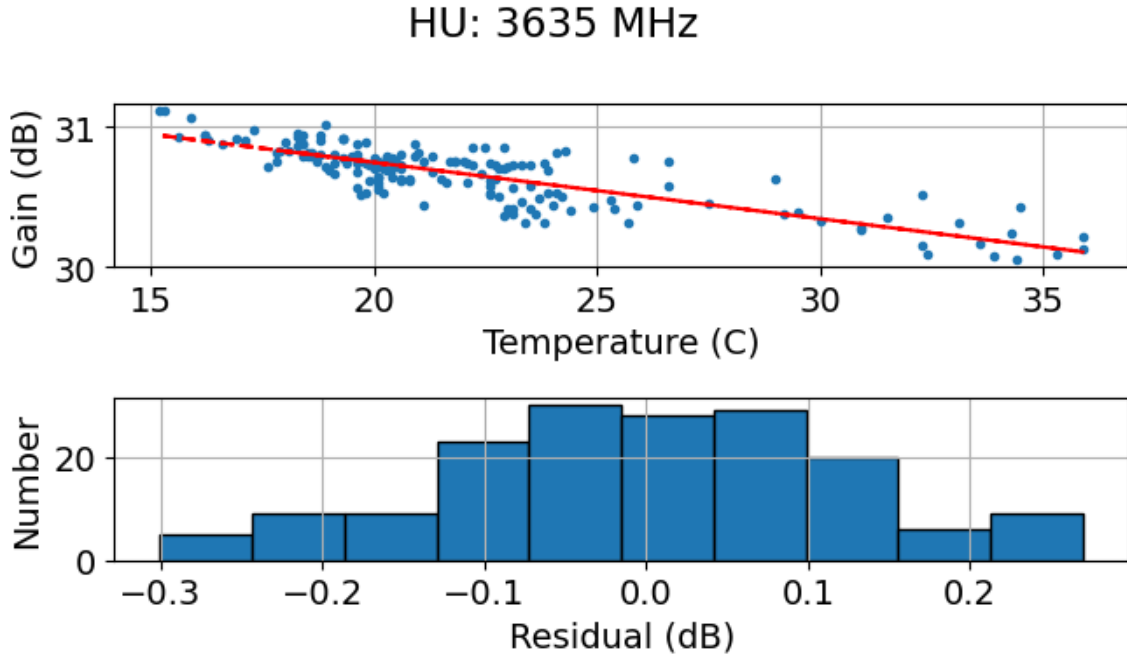


Fig. 33. Example of Fit for Reduced Time Linear Fit of Gain

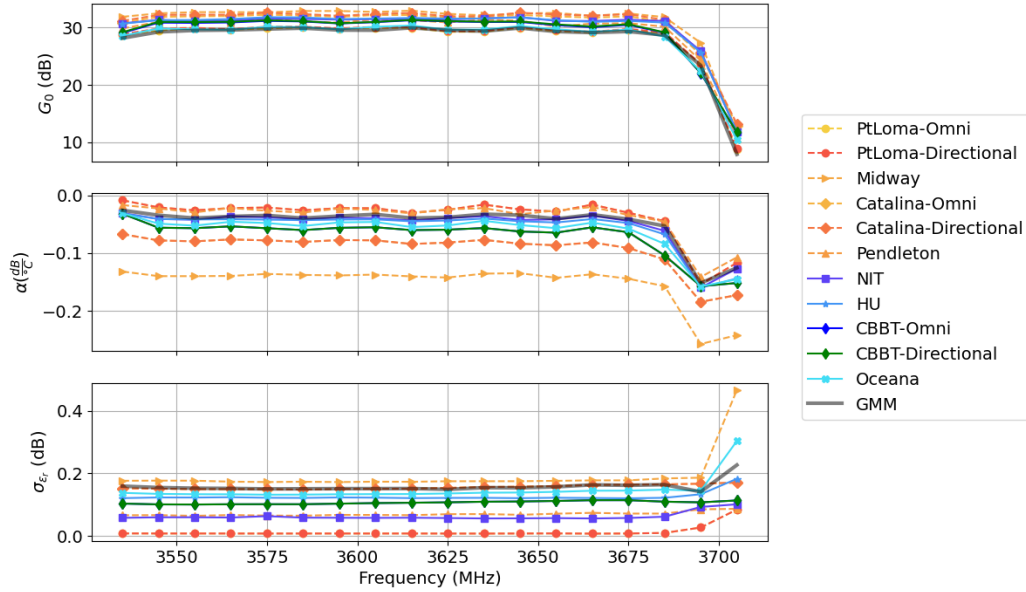


Fig. 34. Summary of Fit Parameters for Reduced Time Linear Fit of Gain

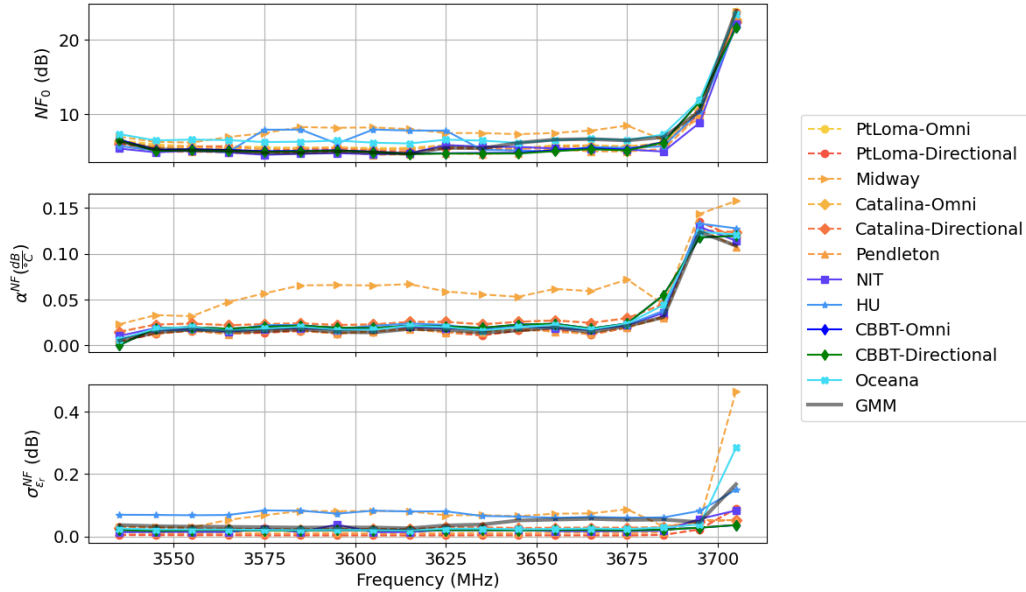


Fig. 35. Summary of Fit Parameters for Reduced Time Linear Fit of Noise Figure

6.6. Long Term Production Sensor Calibration Results

6.6.1. All Field Calibration Results

As noted above, every 10 MHz channel for every sensor generates 8,760 calibration measurements per year. This has generated in excess of 2 million measurements of gain and

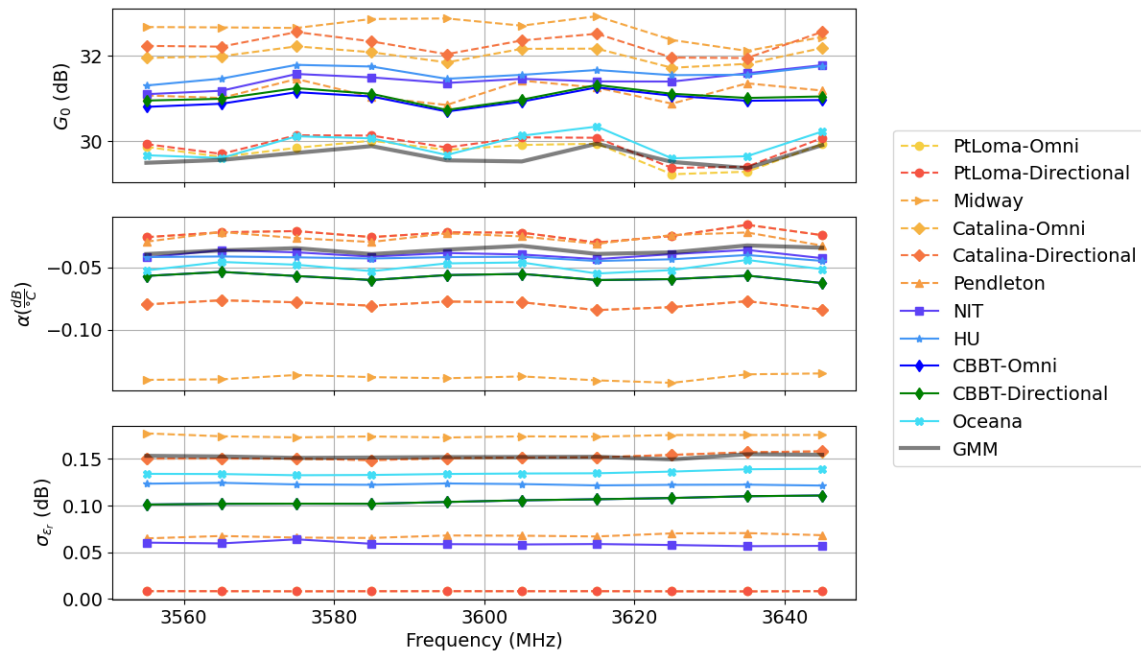


Fig. 36. Summary of Fit Parameters for Reduced Time Linear Fit of Gain for 3550-3650 MHz

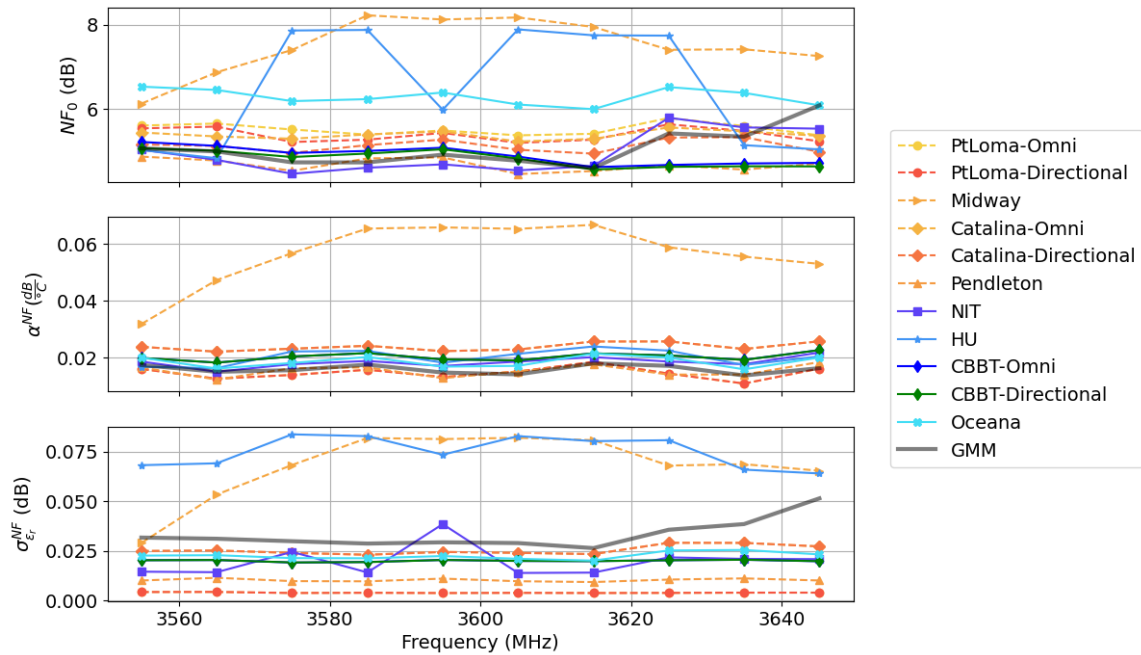


Fig. 37. Summary of Fit Parameters for Reduced Time Linear Fit of Noise Figure for 3550-3650 MHz

noise figure during the [NASCTN CBRS SEA](#) program. This has revealed that temperature drives the primary change in gain and noise figure in time. The magnitude of this change varies from sensor to sensor and we hypothesize that it is a combination of intrinsic temperature variation coefficient in the amplifiers present and the external temperature sensitivity specific to installation site. For example, in Figure 37, it can be observed that α is approximately $-0.05 \frac{dB}{^\circ C}$ for most East Coast sensors. However, we observe extremes for the sensors installed on the USS Midway and at PtLoma. Although the preselectors in both sensors are similar, the [SPUs](#) are installed in different locations. The Midway installation is in a shaded but an environmentally uncontrolled setting without a [TEC](#). In contrast, the Pt. Loma installation is in a temperature controlled space.

6.6.2. Long Term Temperature Trends

When considering the long term range of calibrations from the [NASCTN CBRS SEA](#) program, an interesting feature arises. Unlike the limited time calibration results in section 6.5, some of the extended range calibrations show a bilinear trend (see Figure 38). This trend is a result of the preselector and [SPU](#) being at different temperatures. In particular, the design covered in [16], regulates the temperature of the [SPU](#) using a [TEC](#). The [TEC](#) is used to prevent exceeding temperature ranges for components in the [SPU](#). For ambient temperatures that do not cause the internal [SPU](#) temperature to exceed these criteria, the [TEC](#) remains inactive. For a synopsis of the internal temperatures of the preselector and [SPU](#) see [16, Figures 22 and 32]. To be more precise, a model that accounts for both the temperature of the preselector and [SPU](#) is required. Our model is a bilinear extension of equations 24 and 25. Equations 26 and 27 represent this model.

$$G(T_{preselector}, T_{SPU}) = \alpha T_{preselector} + \beta T_{SPU} + G_0 + \epsilon_r \quad (26)$$

Where α is the change in log gain versus preselector temperature, β is the change in log gain versus [SPU](#) temperature, G_0 is the gain at $0^\circ C$, and ϵ_r is a normally distributed variable with zero mean.

$$NF(T_{preselector}, T_{SPU}) = \alpha^{NF} T_{preselector} + \beta^{NF} T_{SPU} + NF_0 + \epsilon_r^{NF} \quad (27)$$

Where α^{NF} is the change in noise figure versus preselector temperature, β^{NF} is the change in noise figure versus [SPU](#) temperature, NF_0 is the noise figure at $0^\circ C$, and ϵ_r^{NF} is a normally distributed variable with zero mean.

To create a long time perspective the calibration results were organized by schedule id, or the set of instructions that dictate the capture parameters. The schedule id is found in the `metadata/global/ntia-scos:schedule/id` field. In Figures 39 and 40, the fitting results for most sensors for all frequencies are displayed, see Table 11 and Table 12 for all residual standard deviations. In Figures 41 and 42, the results for the frequency range 3550-3650 MHz are displayed. The residual standard deviations from the linear and bi-linear

fits are compared for selected sensors in Figures 43 and 44. The resulting residual distributions are in good agreement for all but the Pt. Loma sensor. The standard deviation of the residuals for the Pt. Loma sensor is approximately 0.01 dB for the short-term linear fit and 0.15 dB for the extended-time bilinear fit for frequencies 3550-3650 MHz. In this case, the sensor does not experience enough temperature variation to accurately capture the true field deployed variance. This illustrates that in situations with small temperature excursions the residual distribution can have a very small standard deviation (<0.03 dB). We therefore conclude a short-term linear fit is an accurate estimate of the uncertainty and temperature behavior as long as excursions of the temperature are sufficient. The magnitude of this temperature excursion is estimated to be >12 °C with multiple instances of rates of change of $>1 \frac{^{\circ}\text{C}}{\text{hr}}$. It is notable that the temperature dependent coefficients in both type of fits is typically between $0.04 \frac{\text{dB}}{^{\circ}\text{C}}$ and $0.07 \frac{\text{dB}}{^{\circ}\text{C}}$, indicating that calibrations preformed with temperature difference of 1°C would result in a gain difference of 0.04 dB to 0.07 dB, roughly the uncertainty of the radiometer characterized noise diodes.

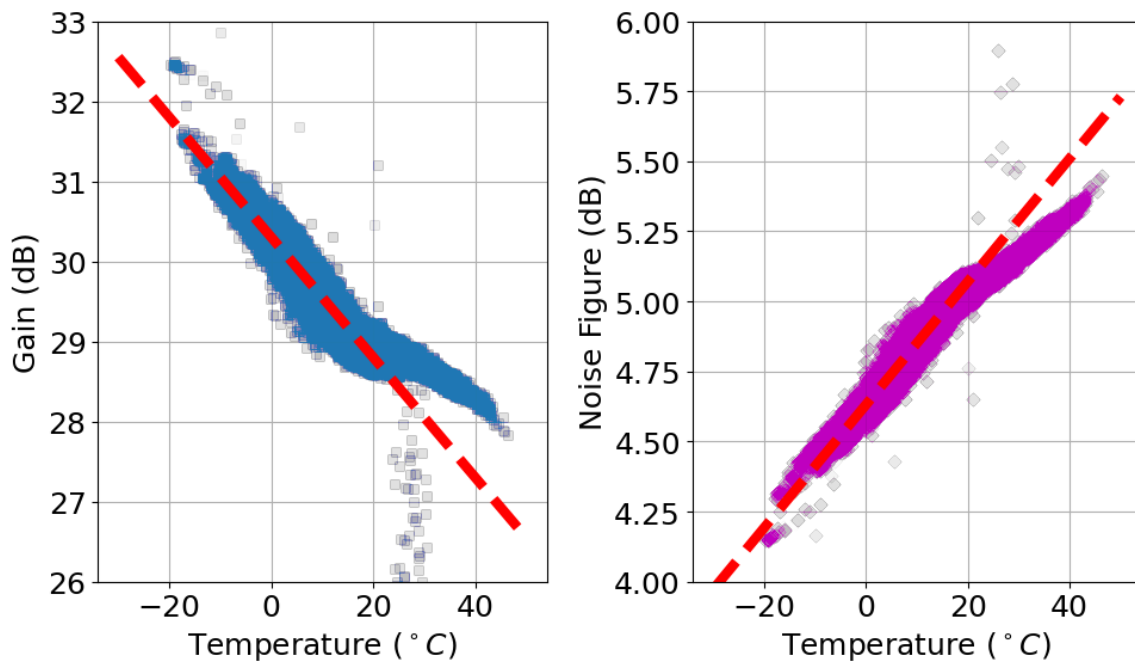


Fig. 38. Gain Values for GMM Versus Preselector Temperature. Red dashed lines are guides for the eye.

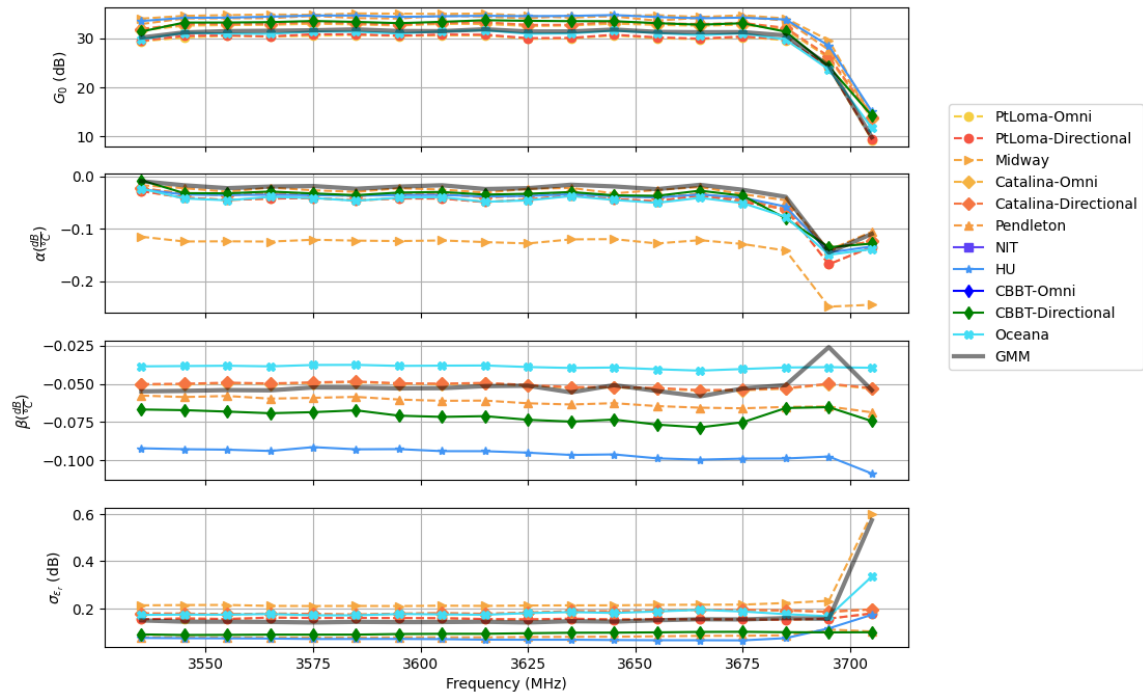


Fig. 39. Summary of Fit Parameters for Extended Time Bi-Linear Fit of Gain

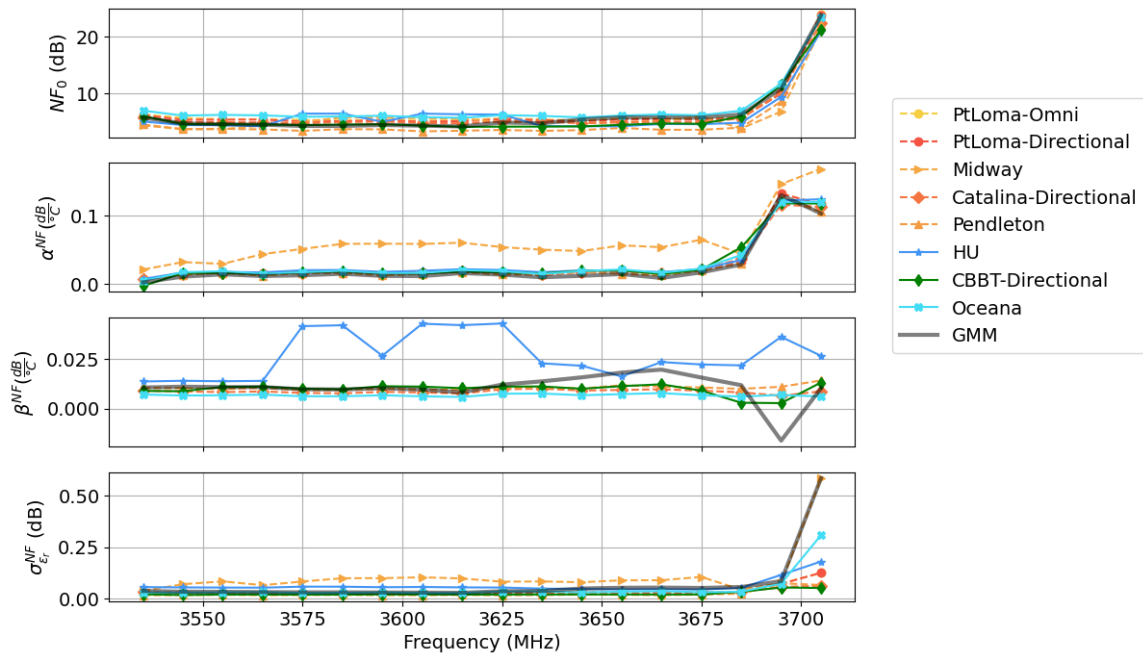


Fig. 40. Summary of Fit Parameters for Extended Time Bi-Linear Fit of Noise Figure

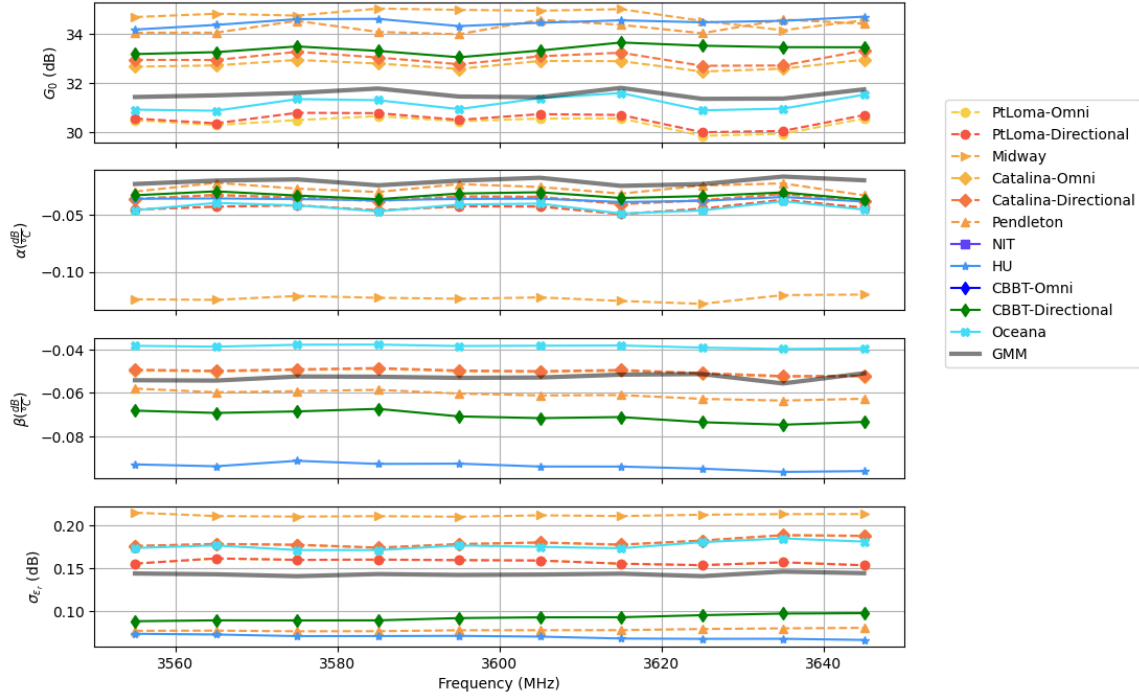


Fig. 41. Summary of Fit Parameters for Extended Time Bi-Linear Fit of Gain for 3550-3650 MHz

6.7. Uncertainty of Power Measurements

The point of calibrating any sensor system is to be confident of the returned value from that sensor system. This returned value cannot be perfect, that is, it always has some uncertainty. For example, the minimum uncertainty that is expected from a Y-factor measurement, is never smaller than the uncertainty of the calibration artifact (noise diode). This is approximately 0.2 dB for COTS devices characterized by the manufacturer, and 0.05 dB for the same device characterized by the NIST radiometer, see section 5. The total uncertainty in the logarithmic gain is the uncertainty in the logarithmic power for the NASCTN CBRS SEA sensors. This uncertainty is dependent on the system and the frequency of the band being monitored. The uncertainty in noise figure represents the minimum trusted power measurement of the sensors. The noise figure analysis is included in Appendix A for completeness. The major sources of gain uncertainty of this type of sensor are:

- The uncertainty of the calibration standard. The radiometer standard uncertainty was **0.05 dB**. We assume this to be a (log) uniformly distributed uncertainty.
- The band flatness change in gain or noise figure across the calibration bandwidth. For gain in the frequency range of 3550-3650 MHz this value is typically between **0.1 dB** and **0.6 dB**. The complete frequency dependent table is available in Appendix A Table 13. This table uses the maximum of the difference of the median values for

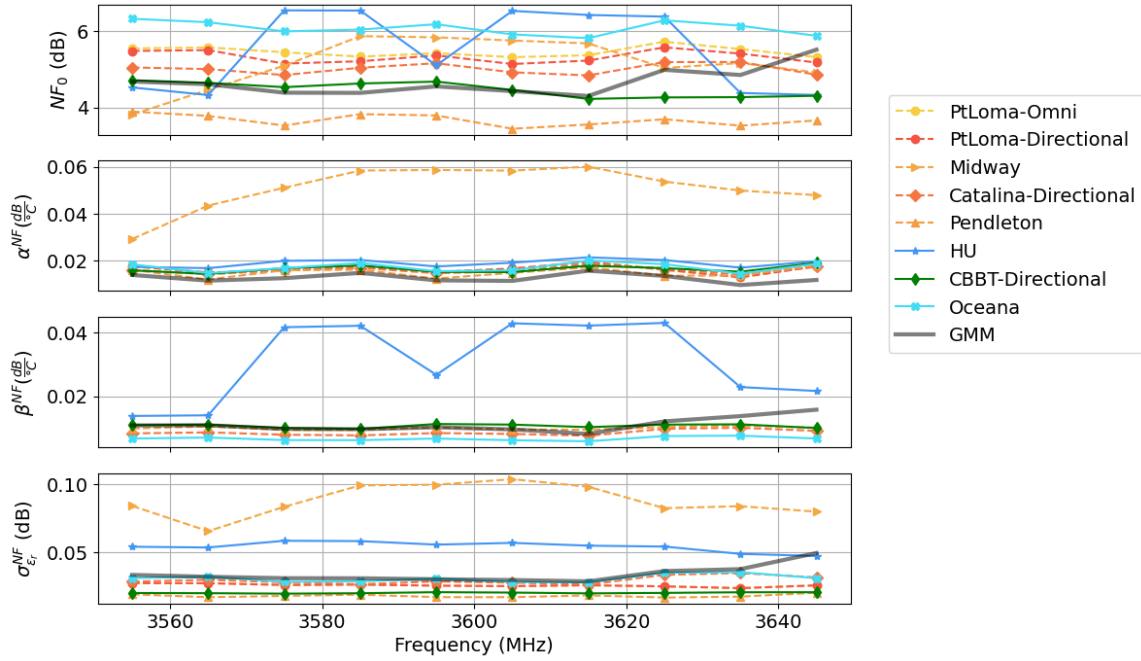


Fig. 42. Summary of Fit Parameters for Extended Time Bi-Linear Fit of Noise Figure for 3550-3650 MHz

-19 dBm reference value and a 6 dB Attenuation in adjacent bands. We assume this to be a (log) uniformly distributed variable. This is frequently the largest source of uncertainty.

- Changes in temperature between calibrations. We use the error function in the temperature dependent fit to estimate this. For gain in the frequency range of 3550-3650 MHz this value is typically between **0.1 dB** and **0.2 dB**. The complete frequency dependent table is available in Appendix A Table 11. We assume this to be a (log) normal distribution.
- The uncertainty of a Y-factor type calibration. This includes the random variation in measurements estimated in Section 6.3 **0.05 dB** and any unaccounted network mismatch estimated in Section 6.1 as **0.20 dB**. We assume the random variation to be a (log) normal distribution and the network mismatch to be a (log) uniform variable.

To estimate the total uncertainty, the values above were used in a Monte Carlo simulation with 10,000 trials. The total uncertainty for gain for each sensor in the 3550-3650 MHz frequency range is shown in Figure 45. Similarly, the total uncertainty for noise figure for each sensor in the in the 3550-3650 MHz frequency range is shown in Figure 46. The uncertainty for all sensors and frequencies for gain is reported in Table 7, and for noise figure

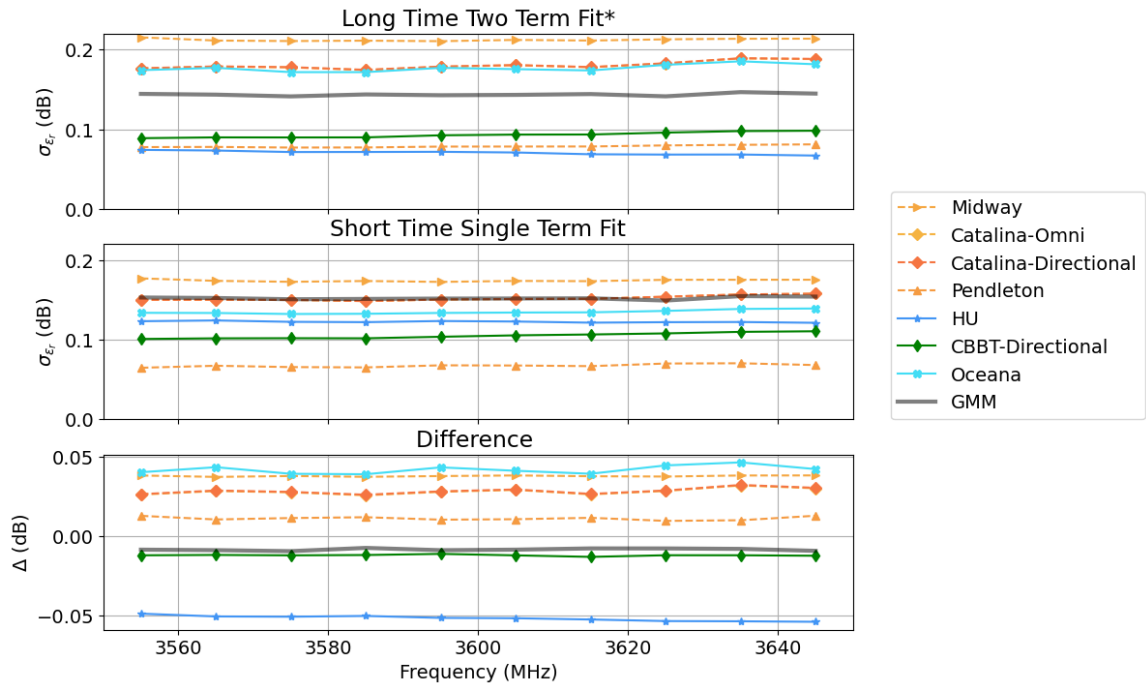


Fig. 43. Comparison of Standard Deviation of Residuals From Linear and Bi-Linear Fits for Gain for 3550-3650 MHz

is reported in Table 8, in Appendix A. The total uncertainty in the 3550-3650 MHz range is ≈ 0.5 dB for gain and ≈ 0.35 dB for noise figure in most sensor-frequency combinations.

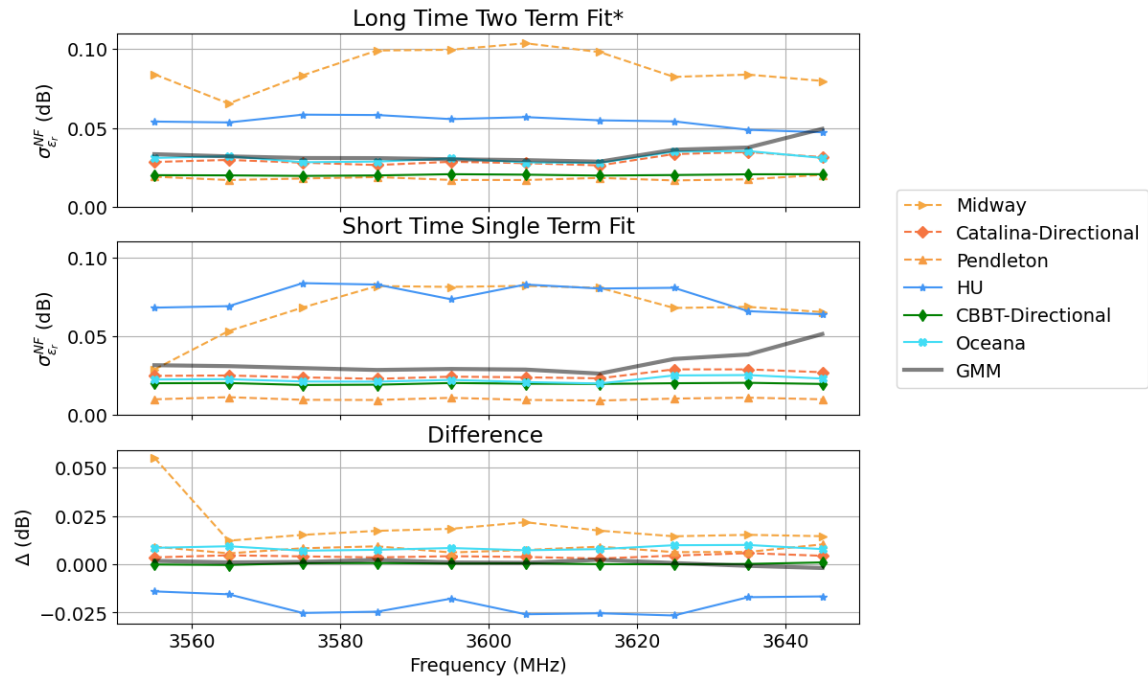


Fig. 44. Comparison of Standard Deviation of Residuals From Linear and Bi-Linear Fits for Noise Figure for 3550-3650 MHz

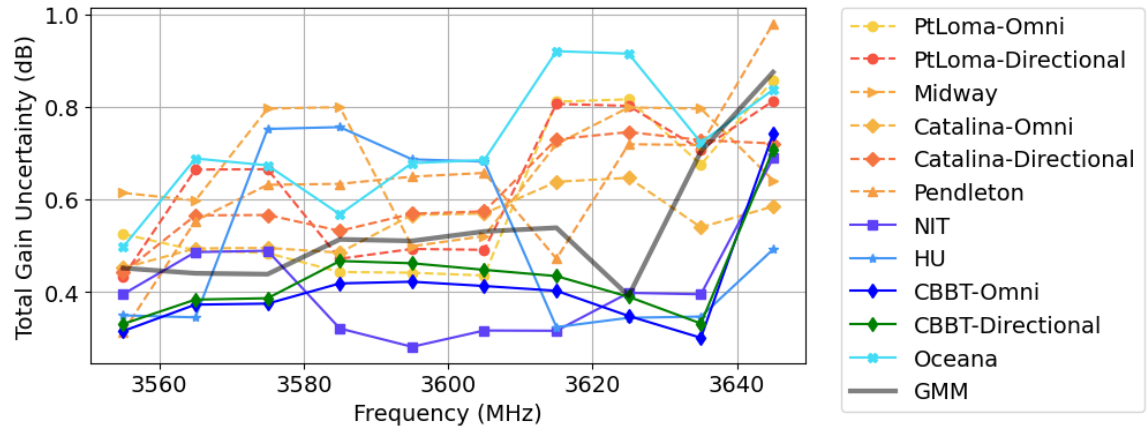


Fig. 45. Total Uncertainty for Gain for 3550-3650 MHz

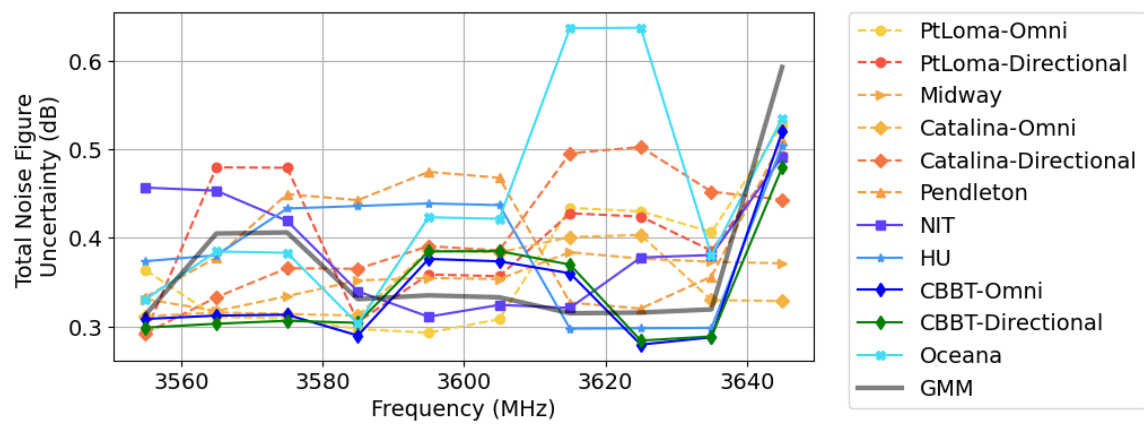


Fig. 46. Total Uncertainty for Noise Figure for 3550-3650 MHz

7. Conclusion

In conclusion, this technical note provides techniques and data related to y-factor calibrations of sensors in the 1-6 GHz operating frequency range. Section 1 provides an introduction to the y-factor method, the definition of sensor used in the note, and the program that deployed sensors over extended periods of time. Section 2, is a general discussion of the properties of noise based calibrations. It also identifies common issues with this type of calibration. Section 3 describes how to build an amplified programmable noise source. Section 4, describes common methods of transferring power measurements from a calibrated noise source to an unknown. Section 5 documents the characterization of many noise diodes and calibration probes using a radiometer. Section 6, is an extensive review of the calibration of field deployed sensors. These sections provide a variety of insights,

- Section 2 : The AMPR can be used to eliminate calibrations contaminated by unwanted signals.
- Section 3 : An amplified programmable noise source provides a way to calibrate in the field for a sensor with large gain differences.
- Section 5 : NIST traceable measurements of a large number of noise diodes shows agreement with the manufacturers specifications and reduces the uncertainty.
- Section 6 : Temperature is the primary driver for gain and noise figure differences in field sensors.
- Section 6 : The temperature dependence in gain for the sensors in this report is $\approx 0.06 \frac{dB}{^{\circ}C}$, indicating a calibration every $1^{\circ}C$ would be required to match the approximate uncertainty of a radiometer calibrated noise diode ($0.05 dB$).
- Section 6 : Frequency dependent changes in gain and noise figure dominate the uncertainty in most cases. This could be decreased by decreasing the bandwidth of channelization (10 MHz) or taking great care to reduce frequency variations in the RF front-end.
- Section 6 : Short-time linear fits to temperature behavior creates an accurate representation as long as excursions of the temperature are sufficient, estimated for these sensors and hourly calibration schedules to be $>12^{\circ}C$ with multiple instances of rates of change of $>1 \frac{^{\circ}C}{hr}$.
- Section 6 : The total uncertainty in the 3550-3650 MHz range is $\approx 0.5 dB$ for gain and $\approx 0.35 dB$ for noise figure in most sensor-frequency combinations.

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Appendix A. Parameters Used in Uncertainty Analysis

The uncertainty analysis for gain and noise figure uses a combination of type A (random uncertainty) and type B (systematic or determined by another method uncertainty) [32]. The tables that follow are summaries of the uncertainties that are sensor and frequency dependent. The short-term linear residual fits for gain (Table 9) and noise figure (Table 10) are included for comparison, however the long-term bi-linear residuals for gain (Table 11) and noise figure (Table 12) are used in the final uncertainty analysis. The band-flatness for gain (Table 13) and noise figure (Table 14) are found to be the largest contribution to uncertainty for most sensor-frequency combinations. There are a few exceptions to this, the next largest uncertainty tends to be a combination of impedance mismatch and extended bi-linear residuals. The uncertainties that are not dependent on sensor and frequency are:

- Calibration Artifact Uncertainty (**0.05 dB** gain and noise figure), assumed to be (log) uniformly distributed.
- Impedance Mismatch Difference Between Calibration Probe and Antenna (**0.2 dB** gain and noise figure). Although this mismatch term is frequency dependent, it is also location dependent and not always easily accounted for because of uncontrolled changes in the location of installation. Thus a single value that represents the median value of the correction using multiple measurements has been adopted. Assumed to be (log) uniformly distributed.
- Uncertainty in y-factor Calibrations (**0.05 dB** gain and **0.07 dB** noise figure). Assumed to be (log) normally distributed.

These uncertainties represent an estimate and uncertainties are assumed symmetric in log space. This approximation has limitations, specifically for powers close to the saturation point or very close to the noise floor the sensor.

Table 7. Total Gain Uncertainty

Frequency (MHz)	3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
Sensor																		
CBBT-Directional	1.63	1.62	0.34	0.38	0.39	0.46	0.47	0.44	0.44	0.39	0.34	0.70	0.70	0.44	2.49	9.34	11.90	11.88
CBBT-Omni	1.60	1.61	0.32	0.37	0.37	0.42	0.43	0.41	0.40	0.35	0.30	0.73	0.74	0.41	2.46	9.31	11.84	11.83
Catalina-Directional	1.03	1.03	0.44	0.57	0.56	0.54	0.56	0.57	0.73	0.76	0.72	0.72	0.61	0.48	1.64	8.35	14.11	14.06
Catalina-Omni	0.99	0.99	0.45	0.50	0.50	0.48	0.57	0.56	0.64	0.64	0.54	0.58	0.59	0.48	1.62	8.33	14.02	13.95
GMM	1.14	1.13	0.45	0.44	0.44	0.51	0.51	0.54	0.54	0.39	0.70	0.88	0.88	0.43	0.96	7.30	18.43	18.29
HU	0.58	0.58	0.35	0.35	0.75	0.75	0.69	0.69	0.33	0.34	0.34	0.50	0.50	0.44	0.94	8.32	15.01	15.15
Midway	0.62	0.63	0.61	0.61	0.80	0.79	0.50	0.53	0.72	0.79	0.79	0.63	0.57	0.58	0.71	7.64	17.43	17.47
NIT	0.54	0.53	0.39	0.49	0.49	0.32	0.29	0.32	0.32	0.40	0.40	0.68	0.68	0.29	0.75	8.53	15.57	15.63
Oceana	1.34	1.34	0.49	0.69	0.67	0.57	0.68	0.68	0.92	0.91	0.71	0.82	0.83	0.60	1.91	8.39	13.58	13.63
Pendleton	1.15	1.15	0.32	0.55	0.63	0.63	0.65	0.65	0.47	0.72	0.71	0.97	0.98	0.83	1.03	8.94	15.00	15.03
PtLoma-Directional	0.92	0.92	0.44	0.67	0.66	0.47	0.49	0.49	0.81	0.80	0.70	0.81	0.82	0.45	1.05	9.08	16.10	16.14
PtLoma-Omni	0.69	0.69	0.52	0.49	0.49	0.44	0.44	0.44	0.81	0.82	0.68	0.86	0.86	0.44	1.10	8.94	16.15	16.22

Table 8. Total Noise Figure Uncertainty

Frequency (MHz)	3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
Sensor																		
CBBT-Directional	1.25	1.25	0.30	0.30	0.31	0.31	0.30	0.39	0.37	0.28	0.29	0.48	0.48	0.35	1.91	7.64	11.59	11.55
CBBT-Omni	1.24	1.25	0.31	0.31	0.31	0.31	0.29	0.37	0.36	0.28	0.29	0.52	0.52	0.34	1.84	7.62	11.72	11.55
Catalina-Directional	0.80	0.79	0.29	0.33	0.37	0.37	0.39	0.39	0.50	0.50	0.45	0.44	0.35	0.30	1.11	6.96	13.96	14.01
Catalina-Omni	0.74	0.75	0.31	0.32	0.31	0.31	0.39	0.38	0.40	0.40	0.33	0.33	0.33	0.31	1.08	6.96	13.98	13.85
GMM	1.03	1.03	0.31	0.40	0.41	0.33	0.34	0.33	0.31	0.32	0.32	0.59	0.60	0.30	0.67	6.05	16.87	16.84
HU	0.59	0.59	0.37	0.38	0.43	0.44	0.44	0.44	0.30	0.30	0.30	0.50	0.51	0.29	0.69	6.67	14.64	14.60
Midway	0.58	0.59	0.33	0.32	0.33	0.35	0.35	0.35	0.38	0.38	0.37	0.37	0.38	0.35	0.52	5.81	17.02	16.87
NIT	0.47	0.47	0.46	0.45	0.42	0.31	0.31	0.32	0.32	0.38	0.38	0.49	0.49	0.32	0.64	6.88	15.06	15.19
Oreana	0.78	0.78	0.33	0.38	0.38	0.30	0.42	0.42	0.64	0.64	0.38	0.54	0.53	0.32	1.35	6.64	13.38	13.36
Pendleton	0.85	0.84	0.33	0.38	0.45	0.44	0.47	0.47	0.33	0.32	0.36	0.51	0.52	0.52	0.69	7.35	15.01	15.00
PtLoma-Directional	0.77	0.77	0.29	0.48	0.48	0.31	0.36	0.36	0.43	0.42	0.39	0.49	0.49	0.28	0.70	7.30	15.29	15.27
PtLoma-Omni	0.53	0.53	0.36	0.31	0.31	0.30	0.29	0.31	0.43	0.43	0.41	0.53	0.53	0.30	0.76	7.16	15.51	15.54

Table 9. Standard Deviation of Single Term Gain Fit Residuals

Frequency (MHz)	3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
Sensor																		
CBBT-Directional	0.10	0.10	0.10	0.10	0.10	0.10	0.10	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11
CBBT-Omni	0.10	0.10	0.10	0.10	0.10	0.10	0.10	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11	0.11
Catalina-Directional	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.16	0.16	0.16	0.16	0.16	0.17	0.17	0.17
Catalina-Omni	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.16	0.16	0.16	0.16	0.16	0.17	0.17	0.17
GMM	0.16	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.15	0.16	0.16	0.16	0.16	0.14	0.23
HU	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.12	0.13	0.18
Midway	0.18	0.18	0.18	0.17	0.17	0.17	0.17	0.17	0.17	0.18	0.18	0.18	0.18	0.18	0.18	0.18	0.19	0.47
NIT	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.09	0.10
Oceana	0.14	0.13	0.13	0.13	0.13	0.13	0.13	0.13	0.13	0.14	0.14	0.14	0.14	0.14	0.15	0.15	0.14	0.30
Pendleton	0.07	0.07	0.06	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.09	0.09
PtLoma-Directional	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.03	0.08
PtLoma-Omni	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.03	0.08

Table 10. Standard Deviation of Single Term Noise Figure Fit Residuals

Frequency (MHz)	3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
Sensor																		
CBBT-Directional	0.021	0.018	0.020	0.020	0.020	0.019	0.020	0.020	0.020	0.020	0.020	0.020	0.021	0.021	0.019	0.022	0.028	0.037
CBBT-Omni	0.021	0.018	0.020	0.020	0.020	0.019	0.020	0.020	0.020	0.020	0.020	0.020	0.021	0.021	0.019	0.022	0.028	0.037
Catalina-Directional	0.027	0.025	0.025	0.025	0.024	0.024	0.024	0.024	0.023	0.029	0.029	0.027	0.028	0.028	0.028	0.030	0.048	0.054
Catalina-Omni	0.027	0.025	0.025	0.025	0.024	0.023	0.024	0.024	0.023	0.029	0.029	0.027	0.028	0.028	0.028	0.030	0.048	0.053
GMM	0.037	0.033	0.032	0.031	0.030	0.029	0.029	0.029	0.026	0.036	0.038	0.051	0.054	0.057	0.053	0.054	0.046	0.168
HU	0.070	0.069	0.068	0.069	0.084	0.083	0.073	0.083	0.080	0.081	0.066	0.064	0.060	0.063	0.060	0.061	0.082	0.150
Midway	0.030	0.030	0.029	0.053	0.068	0.082	0.081	0.082	0.081	0.068	0.069	0.065	0.073	0.074	0.087	0.032	0.048	0.466
NIT	0.014	0.014	0.014	0.014	0.024	0.014	0.038	0.014	0.014	0.022	0.021	0.021	0.017	0.017	0.017	0.019	0.056	0.083
Oreana	0.026	0.023	0.023	0.023	0.021	0.021	0.022	0.021	0.020	0.025	0.025	0.023	0.024	0.026	0.024	0.029	0.038	0.285
Pendleton	0.012	0.011	0.010	0.011	0.010	0.010	0.011	0.010	0.009	0.010	0.011	0.010	0.012	0.012	0.011	0.013	0.053	0.052
PtLoma-Directional	0.005	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.003	0.005	0.023	0.090
PtLoma-Omni	0.005	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.004	0.003	0.005	0.023	0.090

Table 11. Standard Deviation of Two Term Gain Fit Residuals

Sensor	Frequency (MHz)		3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
	Schedule																			
CBRT-Directional CBRT-Omni Catalina-Directional Catalina-Omni GMM	20250419_seadog16_dataproduct		0.09	0.09	0.09	0.09	0.09	0.09	0.09	0.09	0.09	0.10	0.10	0.10	0.10	0.10	0.10	0.10	0.10	0.10
	20250224_seadog06_dataproduct		0.06	0.05	0.05	0.06	0.05	0.05	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.06	0.07	0.07
	20250419_seadog22_dataproduct		0.18	0.18	0.18	0.18	0.18	0.17	0.18	0.18	0.18	0.18	0.19	0.19	0.19	0.20	0.19	0.19	0.19	0.20
	20250419_seadog12_dataproduct		0.18	0.18	0.18	0.18	0.18	0.17	0.18	0.18	0.18	0.18	0.19	0.19	0.19	0.20	0.19	0.19	0.19	0.20
	20250419_seadog14_dataproduct		0.15	0.14	0.14	0.14	0.14	0.14	0.14	0.14	0.14	0.14	0.15	0.14	0.15	0.15	0.15	0.16	0.16	0.17
HU Midway NIT Oreana Pendleton PtLoma-Directional PtLoma-Omni	20250419_seadog04_dataproduct		0.08	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.08	0.12	0.17
	20250419_seadog03_dataproduct		0.21	0.21	0.22	0.21	0.21	0.21	0.21	0.21	0.21	0.21	0.21	0.21	0.22	0.22	0.22	0.22	0.23	0.60
	20240828_ContinuousMeasureSeadog05		0.05	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.06	0.10	0.09
	20250419_seadog10_dataproduct		0.17	0.17	0.17	0.18	0.17	0.17	0.18	0.18	0.17	0.18	0.19	0.18	0.19	0.19	0.19	0.18	0.17	0.34
	20250419_seadog08_dataproduct		0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.08	0.09	0.11	0.10
	20250715_seadog32_6dB_panel		0.15	0.16	0.16	0.16	0.16	0.16	0.16	0.16	0.16	0.15	0.16	0.15	0.15	0.15	0.15	0.15	0.16	0.18
	20250715_seadog32_6dB_omni		0.15	0.16	0.16	0.16	0.16	0.16	0.16	0.16	0.16	0.15	0.16	0.15	0.15	0.15	0.15	0.15	0.16	0.18

Table 12. Standard Deviation of Two Term Noise Figure Fit Residuals

Sensor	Frequency (MHz)		3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
	Schedule																			
CBRT-Directional CBRT-Omni Catalina-Directional Catalina-Omni GMM	20250419_seadog16_dataproduct		0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.03	0.05	0.05
	20250224_seadog06_dataproduct		0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.01	0.02	0.03	0.04
	20250419_seadog22_dataproduct		0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.06	0.06
	20250419_seadog12_dataproduct		0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.06	0.06
	20250419_seadog14_dataproduct		0.04	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.04	0.04	0.05	0.05	0.05	0.05	0.06	0.08	0.58
HU Midway NIT Oreana	20250419_seadog04_dataproduct		0.06	0.05	0.05	0.05	0.06	0.06	0.06	0.06	0.05	0.05	0.05	0.05	0.05	0.05	0.04	0.05	0.12	0.18
	20250419_seadog03_dataproduct		0.04	0.07	0.08	0.07	0.08	0.10	0.10	0.10	0.10	0.08	0.08	0.08	0.09	0.09	0.11	0.04	0.08	0.58
	20240828_ContinuousMeasureSeadog05		0.04	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.04	0.03	0.03	0.03	0.03	0.04	0.09	0.09
	20250419_seadog10_dataproduct		0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.04	0.03	0.03	0.04	0.03	0.03	0.07	0.31
	20250419_seadog08_dataproduct		0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.02	0.07	0.07
PtLoma-Directional PtLoma-Omni	20250715_seadog32_6dB_panel		0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.02	0.02	0.03	0.03	0.03	0.03	0.03	0.07	0.13
	20250715_seadog32_6dB_omni		0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.03	0.02	0.02	0.03	0.03	0.03	0.03	0.03	0.07	0.13

Table 13. Gain Band Flatness

Frequency (MHz)	3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
Sensor																		
CBBT-Directional	1.37	1.37	0.10	0.20	0.20	0.30	0.30	0.27	0.26	0.19	0.07	0.54	0.54	0.25	2.15	8.07	10.23	10.23
CBBT-Omni	1.37	1.37	0.13	0.21	0.21	0.27	0.27	0.26	0.25	0.18	0.08	0.59	0.59	0.27	2.10	8.05	10.26	10.26
Catalina-Directional	0.81	0.81	0.04	0.31	0.31	0.31	0.31	0.31	0.52	0.52	0.49	0.49	0.35	0.10	1.36	7.20	12.19	12.19
Catalina-Omni	0.76	0.76	0.10	0.20	0.20	0.20	0.31	0.31	0.40	0.40	0.24	0.31	0.31	0.11	1.34	7.18	12.13	12.13
GMM	0.92	0.92	0.20	0.19	0.19	0.29	0.29	0.32	0.32	0.08	0.51	0.68	0.68	0.13	0.75	6.37	15.91	15.91
HU	0.43	0.43	0.16	0.16	0.60	0.60	0.54	0.54	0.13	0.16	0.16	0.35	0.35	0.29	0.77	7.25	13.08	13.08
Midway	0.33	0.33	0.30	0.30	0.54	0.54	0.06	0.15	0.45	0.54	0.54	0.34	0.24	0.24	0.44	6.63	15.10	15.10
NIT	0.39	0.39	0.25	0.35	0.35	0.15	0.07	0.14	0.14	0.25	0.25	0.55	0.55	0.09	0.61	7.38	13.51	13.51
Oceana	1.09	1.09	0.20	0.46	0.46	0.32	0.46	0.46	0.70	0.70	0.48	0.61	0.61	0.33	1.62	7.21	11.76	11.76
Pendleton	0.97	0.97	0.08	0.40	0.48	0.48	0.50	0.50	0.31	0.56	0.56	0.81	0.81	0.67	0.85	7.77	13.03	13.03
PtLoma-Directional	0.72	0.72	0.14	0.45	0.45	0.20	0.23	0.23	0.60	0.60	0.49	0.62	0.62	0.17	0.84	7.86	13.96	13.96
PtLoma-Omni	0.49	0.49	0.28	0.23	0.23	0.14	0.14	0.13	0.61	0.61	0.47	0.66	0.66	0.16	0.89	7.77	14.05	14.05

Table 14. Noise Figure Band Flatness

Frequency (MHz)	3535	3545	3555	3565	3575	3585	3595	3605	3615	3625	3635	3645	3655	3665	3670	3685	3695	3705
Sensor																		
CBBT-Directional	1.06	1.06	0.09	0.11	0.11	0.10	0.23	0.23	0.21	0.05	0.06	0.34	0.34	0.18	1.63	6.64	10.03	10.03
CBBT-Omni	1.05	1.05	0.12	0.13	0.13	0.07	0.22	0.22	0.20	0.04	0.08	0.38	0.38	0.18	1.58	6.62	10.06	10.06
Catalina-Directional	0.64	0.64	0.07	0.15	0.20	0.20	0.23	0.23	0.36	0.36	0.30	0.30	0.18	0.09	0.93	6.02	12.13	12.13
Catalina-Omni	0.60	0.60	0.12	0.12	0.12	0.12	0.23	0.23	0.25	0.25	0.15	0.14	0.14	0.09	0.90	6.01	12.07	12.07
GMM	0.86	0.86	0.12	0.25	0.25	0.15	0.15	0.15	0.12	0.12	0.12	0.44	0.44	0.05	0.52	5.23	14.57	14.57
HU	0.44	0.44	0.20	0.21	0.27	0.27	0.28	0.28	0.04	0.04	0.03	0.36	0.36	0.02	0.54	5.78	12.63	12.63
Midway	0.44	0.44	0.06	0.06	0.08	0.08	0.09	0.09	0.16	0.16	0.16	0.16	0.16	0.12	0.35	5.03	14.66	14.66
NIT	0.33	0.33	0.31	0.31	0.27	0.16	0.10	0.14	0.14	0.22	0.22	0.35	0.35	0.13	0.50	5.92	13.00	13.00
Oceana	0.63	0.63	0.15	0.23	0.23	0.09	0.28	0.28	0.49	0.49	0.22	0.40	0.40	0.12	1.14	5.73	11.51	11.51
Pendleton	0.70	0.70	0.16	0.22	0.30	0.30	0.33	0.33	0.15	0.15	0.19	0.37	0.37	0.38	0.54	6.41	13.07	13.07
PtLoma-Directional	0.62	0.62	0.07	0.33	0.33	0.11	0.19	0.19	0.28	0.28	0.23	0.35	0.35	0.02	0.55	6.32	13.33	13.33
PtLoma-Omni	0.39	0.39	0.20	0.11	0.11	0.08	0.07	0.11	0.29	0.29	0.26	0.39	0.39	0.09	0.61	6.22	13.43	13.43

Appendix B. Example Radiometer Calibration of Noise Diode



National Institute of Standards and Technology
Communications Technology Laboratory
325 Broadway
Boulder, CO 80305-3328

Report of Special Test

Noise Source

Device Description:	Keysight Model 346B Serial No. SG61410009
Submitted by:	National Advanced Spectrum and Communication Test Network National Institute of Standards and Technology 325 Broadway MS 67501 Boulder CO 80305
Quantity Measured:	Noise Temperature
Measurements Performed by:	Verified by:
Xifeng Lu 303-497-3049	Dazhen Gu 303-497-3939

The measurements were performed at $23.0\text{ }^{\circ}\text{C} \pm 5\text{ }^{\circ}\text{C}$ and $40\% \pm 5\%$ relative humidity. The applied DC voltage was $28\text{ V} \pm 0.03\text{ V}$ (expanded uncertainties). The estimated uncertainty in the center calibration frequency is 0.01 %. If the submitted noise source is an assembly, composed of a noise source plus an adapter, the assembly must remain intact for the results to be valid. Reported in Table 1 are the Noise Temperature (T), the Excess Noise Ratio (ENR₀), and the associated expanded uncertainties (U) of the device. These results are traceable to the SI (International System of Units) Kelvin through NIST Primary-Standard Noise Sources. The results exclude any contribution of the output connector except as it differs from the NIST primary standard's connector. The results reported here reflect the measurements and our state of knowledge at the time of this calibration or test.

INFORMATION SHEET

The noise temperature measured and reported is the *available* noise temperature, defined to be the available noise power per unit bandwidth divided by Boltzmann's constant. For noise temperatures above $T_0 = 290\text{ K}$, we also report the excess noise ratio delivered into a reflectionless load (ENR₀). It is defined by

$$ENR_0(dB) = 10\log_{10} \left[\frac{T(1-|\Gamma|^2) - T_0}{T_0} \right],$$

where T is the available noise temperature, and Γ is the reflection coefficient of the device under test. If the available ENR is desired, it can be computed directly from the available noise temperature by

$$ENR_{av}(dB) = 10\log_{10} \left(\frac{T - T_0}{T_0} \right).$$

For most devices, the difference between ENR₀ and ENR_{av} is very small. NIST noise-temperature measurements are performed on total-power radiometers, using two primary thermal noise standards, one of which is at ambient temperature and one of which is at cryogenic (liquid nitrogen) temperature. These national primary standards are traceable to the SI unit kelvin (through calibrated thermistors and the boiling temperature of liquid nitrogen). For measurements at 30 MHz and 60 MHz, tunable coaxial standards [1] are used. From 1 GHz to 12.4 GHz, coaxial standards [2] are used, and for 12.4 GHz and above, waveguide/horn standards [3, 4] are used. The radiometers themselves are described in references [1, 5-7]. The NIST radiometers are double-sideband, total-power radiometers. The IF frequency is 0 (*i.e.*, the LO frequency is set to the measurement frequency), and the IF bandwidth B_{IF} ranges from 5 MHz to 20 MHz, depending on the particular radiometer. Thus the reported noise temperature represents an average over a frequency range of $2B_{IF}$ centered at the measurement frequency.

At least three independent measurements (including separate system calibrations, where applicable) of the noise temperature are made at each frequency. The noise source is allowed to warm up before any measurements are made. For many connector types and frequencies, the measurements are made through adapters. The procedure for characterizing the adapter and removing its effect is described in references [8, 9].

The combined standard uncertainty is composed of type-A and type-B uncertainties [10, 11]. Type-A uncertainties (u_A) are those that are measured and determined by statistical methods, such as the standard deviation of the mean of several independent measurements of the quantity of interest. Type-B uncertainties (u_B) are those determined by other means, such as estimates of systematic uncertainties. The uncertainty reported is the expanded uncertainty, given by

$$U = 2\sqrt{u_A^2 + u_B^2}$$

This corresponds approximately to a 95% confidence level. Details of the uncertainty analysis can be found in references [5-7, 12].

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Noise Source
Keysight
Model 346B, SN SG61410009

Table 1
Measurement results for
Noise Source
Model 346B, S/N SG61410009

Center Calibration Frequency GHz	T Kelvins	U Kelvins	ENR0 dB	U dB
3.535	9190.39	96.72	14.87	0.05
3.545	9193.51	97.42	14.87	0.05
3.555	9194.08	97.1	14.87	0.05
3.565	9172.71	97.1	14.86	0.05
3.575	9157.18	96.62	14.85	0.05
3.585	9141.88	98.14	14.84	0.05
3.595	9135.08	99.15	14.84	0.05
3.605	9116.99	96.52	14.83	0.05
3.615	9101.7	97.25	14.82	0.05
3.625	9112.38	101.72	14.83	0.05
3.635	9111.46	96.52	14.83	0.05
3.645	9127.49	98.53	14.84	0.05
3.655	9124.48	96.68	14.84	0.05
3.665	9144.02	97.14	14.85	0.05
3.675	9148.53	97.27	14.85	0.05
3.685	9128.88	98.23	14.84	0.05
3.695	9107.52	97	14.83	0.05
3.705	9085.71	96.53	14.82	0.05

Appendix C. Gain and Noise Figure Distributions



Fig. 47. Gain Distribution of HU Sensor



Fig. 48. Noise Figure Distribution of HU Sensor

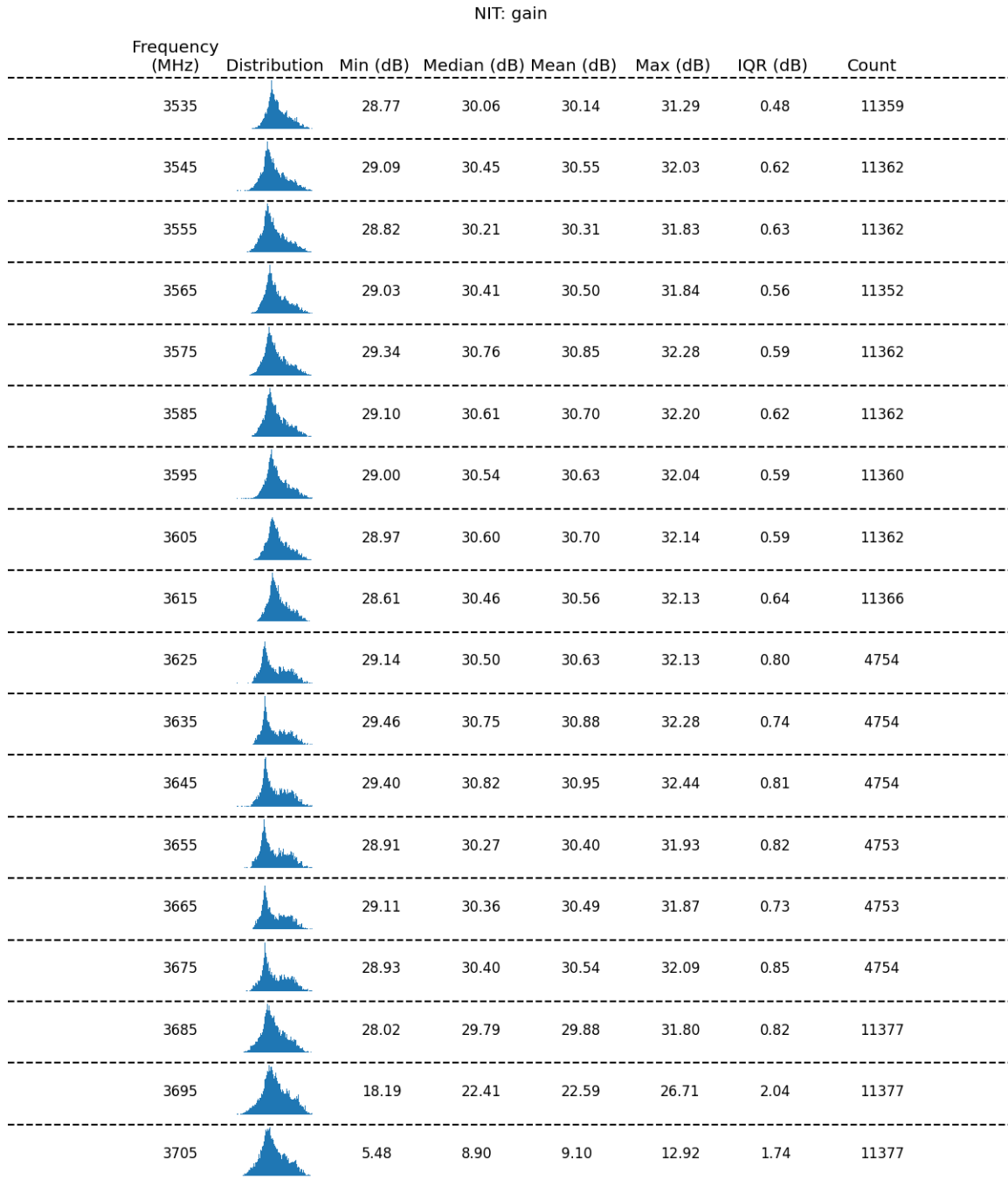


Fig. 49. Gain Distribution of NIT Sensor

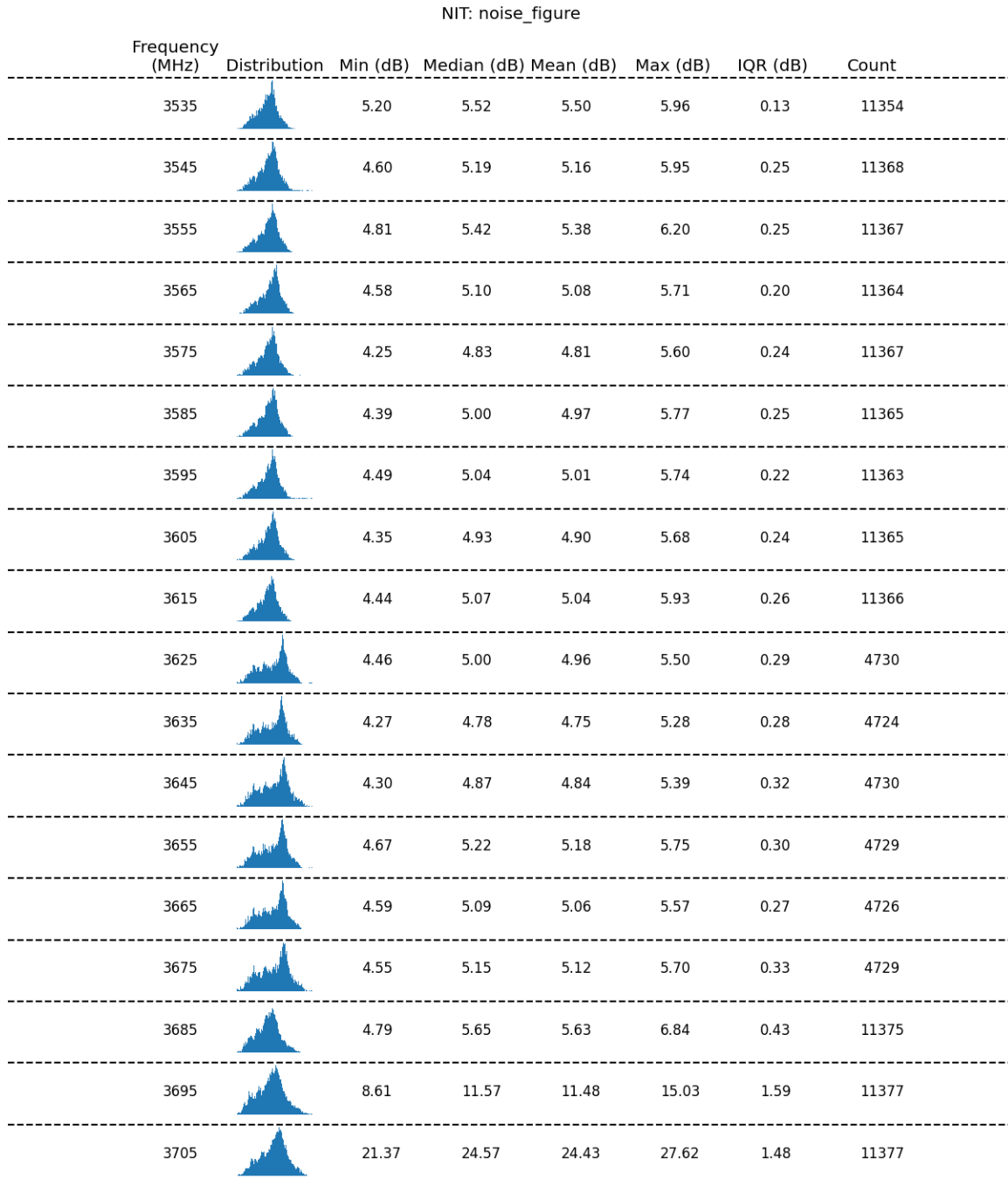


Fig. 50. Noise Figure Distribution of NIT Sensor

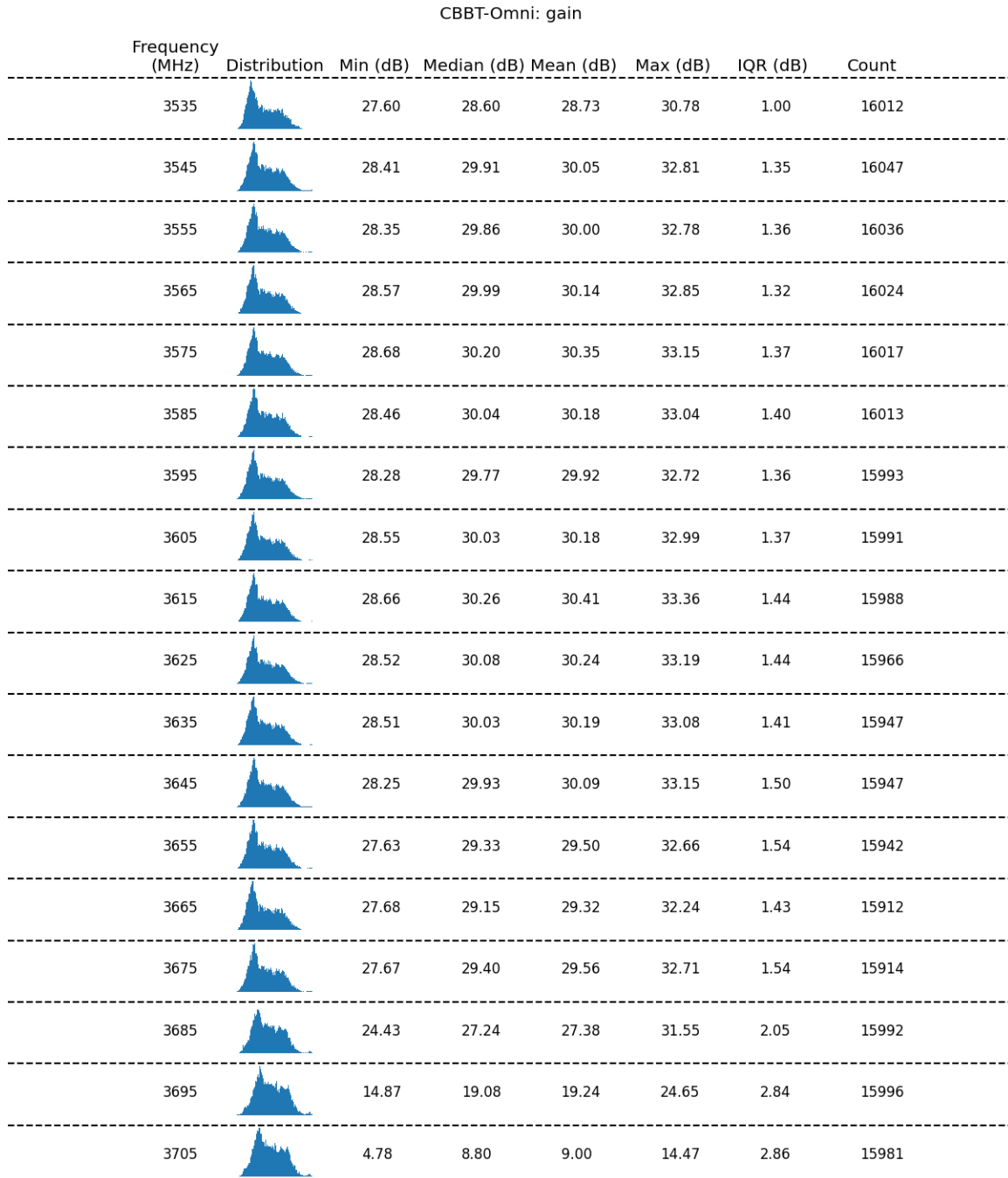


Fig. 51. Gain Distribution of CBBT-Omni Sensor



Fig. 52. Noise Figure Distribution of CBBT-Omni Sensor

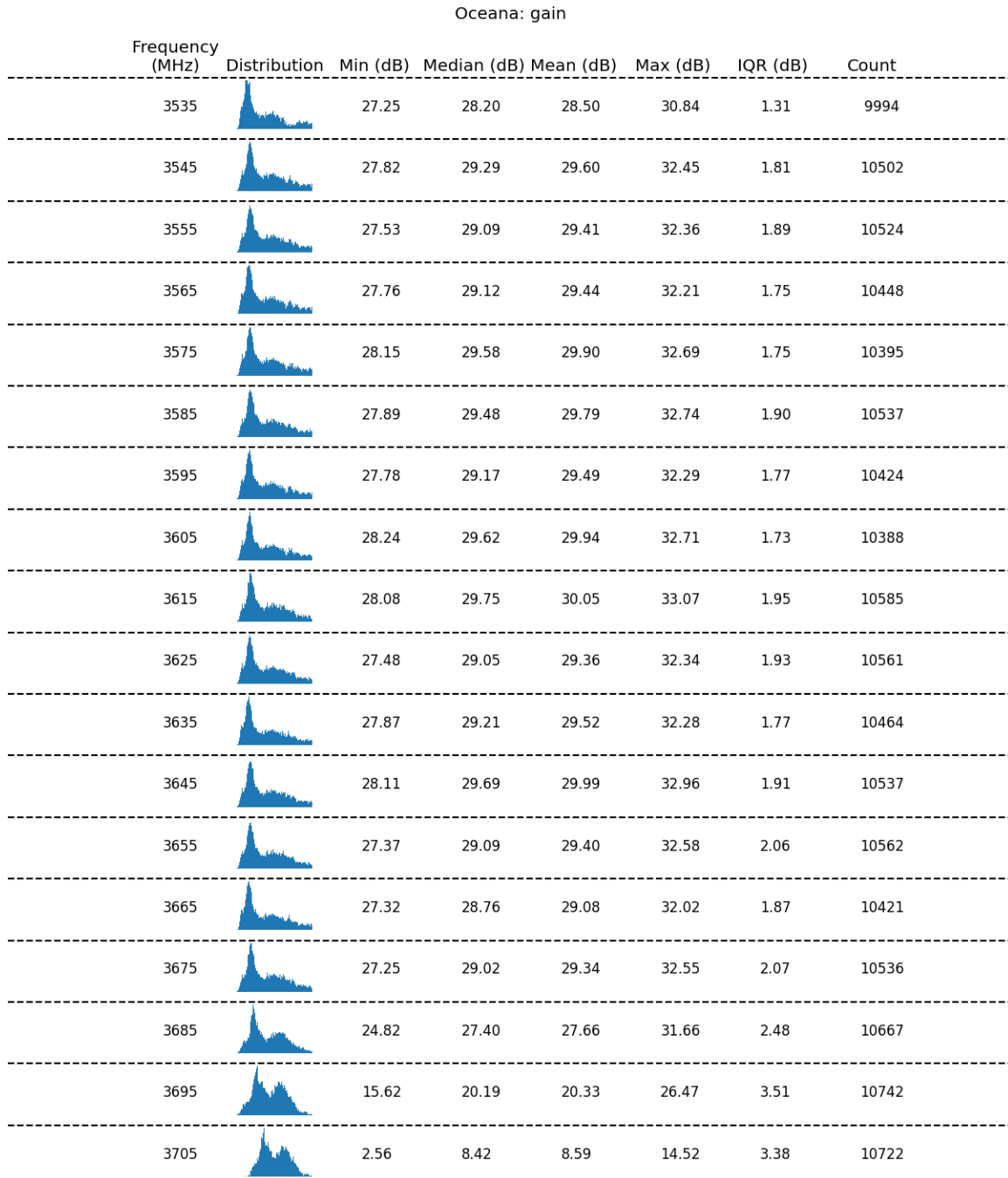


Fig. 53. Gain Distribution of Oceana Sensor



Fig. 54. Noise Figure Distribution of Oceana Sensor

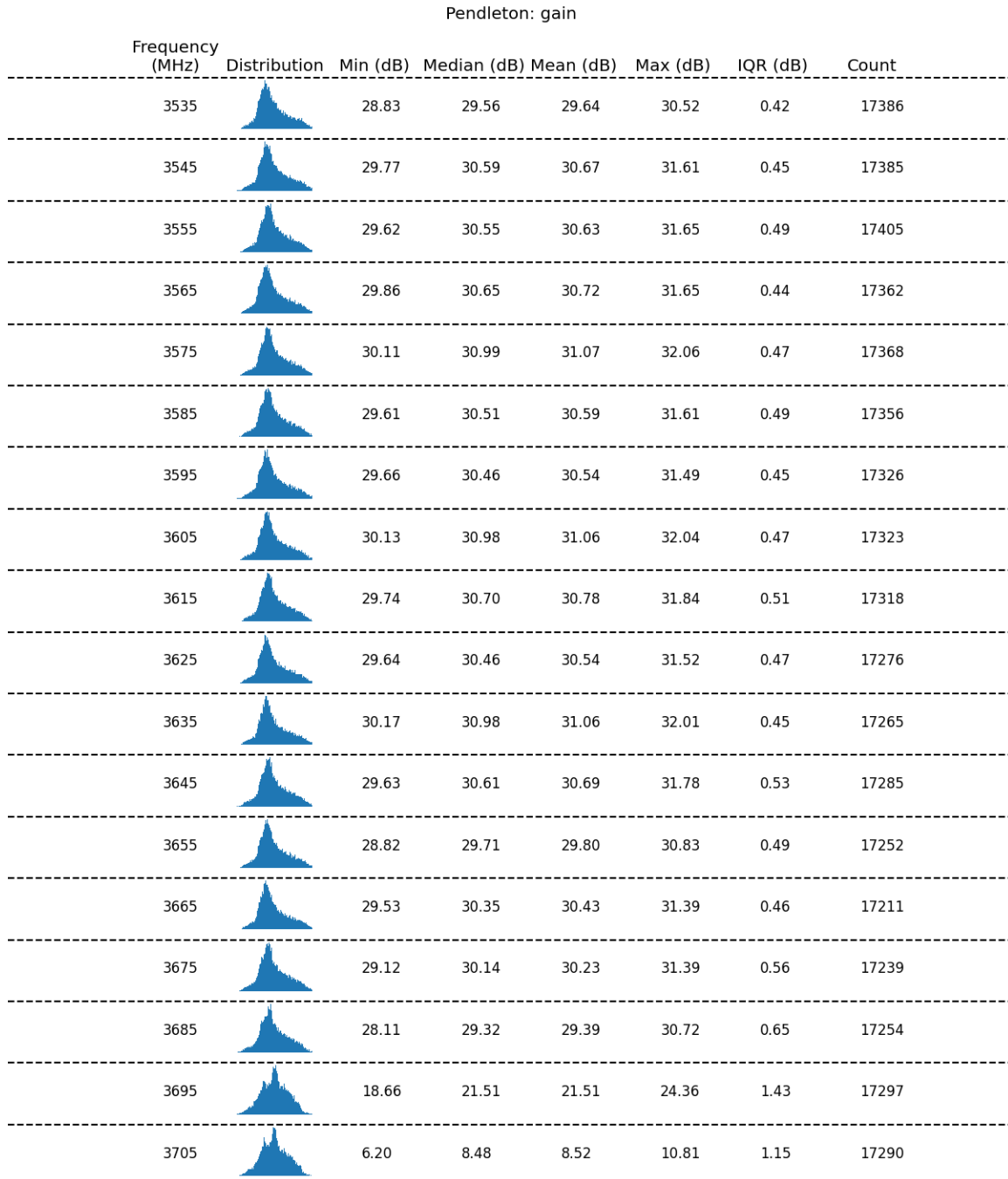


Fig. 55. Gain Distribution of Pendleton Sensor



Fig. 56. Noise Figure Distribution of Pendleton Sensor

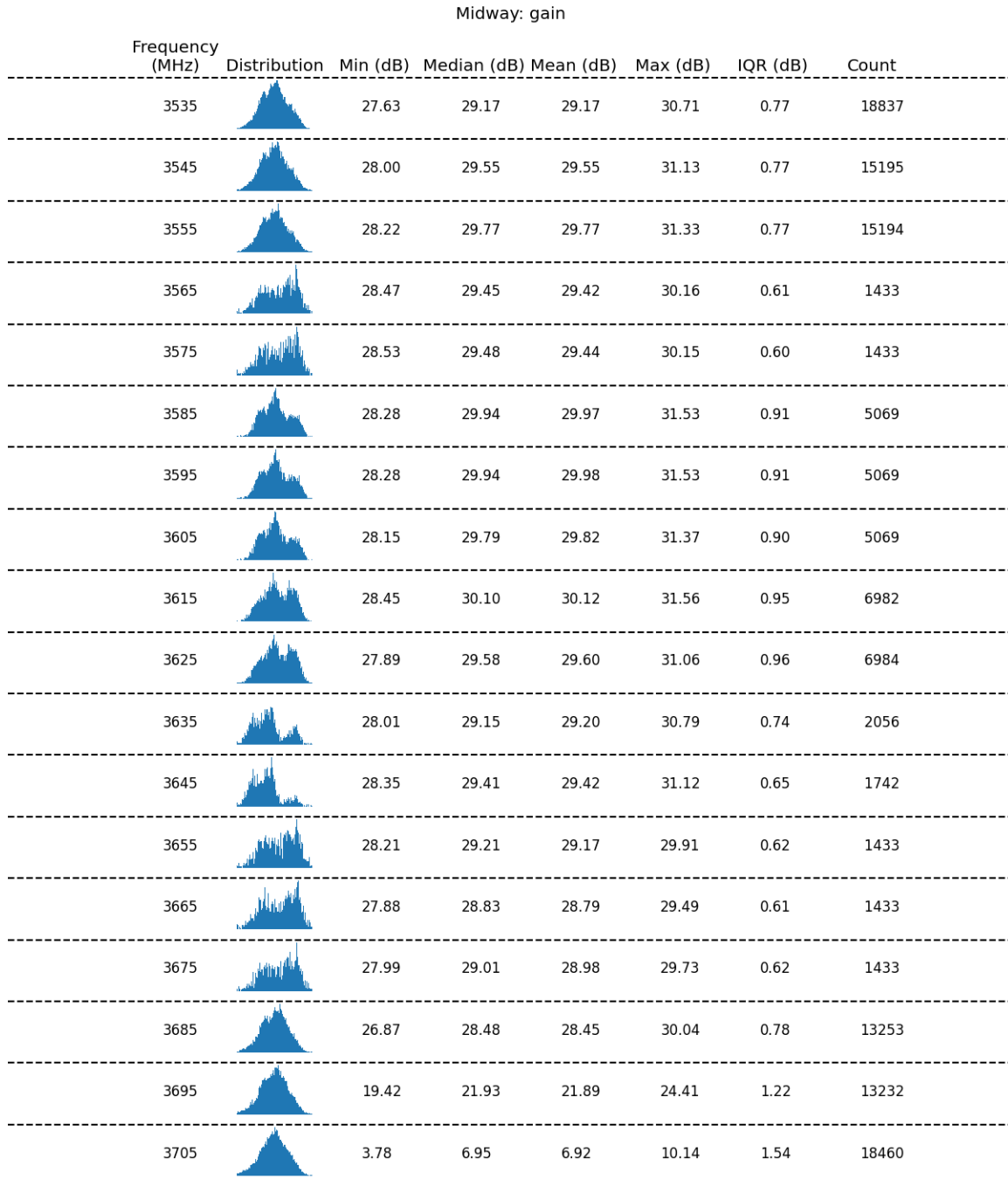


Fig. 57. Gain Distribution of Midway Sensor



Fig. 58. Noise Figure Distribution of Midway Sensor

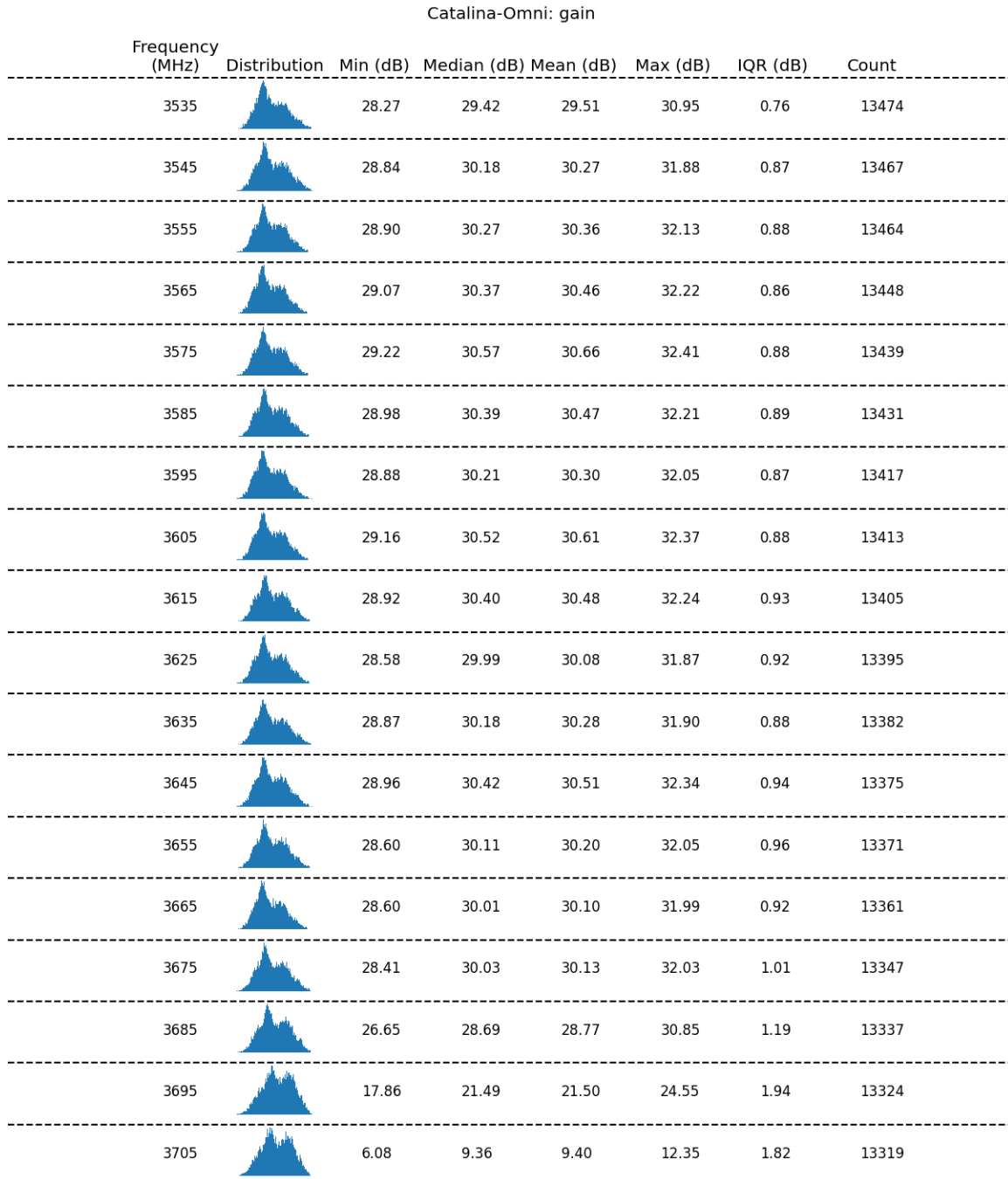


Fig. 59. Gain Distribution of Catalina-Omni Sensor



Fig. 60. Noise Figure Distribution of Catalina-Omni Sensor

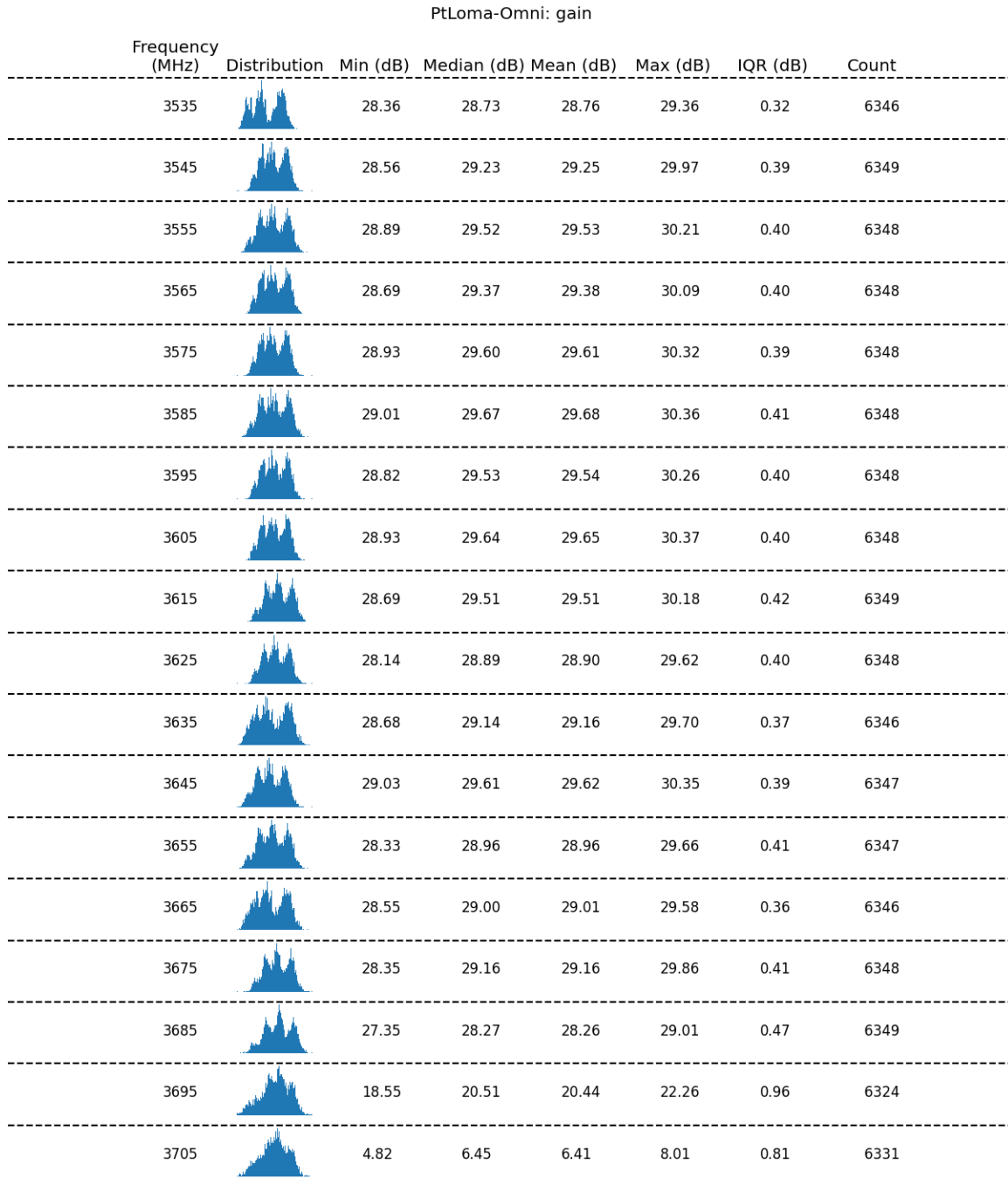


Fig. 61. Gain Distribution of PtLoma-Omni Sensor



Fig. 62. Noise Figure Distribution of PtLoma-Omni Sensor

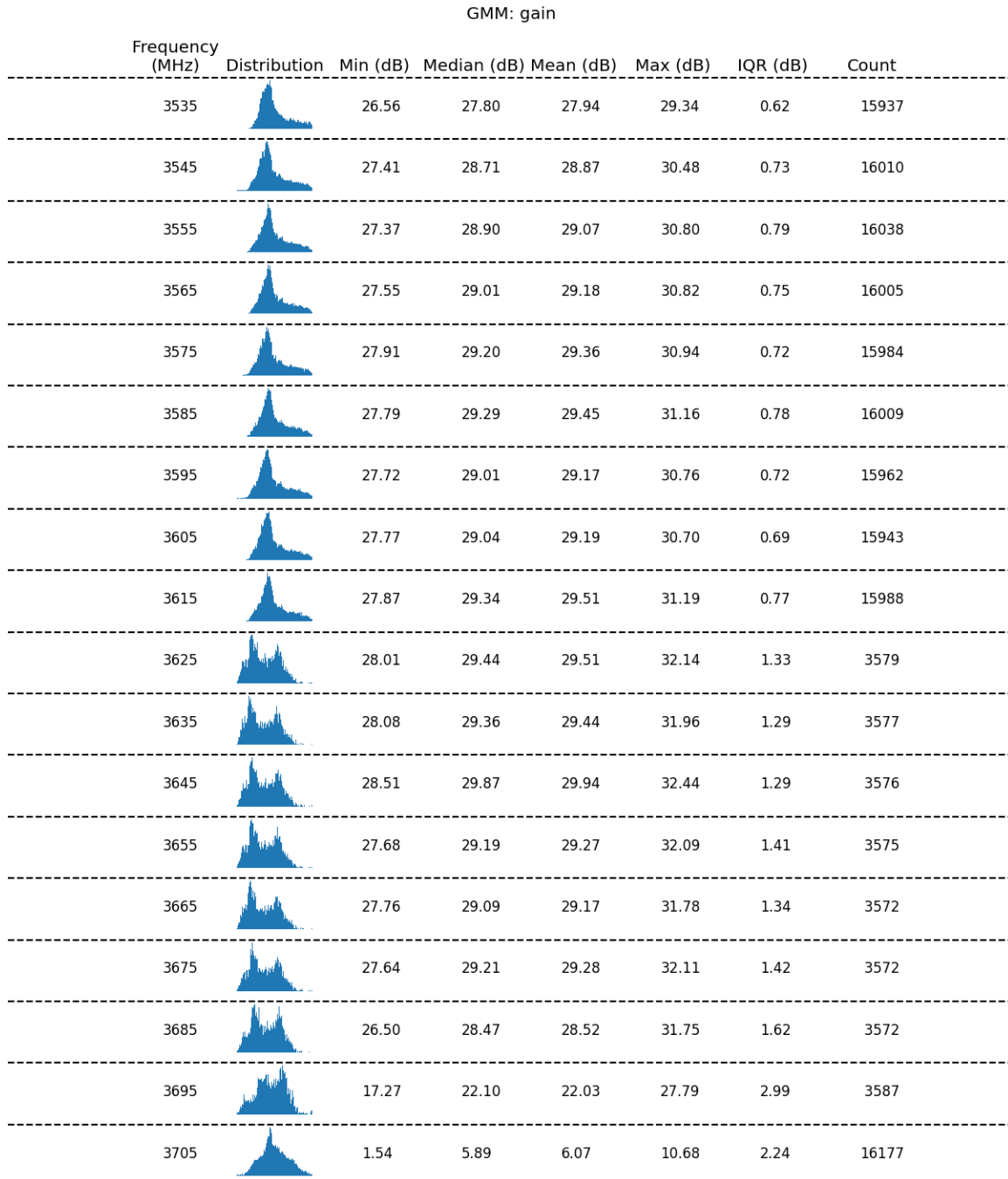


Fig. 63. Gain Distribution of GMM Sensor

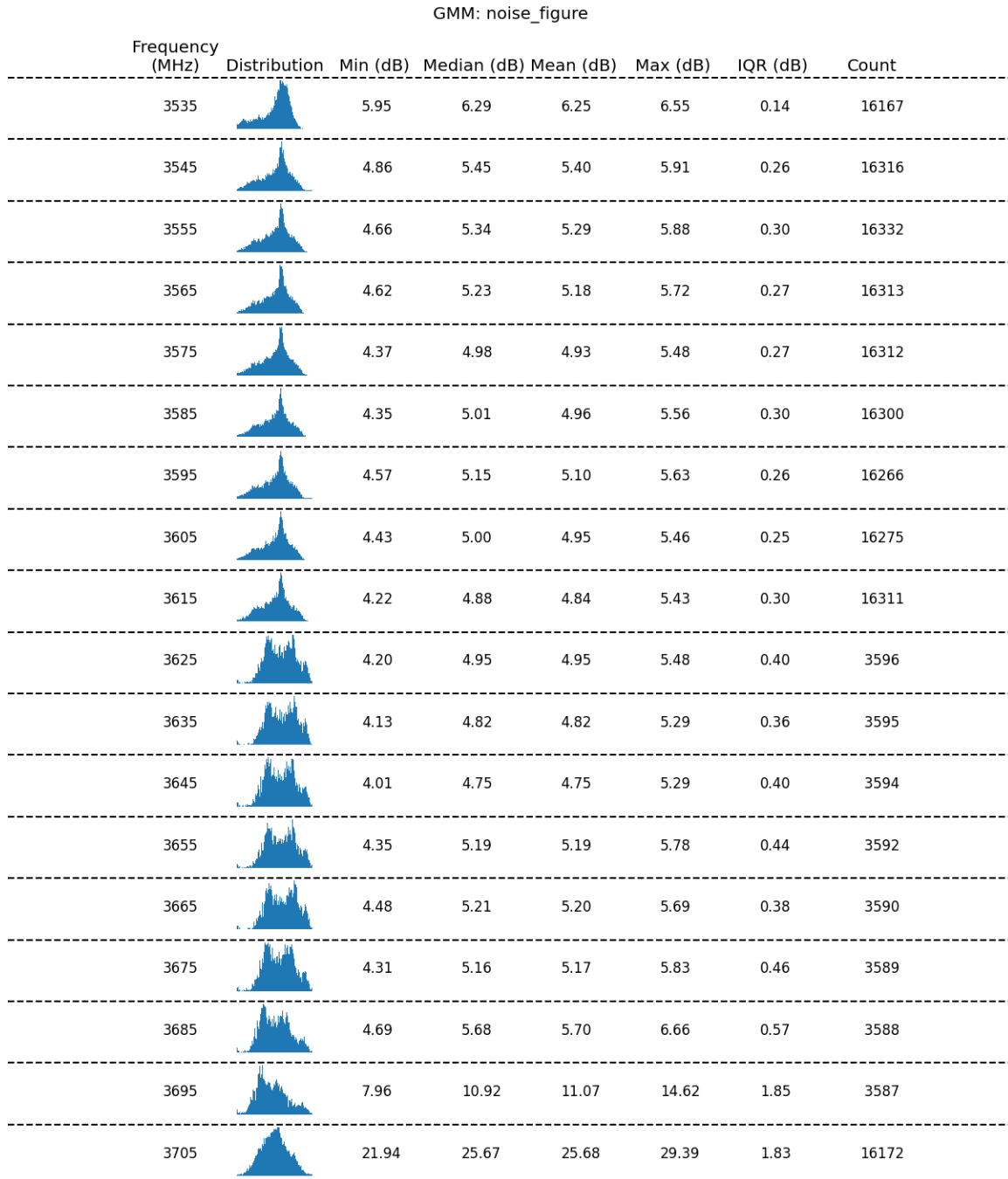


Fig. 64. Noise Figure Distribution of GMM Sensor