ACCURACY OF NOISE MEASUREMENTS

by

Judith Ann Furlong

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ABSTRACT

Two systems that measure the white noise spectrum of voltage reference diodes, in the frequency range from 1 kHz to 100 kHz, were analyzed to determine their accuracy. Limitations to the accuracy of each of these system were identified. Recommendations were make for improving the accuracy of these existing systems.

The results of the analysis on these systems show that the system, which used the HP 3562A Dynamic Signal Analyzer to measure noise, had an accuracy of one-half percent. The other system, which used the Fluke 8506A Thermal RMS Multimeter to measure noise, was expected to have the same, if not slightly better, accuracy.

Thesis Supervisor: Prof. Hermann Haus

Title: Professor of Electrical Engineering

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Judith A. Furlong

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Section 1 INTRODUCTION

1.1 Introduction

This thesis is being conducted as part of a research project in which noise is being used to study the physics of voltage reference diodes. The noise the diode produces reflects the ratio of tunneling to avalanche current within the diode. The tunneling and avalanche mechanisms of these diodes have neutron radiation coefficients of opposite sign. The goal of the project is to see if it is possible to correlate the noise characteristics of the diode with its radiation characteristics.

If correlation between the noise and radiation characteristics exist, it may be possible to use noise measurements to screen production diodes. Assume a manufacturer has a lot of radiation-hard diodes and wishes to screen these devices to sell only those which meet certain specifications. The manufacturer makes a measurement of the noise of all the diodes in the lot. The diodes will be grouped by the amount of noise they display. Samples from each of the groups will be radiated and their radiation characteristics will be determined. The manufacturer will check to see if diodes from the same group exhibit the same radiation characteristics. If this is true, the manufacturer can assume that the other diodes from the group, which were not radiated, will display the same radiation characteristics. Diodes from different groups are not expected to have similar radiation characteristics. The manufacturer will sell only diodes from groups that meet the specifications.

To be able to group diodes using their noise characteristics and determine if there is correlation between noise and radiation characteristics, it will be necessary to make accurate noise measurements. At this point, it is uncertain how accurate the noise measurements must be; however, one opinion suggests the measurements must be highly accurate. In any case, it will be necessary to determine the accuracy of our noise measurements.

Measuring noise to a high degree of accuracy is quite difficult. The most accurate noise measurements to date were performed by W. Lukaszek as part of his doctoral thesis at the University of Florida in 1974.[1] His measurements, which we consider state-of-the-art, had two percent accuracy in the sense that he could measure noise from resistors and determine their accuracy to two percent based on the noise measurements.

This thesis will look at the problems of obtaining accurate noise measurements, particularly with the measurement systems built for this project. Two different noise measurement systems will be evaluated. The limitations of noise measurements will be explored. Various methods of noise measurement will be studied. Thus, the goal of the thesis is threefold: to determine the accuracy of two systems; to identify which system makes the most accurate measurements; and to recommend changes to the existing systems which would improve the accuracy of their measurements.

1.2 Organization of Thesis

Section 2 presents the necessary background material for this thesis. There are five major topics covered in this section. First, more detail about the tunneling and avalanche mechanisms of diodes is presented. Second, the limitations of noise measurements along with ways to overcome these limitations are presented. Third, several noise measurement methods are discussed. Fourth, the difference between conventional and state-of-the-art noise measurements is explained. Fifth, a more detailed description of the state-of-the-art noise measurements conducted by Weislaw Lukaszek [1] is presented.

In section 3, descriptions of the two noise measurement systems, to be studied in this thesis, can be found. The description of each system begins with a generalized description of the block diagram of the system and proceeds to more detailed descriptions of the circuit portion of the system, the commercial equipment used in the system and the measurement procedure used with the system. Included in these descriptions are explanations of why a particular type of circuit or piece of equipment is used in the system. Many of these explanations reflect low-noise design considerations and techniques. In the case of commercial equipment, pertinent specifications as well as brief explanations of how the device is used are given.

In section 4, the accuracy of the two measurement systems is determined. The accuracy of the first system is discussed separately from that of the second system. The discussion of the accuracy of each system begins with analysis of various aspects of the measurement system that could affect the measurement accuracy. These aspects include design, measurement and calibration procedures, averaging and sampling time. Through these analyses, the limitations to measurement accuracy for the system are identified. A number that describes the accuracy of the system is then determined. Finally, recommendations for improving the accuracy of the system are presented.

In section 5, a generalized discussion of accuracy of noise measurements is presented. Common limitations to accurate noise measurements and recommendations for overcoming some of these limitations are briefly discussed. The section concludes by making an estimate of how accurately an arbitrary noise signal can be measured.

Section 6 summarizes the important conclusions reached about the accuracy of noise measurements made with each system and in general. Recommendations for further study of the accuracy of noise as well as suggestions for other noise measurement systems are included in this section.

A number of appendices is included to describe certain topics in more detail and provide other necessary information. Appendix A. contains a glossary of noise related terms used in the thesis. Appendix B. presents the noise models for the most common circuit components and describes how they are used. Computer programs used with the two measurement systems are included in Appendix C. Appendix D. contains the specifications for the Hewlett-Packard 3562A Dynamic Signal Analyzer and the Fluke 8506A Thermal True RMS Multimeter. Appendix E. contains the specifications for selected components used in the circuit portion of both systems. A step-by-step calculation of the noise produced by portions of the circuits used in both systems appears in Appendix F.

Section 2 BACKGROUND

2.1 Tunneling and Avalanche Breakdown*

A diode or p-n junction is said to break down and conduct large currents when a sufficiently high field is applied to the junction. If a diode is reverse-biased there are two different mechanisms of breakdown: tunneling and avalanching.

Tunneling breakdown, also referred to as Zener breakdown, since it is the type of breakdown that occurs in Zener diodes, takes its name from the quantum mechanical tunneling process that is occurring within the diode. When tunneling occurs, the covalent bonds between neighboring atoms in the depletion region are broken, generating holes and electrons. Valence band electrons "tunnel" through the energy gap as they move from the valence to conduction band. Electron-hole pairs are produced by this process and increase the reverse current of the diode.

The second type of reverse breakdown is avalanche breakdown. Avalanche breakdown occurs when the field applied to the junction speed up the mobile carriers in the space charge layer, so that collisions between the carriers and the lattice of the semiconductor occur. These collisions knock electrons from the covalent bonds free, producing holes and electrons. These new carriers increase the reverse

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* References [4] through [7] were used in writing this section. Consult these references for more detailed information about tunneling and avalanche breakdown.

current of the diode. The new carriers may also produce more free electrons and holes through collisions of their own. With each new carrier knocking out more carriers, the reverse current of the diode is increased or multiplied and can become quite large. Because of this multiplication, avalanche breakdown is sometimes called avalanche multiplication.

When referring to reverse breakdown in diodes, a reverse breakdown voltage is often mentioned. This is the voltage at which breakdown begins to occur in the diode. The type of mechanism that causes the breakdown of the diode can be predicted by the range into which the reverse breakdown voltage falls. For silicon diodes breaking down at reverse biases less than 5 volts, the breakdown mechanisms is tunneling. If the diodes breaks down at voltages between 5 and 7 volts, the breakdown mechanism is a combination of tunneling and avalanching. Finally, if the diode breaks down at a voltage of greater than 7 volts, the breakdown mechanism is avalanching. Semiconductors, other than silicon may have different voltages for the boundaries of these ranges.

In some instances it is desirable to know the amount of current produced by tunneling breakdown and the amount produced by avalanche breakdown. One instance where knowing this ratio may be useful is in the processing of radiation-hard diodes. Tunneling breakdown and avalanche breakdown do have some distinguishing characteristics. These mechanisms have temperature and radiation coefficients of opposite signs. Tunneling has a negative temperature coefficient, while avalanching has a positive temperature coefficient. The noise ratio (See Appendix A for definition) of the diode may be used to distinguish between tunneling and avalanche currents.

2.2 Limitations of Noise Measurements

Measuring noise is different from and often more difficult than measuring other types of electric signals. There are several limitations or problems that one faces in measuring noise that one does not encounter in other types of measurements. In most cases, certain precautions and/or measurement schemes can be used to overcome or to minimize these problems. This section will briefly describe the limitations of noise measurements and propose some ways in which these problems may be surmounted.

The nature and characteristics of noise are responsible for several of the limitations in measuring it. First one must be sure that the noise being measured does not exhibit 1/f noise or low frequency noise. 1/f noise has a spectral density that increases without limit as the frequency decreases and is undesirable to measure because of the inaccuracies it contributes to the average value of the noise. To avoid the problem of 1/f noise, the noise must be measured in a region in which its spectrum is flat. This means that the low frequency components of the signal to be measured have to be eliminated through some type of bandlimiting or filtering. Second, the amplitude of noise is small, usually in the nanovolt range for a noise voltage. So the signal must be amplified to be detected by a meter. Third, the white or broadband nature of noise requires that the signal be band-limited (filtered) in some stage of the measurement system as well as averaged over a long period of time to insure accurate measurement of the noise signal.

Bandlimiting is a necessary requirement for a noise measurement system because it eliminates the 1/f noise and more importantly compensates for the white nature of noise. The white noise signal is spread out in the frequency domain, with energy beyond frequencies where noise amplifiers perform well. One has to chose a frequency domain, so that measurements have acceptably low sensitivity to poorly controlled parameters such as stray capacitance and operational amplifier gain-bandwidth. The way to insure that measurements are made only over a certain range of frequencies is through bandlimiting. Bandlimiting is achieved by filtering the signal to be measured so that only the portion of the signal within the chosen frequency range reaches the system output. Problems with bandlimiting arise from the filters that are necessary to achieve it. The filters may add noise to the system, so care must be taken when building them to limit the amount of noise they contribute to the system. Another problem with filters is their stability. The frequency range that they are band-limiting or the passband gain can shift slightly due to drift in the components used to make them.

Other limitations encountered in measuring noise are a result of noisy measurement systems. Both custom built circuits as well as commercial equipment, used in measurement system produce noise of their own. If this noise is large as compared to the signal being measured, inaccurate measurements could result.

As noted, electronic components, even if they are low-noise, exhibit some noise. The noise they produce will contribute to the overall noise of the measurement system and to the noise being measured at the output of the system. In any measurement system one must understand what one is measuring. One must verify that the final noise estimate is limited to only the noise of the device that you wish to measure. So in some manner, the noise of the measurement system must be subtracted from the noise measured at the output of the system. This should leave just the noise of the device being tested, the quantity that is desired.

Along the same line as component noise is commercial equipment noise. Since commercial equipment is made from electronic components, it too will be a noise source. Usually the noise of meters is not a problem, because they are designed to insure the noise of the meter does not cause inaccuracies in measurements. Other commercial equipment, like a preamp could significantly add to the noise of the system. Most equipment comes with noise specifications so one has a rough estimate of the extra noise contributed by the equipment. However, when one needs to make accurate noise measurements, like we wish to do, one must measure the noise of equipment exactly. This will insure that the correct amount of noise is subtracted from the total noise.

The last limitation to be discussed, is calibration. A calibration procedure is often used with measurement systems. In the case of a noise measurement system, a calibration process could be used to estimate the system noise. The noise of a DUT may be determined by the difference in output when a DUT is placed in the system and when a calibration signal is applied to the system. Problems with calibration arise from several sources. First, one must insure the accuracy and stability of the calibration. Inaccuracy or drift in such a signal degrades the measurements. The accuracy and stability of a signal can be verified by observing such a signal over time. A second problem with calibration is consistency with the calibration process. One must take care that the exact same steps in the exact same order are taken for each calibration. If such a procedure is not followed measurements could be inaccurate.

2.3 Methods of Measuring Noise

There are several methods for measuring noise. Most of these methods were developed to measure the signal-to-noise ratio of a system. Knowing the signal-to-noise ratio (SNR) is desirable, especially in communication systems, since it tells how much the signal being transmitted through the system is degraded by the noise. Even if one wants to measure a noise parameter other than SNR, these methods can still be useful, since all the methods measure either the equivalent input or output noise of the system. These two parameters are related by the gain of the system. Other noise parameters, like noise spectral density and noise ratio may be derived from the output noise of the system. This section will describe three noise measurement methods: the sine wave; the noise generator; and the correlation. Although bandlimiting is not mentioned in any of these methods, it will be necessary for making noise measurements. The noise quantities measured by these methods are in units of Volts. If spectral density is desired, divide the measured quantity by the square-root of the noise equivalent bandwidth.

2.3.1 Sine Wave Method of Noise Measurement

To illustrate how the sine wave method of noise measurement works, the procedure for measuring the equivalent input noise, as described by Motchenbacher and Fitchen [2], will be used. To determine the input noise with the sine wave method, the output noise and the gain of the system must be measured. The exact procedure for finding the input noise is as follows.

1. Measure the transfer voltage gain K_t.
2. Measure the total output noise E_{no}.
3. Calculate the equivalent input noise E_{ni} by dividing the output

noise by the transfer voltage gain. [2]

Figure 2.3.1 shows the block diagram for measuring the input noise. V_s represents the input sine wave signal or sine wave generator. E_{ni} is the equivalent input noise, which is being measured. Z_s is the source impedance. The system is represented by the amplifier symbol. The equivalent output noise, E_{no}, and the output sine wave signal, V_o are measured at the output terminal of the system.

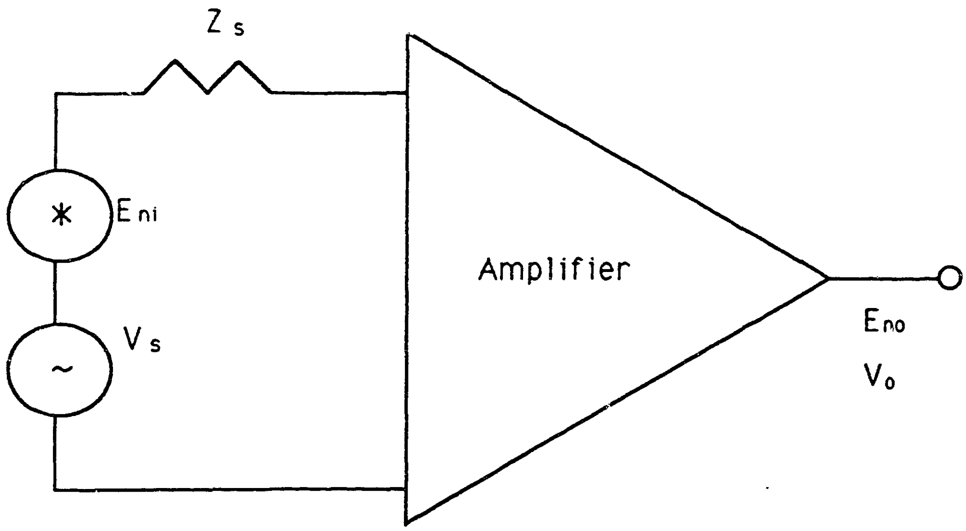

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Figure 2.3.1 Sine Wave Method of Noise Measurement

The gain of the system, K_t, is equal to

$$K_t = \frac{V_o}{V_i} \tag{2.3.1}$$

The gain is measured by inserting the sine wave voltage generator, V_s, in series with the source impedance, Z_s, at the input of the system. The resulting sine wave is measured at the output terminal. The gain is found using equation (2.3.1).

The output noise of the system, E_{no} is measured by removing the signal generator, V_s, and replacing it with a shorting plug. The source impedance, Z_s, is not removed. The noise at the output of the system is measured with an rms voltmeter. Finally, the equivalent input noise is found using the following equation

$$E_{ni} = \frac{E_{no}}{K_t} \tag{2.3.2}$$

The advantage of the sine wave method of measuring noise is that it uses readily available equipment, just a sine wave generator and a rms voltage meter. It is useful, for noise measurements at various frequencies, since the measurement procedure remains the same at all frequencies. The gain of the system at different frequencies is obtained by applying sine waves of different frequencies to the system. This method can be used for low frequency noise measurements. The method may be useful for determining the noise and especially the gain of our noise measurement system.

2.3.2 Noise Generator Method of Noise Measurement

Measurement of the equivalent input noise of a system, will also be used to demonstrate how the noise generator method of noise measurement works. Once again Motchenbacher and Fitchen [2] will be consulted for their description of this measurement method. The input noise measurement procedure is as follows:

1. Measure the total output noise.
2. Insert a calibrated noise signal at the input to increase the output noise voltage by 3 dB.
3. The noise generator signal is now equal to amplifier equivalent input noise. [2]

Figure 2.3.2 shows the block diagram for the noise generator method. E_{ns}, a calibrated noise source is placed in series with a sensor resistor, R_s. E_{ni} represents the equivalent input noise of the system, the quantity that will be found using this procedure. Z_i is the input impedance of the system or amplifier. E_{no} is the output noise of the amplifier and the noise of the generator. An alternative setup for this method is to replace the calibrated noise generator E_{ns} with a high-impedance noise current generator in parallel with the source impedance R_s. Then the equivalent input current noise I_{ni} may

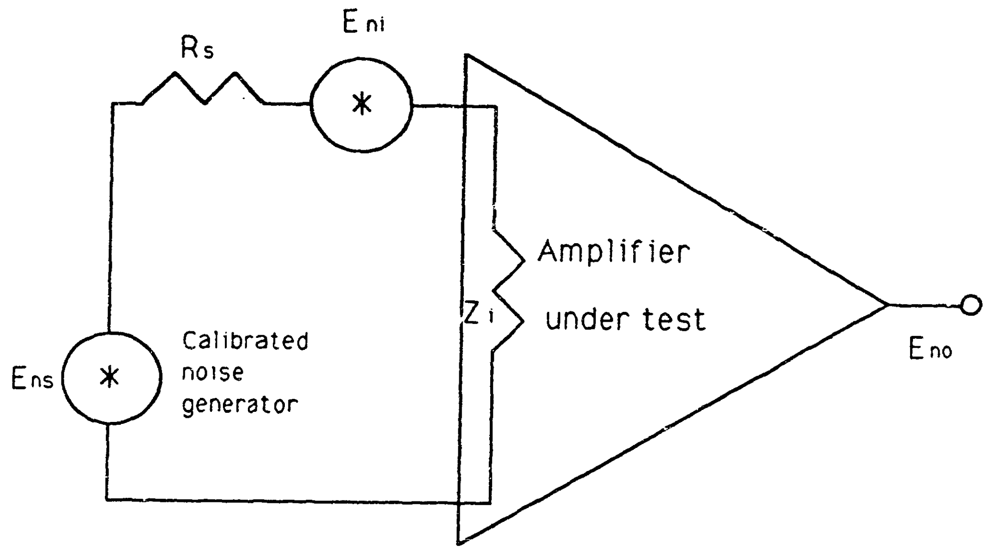

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Figure 2.3.2 Noise Generator Method of Noise Measurement be calculated.

In the noise generator method, the output of the system is measured twice, one time with the noise generator in place and the other with the generator disconnected. The first step in the measurement is to measure the noise at the output of the system with the noise generator disconnected. Then attach the generator to the system. Adjust the output of the generator until the output noise of the system is twice the value it was before, or in other word, 3 dB higher than before. Now measure the output of the noise generator. This value is equal to the equivalent input noise of the system.

The advantage of the noise generator method is the ease of the measurement, just attaching a noise generator to the system and adjusting its value until the output is doubled. Another advantage is that the method is inexpensive, because a low cost diode could be used as the noise generator.

There are disadvantages to this method as well. It is not well suited to low frequency measurements because long measurement times are required. Also pickup of additional noise at the input terminals is more likely to occur in this method, because of the system configuration.

2.3.3 Correlation Method of Noise Measurement

The correlation method of noise measurement is especially useful for measuring very small noise signals. Unlike the other two methods of measurement, this method does not measure or calculate the equivalent input noise or gain of a system. Instead it measures a noise signal. It could be used to measure the output noise of a system, but other methods would have to be employed to find the input noise and gain of the system.

The best description of the correlation method is given by A. van der Ziel [3]. Figure 2.3.3 shows the measurement setup for this method and was taken from A. van der Ziel's book, Noise in Solid State Devices and Circuits. The first step in the measurement is to feed the signal to be measured, V_n, through parallel amplifiers and filters. The amplifiers are represented by A's in the figure and the filters by F's. The resulting signals, V_1 and V_2 are both amplified and filtered. V_1 and V_2 are then put into a crosscorrelator. (See Figure 2.3.3.) The crosscorrelator multiplies the two signals together and its output is the product of the signals, V_1V_2. This signal is then averaged over a certain period of time by the averaging circuit. (See Figure 2.3.3.) The output of the averager is $\overline{V_1V_2}$. When the signal is averaged, the noise of the two amplifiers disappear, since they are uncorrelated. The signal that remains, $\overline{V_1V_2}$,

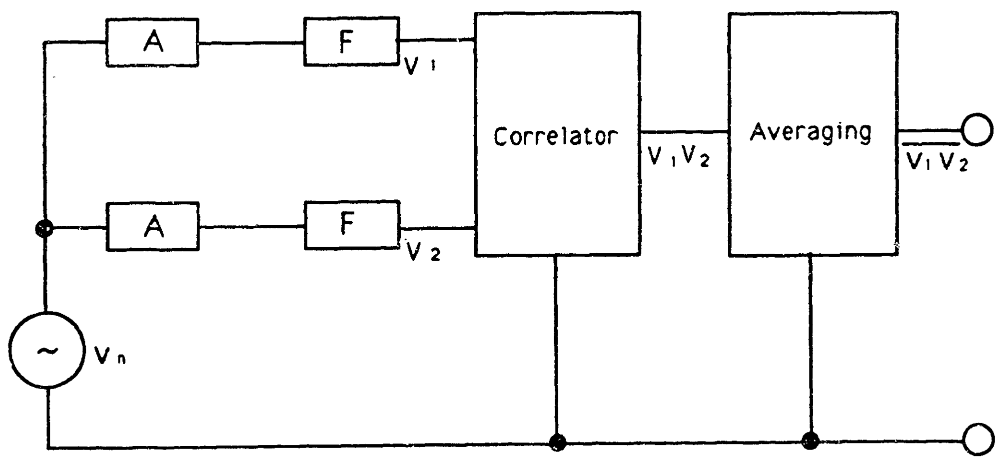

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Figure 2.3.3 Correlation Method for Noise Measurements.
A, Amplifier F, Filter

that remains, V_1V_2, is equal to the signal being measured.

There are several advantages of the correlation method. First, the method allows measurements of very small noise signals. Second, the measurement system for this method is more stable against drift, because the system noise is eliminated from the output noise. For our needs the method may be useful, if the signal we are measuring is small. The method may also be useful, if stability of our measurement system limits the accuracy of our measurements.

2.4 Typical versus State-of-the-Art Noise Measurements

Due to the limited applications, precision noise measurement is not a common area of study. One need for noise measurements is the classification of semiconductor devices. Semiconductor manufacturers often produce low noise components. Applications for these components include systems with low level inputs, audio systems and noise measurement systems. When a manufacturer says a component is low noise, they usually mean that the component is designed to have low noise and that the product sold has a noise level around a value specified on the data sheet. The manufacturers need to perform noise measurements to verify the noise level predicted by design, to determine the average noise level for the component and to screen out any component that does not have the specified noise level.

The noise measurements that manufacturers conduct on their components is what I refer to in this paper as a typical noise measurement. Some manufacturers use special equipment designed to measure noise to perform their measurements. One such device is the Quan-Tech Model 5173 Semiconductor Noise Analyzer. With all available attachments, this device is able to measure noise in a variety of semiconductor devices especially transistors (FETs and bipolars), diodes and operational amplifiers. By inserting a device into the appropriate fixture, a user receives the noise level of the device at five frequency regions (10 Hz, 100 Hz, 1 kHz, 10 kHz and 100 kHz). This particular device is considered state-of-the-art for production screening, but its accuracy is not specified. Resolution is to 3 1/2 digits per reading. The accuracy or more precisely the resolution of the Quan-Tech is not good enough to detect slight differences in the noise levels of similar semiconductors. The ability to detect such a change is necessary for our project.

State-of-the-art noise measurements were conducted by Wieslaw A. Lukaszek in 1974, for his PhD. thesis from the University of Florida [1]. He used noise measurements to investigate the transition from tunneling to avalanche breakdown in silicon p-n junctions. Lukaszek built his own noise measurement system by combining circuitry he built with commercial equipment. We consider his measurements and measurement system state-of-the-art because we know of no diode noise measurements with better accuracy. He characterized the accuracy of his system using resistor noise measurements. The resistance values predicted from noise measurements were within 2% of values obtained from precision bridge measurements.

2.5 Lukaszek's Noise Measurements

For his doctoral research at the University of Florida, Weislaw Lukaszek investigated the transition from tunneling to avalanching in silicon p-n junctions.[1] To study this transition he measured the electric noise produced by the diode (p-n junction). The noise level of the diode reflects the ratio of tunneling to avalanche current within the diode. Lukaszek used V-I measurements to pinpoint the transition between the two types of breakdown as well as to calculate the multiplication factor for avalanche breakdown. The noise measurements he made are the most accurate noise measurements to date.

For our project, we are interested in Lukaszek's noise measurements more than his experimental conclusions for several reasons. First, we are interested in knowing the ratio of tunneling to avalanche current within the diode. Second, the 2% accuracy of his measurements is interesting, because we need to make highly accurate noise measurements. These measurements may have to be better than 2% accurate, but at least by following some of Lukaszek's ideas for measurement we should be able to achieve the 2% degree of accuracy. Improvements in technology in the fifteen years between the measurements may make our measurements more than 2% accurate. Third, since Lukaszek's measurements are also conducted on diodes his work can be used as a reference to see if our system is working as it should.

In this section, Lukaszek's noise measurement system will be described. His measurement procedure will be outlined. In addition, noise ratio and how it can be used to distinguish tunneling from avalanche breakdown will be discussed.

2.5.1 Lukaszek's Noise Measurement System

The system that Lukaszek used to make his noise measurements may be seen in Figure 2.5.1. The General-Radio 1381 Random Noise Generator along with the Hewlett-Packard 350-D attenuator is used to provide a white noise calibrated signal to the system. The 600 Ω resistor, located after the attenuator, is used to provide an impedance match between the attenuator and the rest of the system. The capacitor and resistor in series provide DC and impedance isolation from the rest of the circuit. This isolation is necessary to maintain a constant impedance at the attenuator output terminals regardless of the diode bias network and to convert the noise calibration network into a high impedance current-like source that will not load the diode. The diode bias network is a variable current source. Low noise wire wound resistors, Rb, may be switched in and out of the circuit to provide a range of bias currents to the diode. At the output of the diode is a specially designed preamplifier circuit that utilizes a low noise JFET. After this preamplifier is a selectable gain amplifier which is used to amplify the noise signal so it may be detected by the General-Radio 1925-1926 Real Time Spectrum Analyzer. This instrument consists of 45 third-octave filters, ranging in center frequencies from 3.15 Hz to 8 kHz. The output of each filter is sampled for up to 32 seconds and the rms voltage of each filter, in units of dB, is computed and displayed on the General-Radio 1926 or printed out on the MDS 800 tape printer.

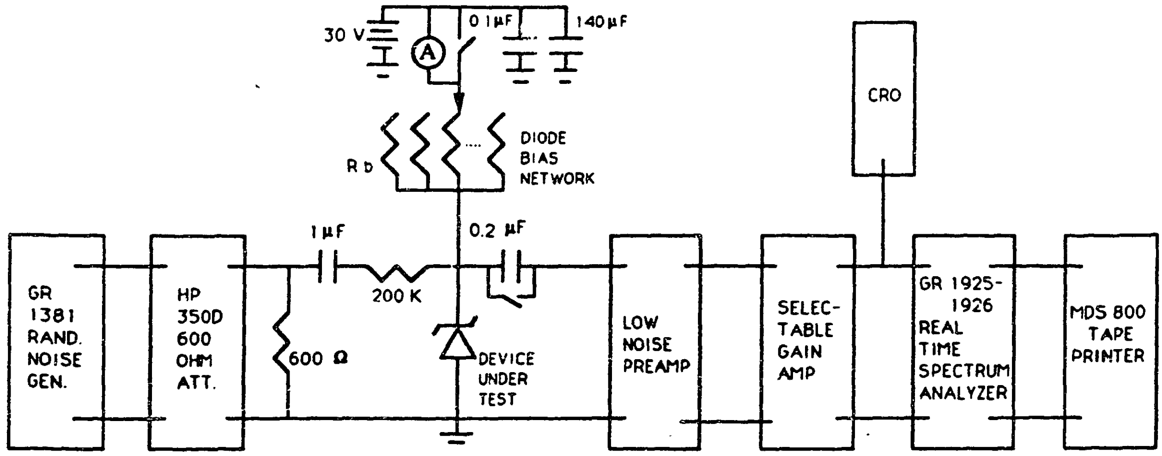

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Figure 2.5.1 Lukaszek's Noise Measurement System

2.5.2 Measurement Procedure

Lukaszek used the following procedure to make his noise measurements. First, he removed the noise calibration signal provided by the General-Radio 1381 and the attenuator from the system by disconnecting the attenuator. He placed a 600 Ω resistor in parallel with the 600 Ω resistor already in the circuit. The diode was biased at a specified reverse current. Then a series of five, 32 second, measurements of the diode noise were made. The next step was to take the second 600 Ω resistor out and reattach the attenuator to the system. The attenuation level was adjusted so that the output noise was 20 dB higher than the diode noise output alone. Another set of five, 32 second, measurements were made. From his noise model for his system and the measurement he made, Lukaszek was able to determine the noise current spectral density, S_{id}, for the diode.

Lukaszek verified the accuracy of his system by measuring resistors in place of diodes. He followed the same procedure as outlined above. Using diodes in the range from 200 Ω to 2 MΩ he was able to predict resistance values from the noise data, that agreed to better than 2% with values obtained by precision bridge measurements.

2.5.3 Noise Ratio

Lukaszek calculated the noise ratio (See Appendix A for definition) for each diode from the noise and reverse current data he measured. The noise ratio indicates whether the breakdown of the p-n junction is caused by tunneling or avalanching. A single step tunneling process has a noise ratio of exactly unity. Multiple step tunneling processes have a noise ratio of less than unity. Noise ratios larger than unity indicate that there is some avalanche breakdown. As one can see from Lukaszek's thesis, noise ratio can be used as an indicator of the transition from tunneling to avalanche breakdown within diodes.

Section 3 DESCRIPTION OF THE NOISE MEASUREMENT SYSTEMS

3.1 The First Measurement System

3.1.1 Block Diagram

The block diagram for the first noise measurement system may be seen in Figure 3.1.1.1. The system has two components, a circuit and

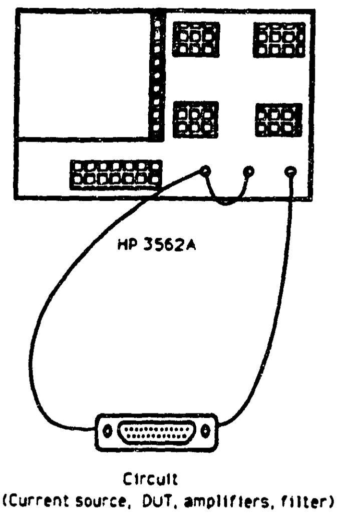


Figure 3.1.1.1 Block Diagram for the First Noise Measurement System

the Hewlett-Packard 3562A Dynamic Signal Analyzer (HP 3562A). The system is not as simple as it appears because each component has several parts and plays several roles in the overall measurement. More detailed descriptions of the components are contained in the following sections. A brief outline of the system is presented here.

The circuit portion of the system has three major parts; a current source, the device under test (DUT), and a noise amplifier. The current source is variable and is used to bias the DUT. The noise amplifiers, as their name suggest, amplify the noise signal produced by the DUT. Amplification is necessary to make the noise signal large enough so that the noise measuring device (in this system the HP 3562A) may detect the signal.

The Hewlett-Packard 3562A Dynamic Signal Analyzer performs several tasks within this measurement system. First, it provides bandlimiting for the system by allowing the user to choose the bandwidth of the noise measurement. Second, the HP 3562A is used in the calibration process for the system. In particular, the HP 3562A supplies a signal to the circuit and makes a measurement of the frequency response of the circuit. The gain of the circuit may be determined from this frequency response. Third, the frequency response measurement capability of the HP 3562A is used to determine the incremental resistance of the DUT. Finally, the HP 3562A is used to measure the noise signal at the output of the circuit portion of the system.

3.1.2 Circuit Description

The circuit portion of the first system consists of a variable current source, a calibration input, a socket for the DUT, three noise amplification stages and a filter. All of these sections are contained in one box. The circuit diagrams for this portion of the

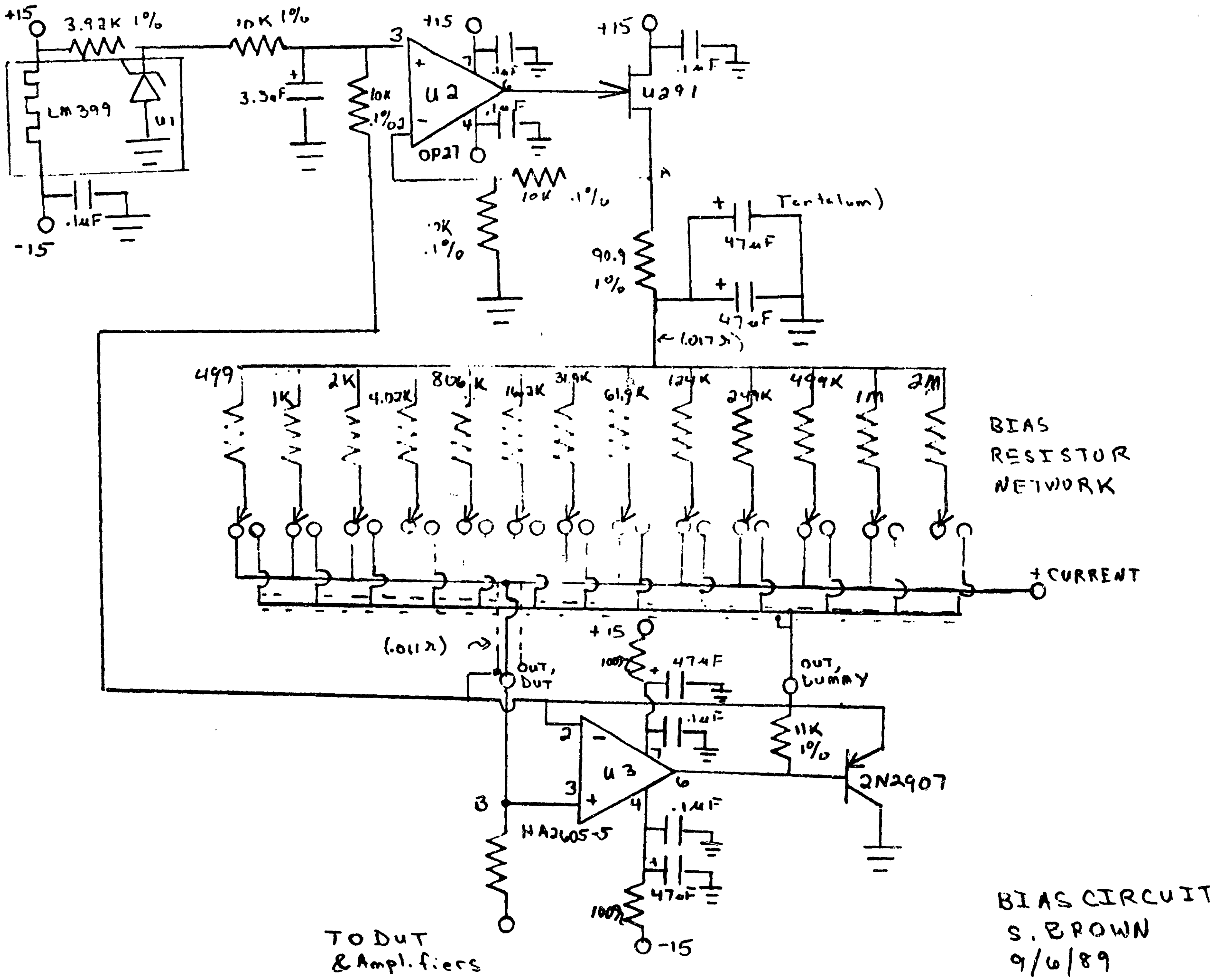


Figure 3.1.2.1 Bias Circuit Diagram

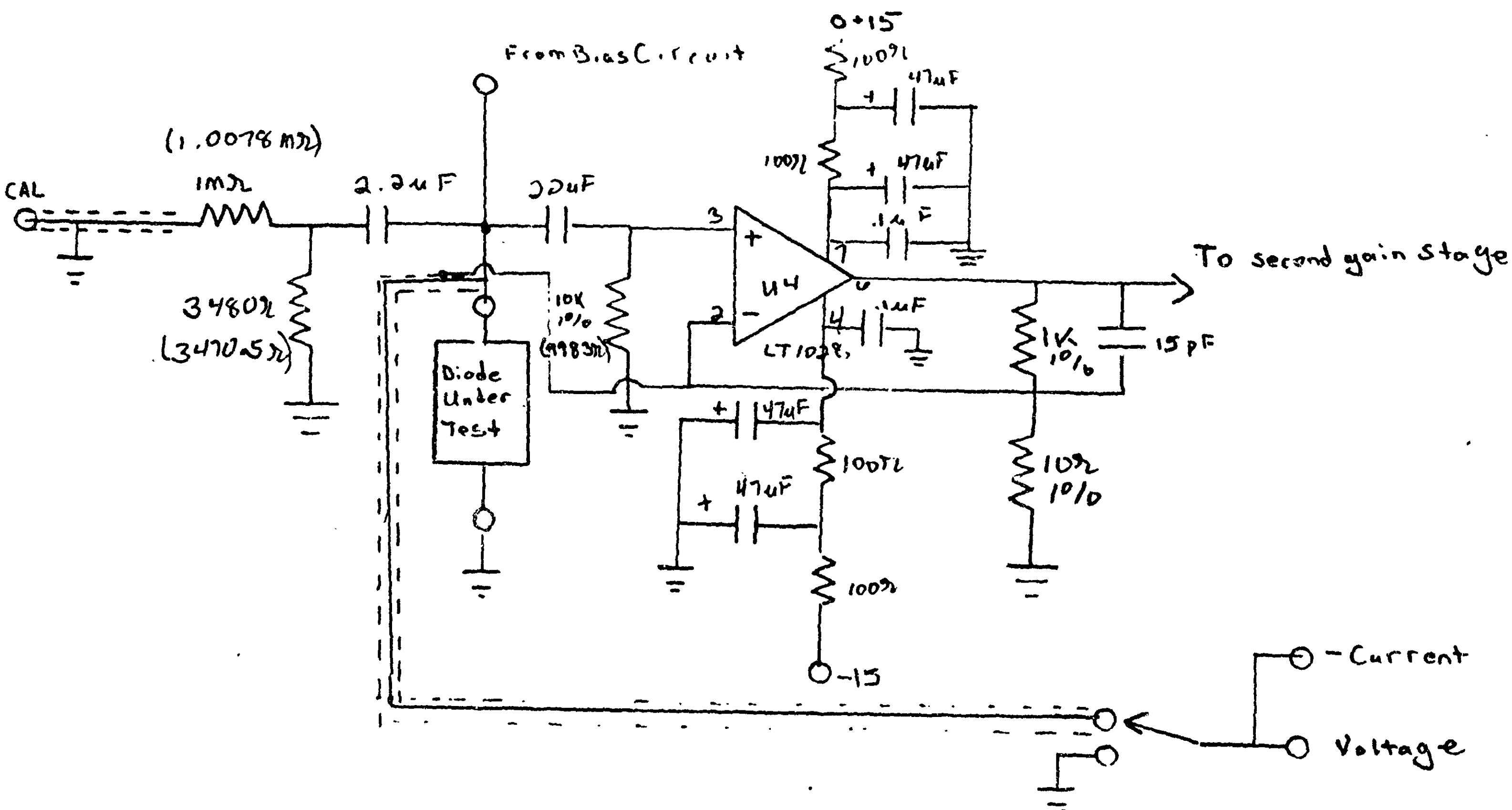


Figure 3.1.2.2 Calibration Input and DUT Socket

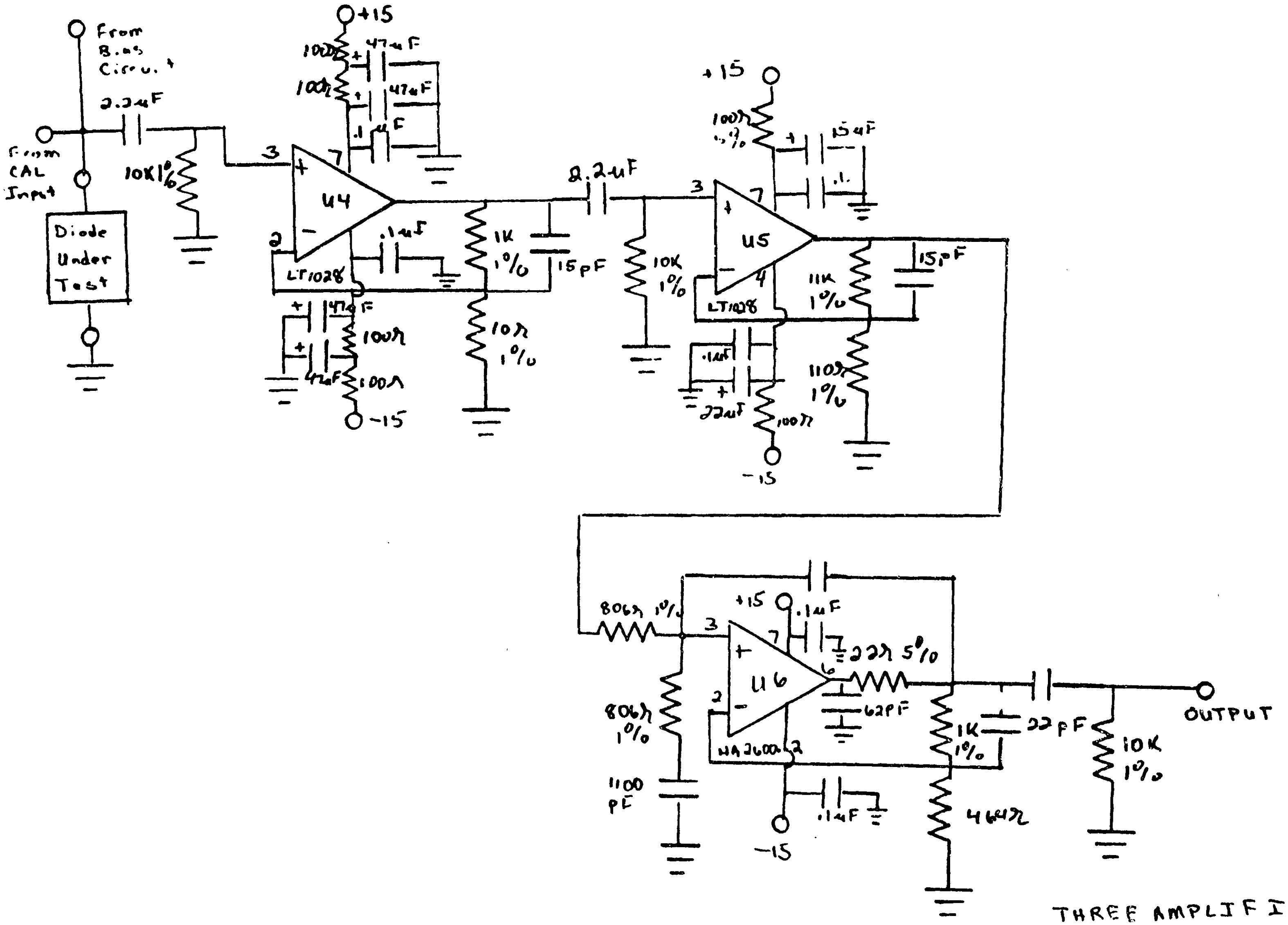


Figure 3.1.2.3 Amplification Stages

THREE AMPLIFICATION STAGES
S. BROWN 9/6/89

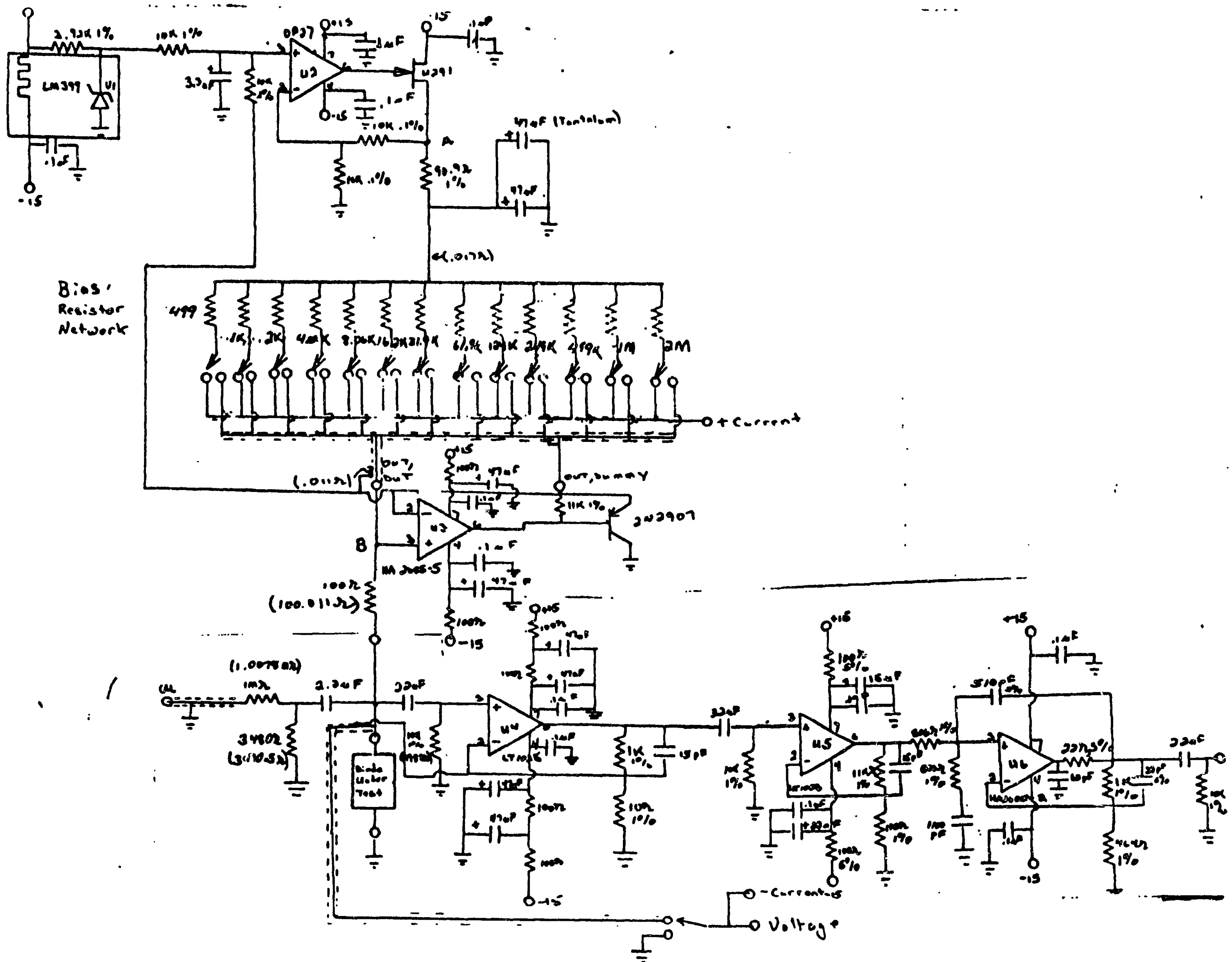


Figure 3.1.2.4 Complete Circuit Diagram

system may be seen in Figures 3.1.2.1 through 3.1.2.4. Figures 3.1.2.1 through 3.1.2.3 show sections of the circuit, while Figure 3.1.2.4 shows the complete circuit. Each section will be described in further detail in the following paragraphs.

The bias circuit was designed to have several features. The circuit has constant power dissipation, independent of bias current. Bias current is insensitive to the voltage of the DUT. The bias current needs a temperature controlled reference voltage for the input in a feedback configuration so it is stable. It is also easy to filter out high frequency noise from this circuit.

The bias circuit may be seen in Figure 3.1.2.1. A National Semiconductor LM399 voltage reference provides a constant voltage of 7 Volts to the bias circuit. The purpose of the complex circuitry that follows the voltage reference is to keep the voltage from the output of the U2 op amp to the noninverting input of the U3 op amp (From points A to B on Figure 3.1.2.1) constant. In other words, the voltage across the bias resistor network is kept constant. Resistor values within this network range in from 499 Ω to 2 MΩ. These resistors may be switched into the circuit to produce a variety of bias currents for the DUT. The selected bias current then flows through a 100 Ω resistor used to measure bias current and into the DUT itself.

The next two parts of the circuit may be seen in Figure 3.1.2.2. They are the cal(ibration) input and the DUT socket. The CAL input as its name implies is the point in the circuit where a calibration signal is applied. The signal is attenuated and filtered before it passes into the amplification stages at the top of the DUT. Since diodes are the type of devices being measured in this system, the DUT socket is one that accommodates such a device. The cathode of the DUT is attached to the 100Ω resistor and the anode is attached to ground. The DUT is reversed biased.

There are three noise amplification stages within the circuit. When designing a noise amplification stage one must design a low noise amplifier because you do not want the noise of the amplifier to swamp the noise signal you are amplifying. To keep the noise of the amplifier down, one can use low noise operational amplifiers, which have low input noise voltages and currents. These low noise operational amplifiers are especially critical in the first amplification stage where the input signal is very small. In this circuit, the Linear Technology LT1028 low noise operational amplifier is used in the first and second amplification stages.

The three amplification stages may be seen in Figure 3.1.2.3. The input of the first amplification stage is the sum of the noise of the DUT and any calibration signal. The first stage has a gain of approximately 101. The second stage also has a gain of approximately 101. The last stage is not only an amplifier, but also a filter. The gain of this stage is 3.16. The filter is a second order low pass filter with a cutoff frequency at about 263.6 kHz. The amplified noise signal passes though a simple high pass filter with a cutoff at 7 Hz. This filter eliminates any DC signal. Then the amplifier signal goes to the output terminal of the circuit.

3.1.3 Commercial Equipment

There is only one piece of commercial equipment used in this measurement system. This is the Hewlett-Packard 3562A (HP 3562A). This piece of equipment is used to calibrate the system, measure the gain of the system, and measure the noise of the system. From the range of functions the HP 3562A performs, it can be seen that this is a very versatile piece of equipment. In this section, the measurement procedures for frequency response and power spectrum measurements with the HP 3562A will be described in detail. Pertinent specification for the HP 3562A will also be presented.

The HP 3562A is capable of making a frequency response measurement on a system. This measurement, sometimes called a transfer function, is the ratio of the system's output to input. From this measurement, the gain and phase shift of the system may be determined.

The basic setup for the frequency response measurement may be seen in Figure 3.1.3.1. The source output terminal of the HP 3562A is

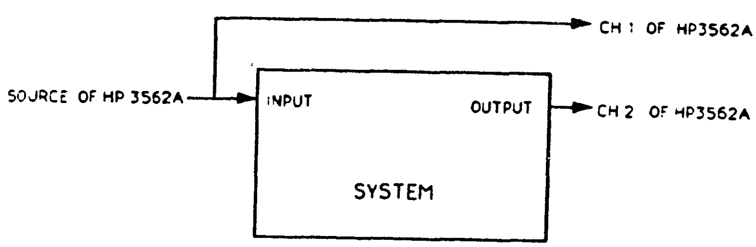


Figure 3.1.3.1 Setup for Frequency Response Measurement

attached to its own channel one and the input of the system. The output of the system is attached to channel two of the HP 3562A.

The HP 3562A offers four types of measurement modes; linear resolution, log resolution, swept sine, and time capture. Frequency response measurements may be made with the first three modes. For our measurement purposes, we conduct frequency response measurements in only the linear resolution and swept sine modes.

In the linear resolution mode, "time data is sampled until a data buffer called the 'time record' is filled with a fixed number of time samples. Once a time record is filled, the fast Fourier transform of the record is computed and the frequency spectrum is displayed." [9:9] In this mode each channel has 801 lines of frequency resolution. The resolution ranges from 125 Hz for a full (100 kHz) frequency span to 12.8 μHz for the smallest (10.24 mHz) frequency span.

In the linear resolution frequency response, a signal is applied by the HP 3562A through its source terminal to the input of the system. There are five types of source signals to select from; random noise, burst random noise, periodic chirp, burst chirp, and fixed sine. The random noise and fixed sine are the most common selections. The source level may also be selected. The range of levels depend on the type of signal being applied.

The averaging capabilities of the HP 3562A may also be used with this measurement. There are four types of averaging available; stable (mean), exponential, peak hold, and continuous peak. We use only the stable (mean) type of averaging. Any number of averages between 1 and 32,767 samples may be selected . The HP 3562A makes a number of measurements equal to the number of averages selected. It averages these measurements and displays the average value.

The frequency span of the measurement may also be selected. The frequency span of the HP 3562A is from 0 Hz to 100 kHz. Various smaller frequency spans may be selected. If the frequency span of the system being measured is known, it should be used as the measurement's frequency span. If the frequency span of the system is unknown, it is best to use the full 100 kHz span.

Once the HP 3562A completes the frequency response measurement, it will display the magnitude (gain) versus frequency on the screen. The scale and units of the display may be changed. The cursors and special marker capabilities of the HP 3562A may be used to determine certain values, like the gain at certain frequencies. If the HP 3562A is attached to a plotter a copy of the screen may be produced.

Frequency response measurements are also made with the swept sine mode. In this mode, "the HP 3562A is reconfigured as a full-function DC to 100 kHz frequency response analyzer." [9:15] This type of product is traditionally used in low frequency network analysis. These products perform the same measurement as a tuned network analyzer, but instead of using low frequency filters they "perform a time domain integration of the input signals to mathematically filter signals at very low frequencies. Measurement results are usually displayed as point-by-point numerical values or on an x-y plotter." [9:15]

In a swept sine frequency response, the source that is applied to the system is a sine wave with a fixed amplitude, that the operator selects, and a varying frequency. The initial frequency of the sine wave is called the start frequency. The frequency of the wave changes at a certain rate called the sweep rate. The final frequency is equal to the start frequency plus the frequency span. The start frequency, sweep rate and frequency span may all be selected by the operator. The operator also has to choose between a linear or log sweep. The difference between the two is that the frequency is either linear or logarithmic. The frequency response for each frequency within the span is calculated and displayed on the screen. The frequency response is drawn on the screen a point at a time.

The swept sine measurement is like the linear resolution method in several ways. Averaging can be utilized, where in this mode measurements at a single frequency are averaged and then displayed. The scale and units of the measurement can be easily changed with a touch of a button. The cursors and special markers may also be used with the measured waveform.

The other type of measurement made with the HP 3562A is the power spectrum measurement. This measurement may be make in the linear resolution and log resolution modes. In our system we only make the measurement in the linear resolution mode. The power spectrum measurement displays the input signal in the frequency domain. It is computed by taking the FFT of the input signal and multiplying it by complex conjugate of the FFT.

To make this measurement, attach the signal to be measured to either one of the HP 3562A channels. Then check to make sure the channel is activated. Display the power spectrum of the channel on the screen. Select the frequency span of the measurement. Decide if averaging is desired and if so select the number of averages to be made. Start the measurement.

Once the measurement is completed, the power spectrum is displayed on the screen. The units and scale can be changed so the desired spectrum is displayed. The cursor and special markers can be utilized to record more detailed data. Since the power spectrum measurement is used to measure the noise of the system, the units of $Volts^2/Hertz$ is selected. If these units are used the displayed waveform is equal to the spectral density of the noise signal.

This section has just briefly described two types of measurements made with the HP 3562A. Along with these descriptions, some of the specifications of the HP 3562A have been given. A complete listing of the specification of this device may be seen in Appendix D. The specifications will be discussed again in section 4, when I evaluate the accuracy of measurements made with this equipment.

3.1.4 Measurement Procedure

A calibration procedure is part of the overall measurement procedure. The calibration procedure measures the gain of the system as well as the noise of the system for three calibration resistors (10, 100 and 1,000 ohms). This data is then fed into a computer program which returns several constants. The constants are ultimately used to determine the dynamic resistance and the noise ratio associated with the DUT.

The circuit and the HP 3562A are the components involved in the calibration procedure. The capabilities of the HP 3562A are utilized in this calibration procedure. In particular, the HP 3562A is used to make frequency response and noise (power spectrum) measurements. The mathematical functions as well as source capabilities are used in conjunction with these measurements.

The first step in the calibration procedure is to measure the frequency response correction waveform. This correction waveform is necessary for measurements made outside the original measurement bandwidth, a bandwidth in which the frequency response of the system was adjusted to be flat. It was discovered that some of the devices being measured by the system displayed 1/f noise above 5 kHz. To measure the noise of such devices, without added 1/f noise, it was necessary to use other bandwidths. A technique, which employs a correction waveform, allows the frequency span to be moved without modifying the hardware of the system.

The setup for the measurement of the correction waveform may be seen in Figure 3.1.4.1. The source of the HP 3562A is attached to the

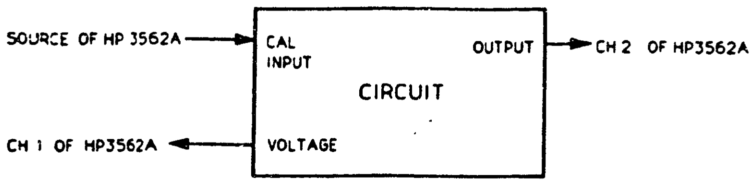


Figure 3.1.4.1 Setup for Correction Waveform Measurement

CAL input of the circuit. The VOLTAGE terminal of the circuit is attached to channel 1 of the HP 3562A, while the OUT terminal of the circuit is attached to channel 2.

The 1 kΩ calibration resistor is used as the DUT in the correction waveform measurements. The 16 kΩ bias resistor is put into place, producing a bias voltage of around 0.4 mV. The HP 3562A is set for a sweep sine frequency response measurement using state 3, which may be seen in Figure 3.1.4.2. The frequency range in this state must be adjusted so it corresponds to the range which will be used in future noise measurements. A series of ten sweep sine measurement, with HP 3562A calibrations between each measurement, is taken. The

```
                    Swept Sine

AVERAGE:  INTGRT TIME              # AVGS
          50.0mS                        1

FREQ:
  START   5 kHz                    SPAN      50.0kHz
  STOP    55 kHz                   RESLTN    31.2 Hz

SWEEP:    TYPE       DIR           EST TIME  EST RATE
          Linear     Up            12.1 Min  68.7 Hz/S

AU GAIN:  Off

INPUT:    RANGE      ENG UNITS     COUPLING
  CH 1    3.99mVpk   1.0 V/EU      AC (Flt)
  CH 2    11.2 Vpk   1.0 V/EU      AC (Flt)

SOURCE:   TYPE                     LEVEL     OFFSET
          Off                      450mVpk   0.0 Vpk
```

Figure 3.1.4.2 State 3, Used for Frequency Response Measurement

purpose is to average calibration effects. This series of measurements is easily made by using the auto sequence titled "Start w/Cal", which may be seen in Figure 3.1.4.3. When all

```
                 Auto Sequence 4 174 Keys Left
Display ON       Label: START W/CAL
  1   START
  2   SAVE RECALL: SAVE DATA# 2
  3   CAL: SINGLE CAL
  4   ASEQ FCTNS: TIMED PAUSE 0 Sec
  5   START
  6   MATH: ADD: SAVED 2
  7   ASEQ FCTNS: LOOP TO 2, 8
  8   MATH: DIV 10
```

Figure 3.1.4.3 Autosequence "Start w/Cal" ten measurements are displayed the frequency response is displayed on the screen of the HP 3562A. The frequency response is manipulated to obtain the absolute magnitude squared gain. The HP 3562A has two display traces, A and B. A measurement may be displayed in either of both of these displays. The manipulation of the frequency response utilizes the dual trace capability of the HP 3562A. First, the frequency response in both trace A and B. Trace A is complex conjugated and multiplied by trace B. The result is the absolute magnitude squared gain of the system for a particular bandwidth. This result is stored in the HP 3562A's memory within Data 1. All future noise measurements made within this frequency range should be divided by this correction waveform. By doing this, the noise data is corrected for any effects of the noise amplifier outside its flat region and is referred back to the DUT.

The next step in the calibration procedure is to measure the gain of the noise amplifier from the CAL input to the OUT terminal at a fixed frequency of 5.2 kHz. The setup for this measurement may be seen in Figure 3.1.4.4. The source of the HP 3562A is attached to its

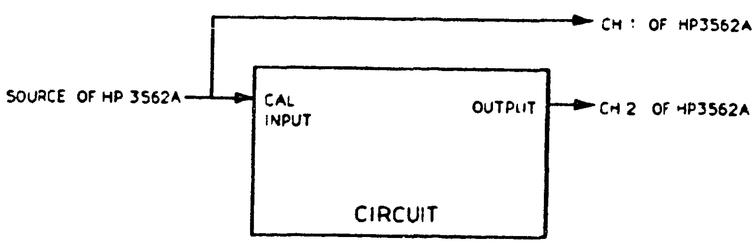


Figure 3.1.4.4 Setup for Measuring Gain at a Fixed Frequency

own channel 1 and to the CAL input of the circuit. The OUTPUT terminal is attached to channel 2 of the HP 3562A.

The 1 kΩ calibration resistor remains as the DUT. The source level of the input sine wave is adjusted so that the sine wave at the output of the circuit is around 1.6 V. This is just a conventional value and does not have to be exact. The HP 3562A is set up for a linear resolution frequency response measurement using state 2, which may be seen in Figure 3.1.4.5. Using the auto sequence "Start w/Cal" a series of ten

Linear Resolution

MEASURE:	CHAN 1 Freq Resp		CHAN 2 Freq Resp	
WINDOW:	CHAN 1 Hanning		CHAN 2 Hanning	
AVERAGE:	TYPE Stable	# AVGS 10	OVERLAP 0%	TIME AVG Off
FREQ:	CENTER 5.2 kHz		SPAN 200 Hz	BW 375mHz
	REC LGTH 4.0 S	Δt 3.91mS		
TRIGGER:	TYPE Freerun	LEVEL 0.0 Vpk	SLOPE Pos	PREVIEW Off
INPUT: CH 1 CH 2	RANGE AutoRng AutoRng	ENG UNITS 1.0 V/EU 1.0 V/EU	COUPLING AC (Gnd) AC (Gnd)	DELAY 0.0 S 0.0 S
SOURCE:	TYPE Fxd Sin	FREQ 5.2kHz	LEVEL 2.25 Vpk	OFFSET 0.0 Vpk

Figure 3.1.4.5 State 2, Used to Measure the Gain at a Fixed Frequency

measurements with calibrations in between are taken and averaged. This average frequency response should be displayed on the screen and the value at 5.2 kΩ should be recorded. This is the gain from the CAL input to the OUT terminal.

The third step in the calibration procedure is to measure the noise of the 1 kΩ calibration resistor using the HP 3562A. The system may be left in the frequency response setup that was seen in Figure 3.1.4.4. The source is not used in this case and only channel 2 is activated. The HP 3562A is set up for this noise (power spectrum) measurement using state 1, which may be seen in Figure 3.1.4.6. The

Linear Resolution

MEASURE:	CHAN 1 Off		CHAN 2 Power Spec	
WINDOW:	CHAN 1 Hanning		CHAN 2 Hanning	
AVERAGE:	TYPE Stable	# AVGS 1000	OVERLAP 0%	TIME AVG Off
FREQ:	CENTER 90 kHz		SPAN 20.0kHz	BW 37.5 Hz
	REC LGTH 40.0mS	Δt 39.1μS		
TRIGGER:	TYPE Freerun	LEVEL 0.0 Vpk	SLOPE Pos	PREVIEW Off
INPUT: CH 1 CH 2	RANGE AutoRng AutoRng	ENG UNITS 1.0 V/EU 1.0 V/EU	COUPLING AC (Gnd) AC (Gnd)	DELAY 0.0 S 0.0 S
SOURCE:	TYPE Off		LEVEL 5.0 Vpk	OFFSET 0.0 Vpk

Figure 3.1.4.6 State 1, Used to Measure Noise

frequency range must be moved to the appropriate frequency bandwidth. The autosequence "Start w/Cal" is used with this measurement setup so a series of ten measurements with calibrations in between are made and averaged. The result is divided by the correction waveform stored in the HP 3562A memory. The special marker "ave value" is pressed to display the average value of the noise for the 1 kΩ calibration resistor. This value is in units of Volts^2/Hertz.

The fourth step in the calibration is to estimate the resistance of the 1 kΩ calibration resistor. This is done by measuring the voltage across the resistor and the current through it with a meter. We use a Fluke 8506A digital volt meter for this job. The voltage is measured at the voltage terminal of the circuit and the current is measured at the current terminal of the circuit. The resistance of the 1 kΩ resistor is estimated using the following formula.

$$R_{DUT} = \frac{V_{VOLTAGE} * R_S}{V_{CURRENT}} \qquad (3.1.4.1)$$

R_s is the sum of the resistor which the current is measured across, which has a resistance of 100 Ω, and the resistance of the wire, which has a resistance of 0.011Ω.

Steps two through four of the calibration procedure are repeated with a 10 Ω and a 100 Ω calibration resistor used as the DUT. After these measurements are complete, a four terminal resistance measurement is made with the Fluke 8506A on all calibration resistors. The gain and noise data obtained from the three calibrations along with the corresponding calibration resistance are fed into the program cal.bas. A copy of this program may be seen in Appendix C. The program solves three linear equations for the calibration constants, C_r, G and K_r. The program also solves three linear equations for the calibration constants A, B, and C. These six constants are used with noise measurement data obtained for diode DUTs to determine the dynamic resistance and noise ratio of the DUTs.

Once a calibration is made, noise measurements on various DUTs can be conducted. The data from the calibration is used to solve for noise ratio and other parameters. Another calibration is not necessary for several months or until something is changed in the circuit. If changes are made to the circuit a calibration should be performed before the system is used again for noise measurements.

The measurement procedure used with this system begins with the installation of a DUT. The DUT is usually a diode. The DUT is biased at a certain current by switching in bias resistors until the desired current is reached. The bias voltage is measured with the Fluke 8506A at the VOLTAGE terminal of the circuit. The bias current is measured at the CURRENT terminal of the circuit. Actually the voltage across a 100 Ω resistor is measured. The bias current is obtained by dividing this voltage by 100 Ω. The bias voltage and current data will be used later in the project to determine if there is correlation between noise and radiation characteristics of the diodes.

The next step in the measurement procedure is to measure the gain from the CAL input to the OUT terminal. This measurement is made at a fixed frequency of 5.2 kHz using state 2 (see Figure 3.1.4.5). This fixed gain is recorded and used along with the three calibration constants, G, K_r, and C_r, to find the admittance, G_d, which is equal to the DUT resistance in parallel with the bias resistors. G_d is found using the following equation.

$$G_d = \frac{K_r}{Gain - C_r} - G \qquad (3.1.4.2)$$

If the resistance is desired, it may be easily obtained by inverting this admittance ($R_d = 1/G_d$).

The last measurement made in this procedure is a measurement of the output noise of the circuit with the HP 3562A. A measurement of the noise spectral density, S_v, is made by performing a power spectrum measurement using state 1 (see Figure 3.1.4.6) on the output of the circuit. S_v is a noise voltage spectral density in units of $Volts^2$/Hertz. It contains a noise contribution of the circuit and the DUT. Dividing the spectral density by the correction waveform refers the noise back to the DUT.

Once the three measurement steps have been completed there is enough information to calculate the noise ratio of the DUT. See Appendix A for a definition of and formula for noise ratio. To calculate noise ratio the noise voltage spectral density referred to the DUT, S_v, will have to be converted into a current spectral density, consisting of only the DUT noise, S_{id}.

The first step in this conversion is to subtract away the noise contributed by the circuit. The noise contributed by the circuit is dependent upon the impedance seen at the input of the noise amplification stage of the circuit. From the calibration process three constants, A, B, and C were found. These constants are the coefficients of an equation that can be used to predict the noise at the DUT node produced by the system. In the case of a diode DUT, the impedance seen at the input of the noise amplifier is the resistance, R_t. R_t is described by the following equation.

$$R_t = (1/R_d + G)^{-1} \qquad (3.1.4.3)$$

Substitute the value for R_t into the following equation

$$e_o^2 = A + B * R_t + C * R_t^2 \qquad (3.1.4.4)$$

allows one to predict the noise of the system, e_o. This output noise is subtracted from S_v.

Recall that the noise measured was a voltage spectral density. For noise ratio calculations, one needs a current spectral density. To obtain the current spectral density just divide by the resistance seen at the DUT node, R_t, squared.

When the predicted noise e_o^2 was subtracted from S_v, e_o^2 included noise contributed by the incremental resistance of the diode. But this resistor is, by convention, modeled as noise free. So, we follow convention by adding a thermal noise current spectral density equal to

$$i_t^2 = 4kT/R_d \qquad (3.1.4.5)$$

to S_{id}.

Thus the overall conversion of S_v to S_{id} may be summarized with the following formula.

$$S_{id} = \frac{S_v - e_o^2}{R_t^2} + i_t^2 \qquad (3.1.4.6)$$

Now with the spectral density, S_{id}, and the bias current, I_r, known, noise ratio may be found using equation (A.6) from Appendix A.

3.2 The Second Measurement System

3.2.1 Block Diagram

The block diagram for the second noise measurement system may be seen in Figure 3.2.1.1. This system has five components; a circuit, a

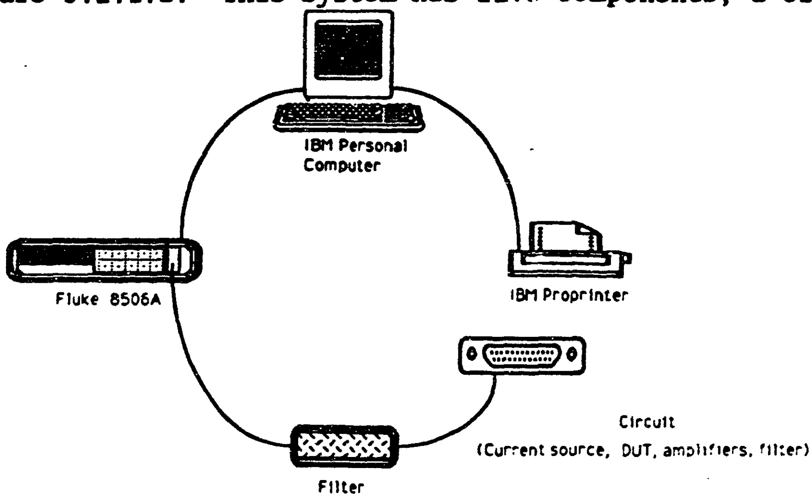


Figure 3.2.1.1 Block Diagram for the Second Noise Measurement System

filter, the Fluke 8506A Thermal RMS Multimeter (Fluke 8506A), the IBM Personal Computer (IBM PC) and the IBM Proprinter (printer). Once again, more detailed descriptions of the components are contained in the following sections. The more generalized description appears in this section.

The noise amplifier circuit used in this system is the same one that is used in the first measurement system. It has four parts; a current source, the DUT, noise amplifiers, and a filter. This measurement system contains an additional filter because the Fluke 8506A does not have built in bandlimiting capabilities and has a very wide measuring bandwidth in the order of megahertz. The filter provides the bandlimiting that is necessary for obtaining noise measurements with the Fluke 8506A that can with accuracy be related to power spectral density.

The Fluke 8506A is used in this system to measure the noise at the output of the filter. The Fluke 8506A was selected to measure the noise because of its capability to measure rms values of random signals very accurately. An additional feature allows for measurements made with the Fluke 8506A to be controlled by a computer program.

The IBM PC is used in this system to run a BASIC program, I wrote, which controls the noise measurements made by the Fluke 8506A. The IBM PC also stores the measurement data for the system. The printer is used to produce a paper copy of the measurement data for the system.

3.2.2 Circuit Description

There are two circuits that are used in this second noise measurement system. The first is the same circuit that was used with the first measurement system. It contains a bias current source and noise amplifiers as well as the socket for the DUT. This circuit has been described in detail in section 3.1.2. The second circuit used in this system is the filter. The filter is necessary to bandlimit the noise signal so that the noise may be measured accurately by the Fluke 8506A, which does not have bandlimiting capabilities. The filter will be described in detail in the following paragraphs.

The filter used in this measurement system was designed as an anti-aliasing filter for a digital system designed, to measure noise, by Lenny Sheet for his combined bachelor and master's thesis from M.I.T. The filter was designed with noise measurement in mind, so it has a low frequency cutoff of 1 kHz. Since most devices that we will measure will not have 1/f noise above 1 kHz, the filter will prevent the measurement of 1/f noise. The filter has a bandwidth of 29 kHz from 1 kHz to 30 kHz. Some type of bandlimiting is required for noise measurements, but as long as the filter cuts off the 1/f noise, the selection of the high frequency cutoff and bandwidth is arbitrary. With this consideration in mind, the filter described here will be fine for our application.

The filter is a fifth order Chebyshev filter. Since it is of a high order it has sharp frequency rolloffs. As noted above it has a bandwidth of 29 kHz, form 1 kHz to 30 kHz. It has a gain of approximately -14 dB with a 1 dB ripple in the passband. A frequency response of the filter, taken with the HP 3562A, may be seen in Figure 3.2.2.1.

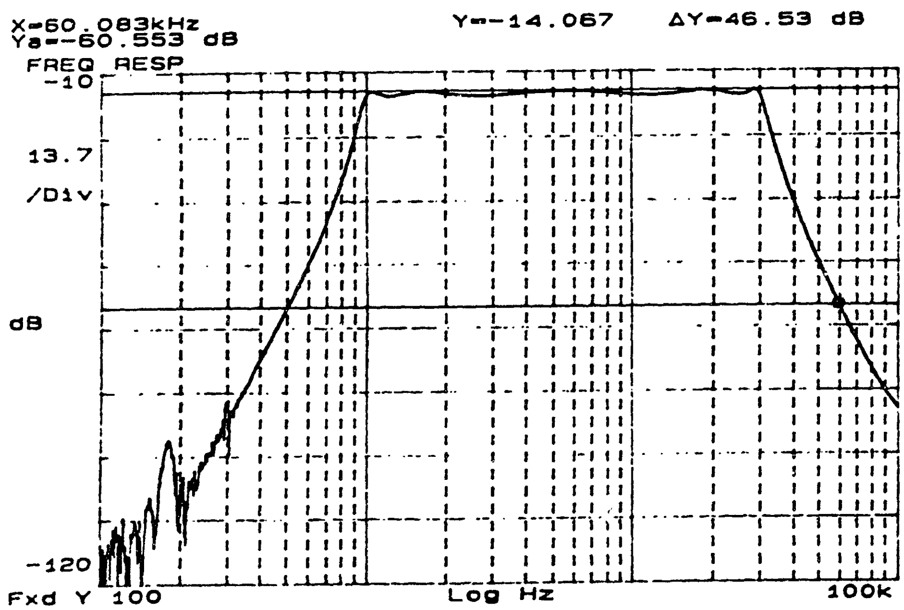


Figure 3.2.2.1 Frequency Response of Filter

The filter consists of five stages all of the same basic design. This design may be seen in Figure 3.2.2.2. Table 1 gives the exact

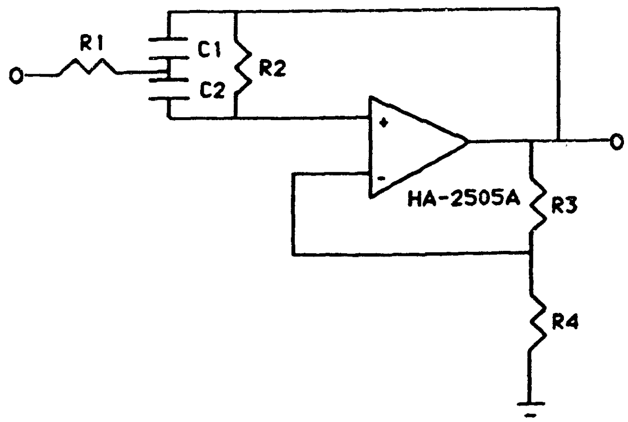


Figure 3.2.2.2 Circuit Diagram for Filter Stage

resistor and capacitor values for the components in each of the five

Table 1 Component Values for the Five Filter Stages

Stage	R1 (kΩ)	R2 (kΩ)	R3 (kΩ)	R4 (kΩ)	C1 (nF)	C2 (nF)	C3 (pF)
1	5.110	60.400	10.000	1.180	0.301	0.301	47.000
2	4.688	52.300	10.000	1.400	10.100	10.100	47.000
3	4.546	14.000	10.000	2.890	0.983	0.983	47.000
4	6.049	19.109	10.000	2.808	9.910	9.910	47.000
5	2.543	3.320	10.000	1.922	10.000	10.000	47.000

stages. The filter is built so that the signal applied to the filter passes through stage five first, proceeds through the other stages in descending numerical order, exiting from stage one.

When the second measurement system was first assembled and tested, some of the smaller signals from the circuit were so attenuated by the filter that they could not be distinguished from the noise of filter. To eliminate this problem, it was necessary to add another gain stage to the system. Since there were several sockets available in the box containing the filter, the gain stage was placed in the box, preceding the filter.

The circuit diagram for this gain stage may be seen in Figure 3.2.2.3. The circuit was designed so that a wide range of signal

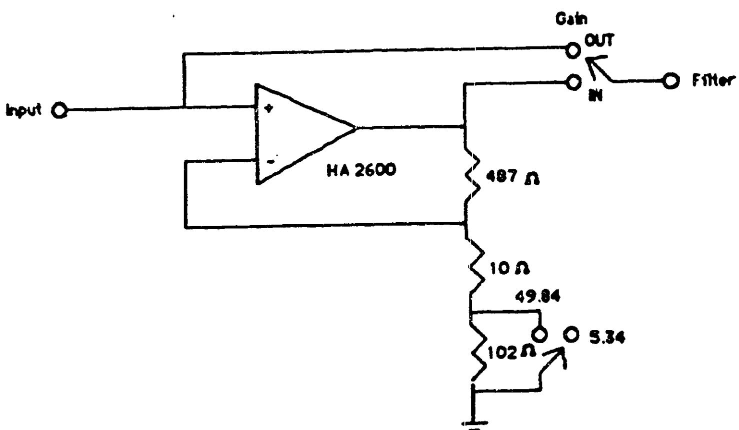


Figure 3.2.2.3 Circuit Diagram of the First Gain Stage

amplitudes could be accommodated. When the amplitude of the signal from the circuit is high the gain stage may be switched out of the circuit. This prevents saturation of the operational amplifiers and allows the signal from the circuit to go directly into the filter. For lower amplitudes, the gain stage can be switched into the circuit. The gains of 5.34 and 47.84 were chosen by looking at the lowest amplitude that was measured by the first system and the noise of the filter with no input. The criteria for choosing the gain was to amplify the smallest signal so that it was sufficiently above the noise of the filter. Thus the gain of approximately 50 was selected. The gain of approximately 5 was selected because it was a factor of ten below the other gain. It can be used for the intermediate amplitudes that need some amplification to be measured precisely, but not as much as the very low signals.

After further testing of the filter circuit, a problem with capacitive loading was found. To eliminate this problem, another gain stage was added at the output of the filter to act as a buffer. A circuit diagram for this gain stage may be seen in Figure 3.3.2.4. The gain of 6 was selected to boost the signal to approximately the same level it was at before going through the filter, negating the one-sixth attenuation of the filter. The gain of 18 was selected to allow further amplification of very small signals.

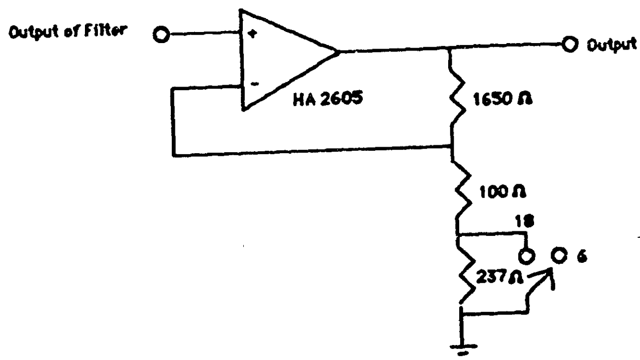


Figure 3.2.2.4 Circuit Diagram of the Second Gain Stage

3.2.3 Commercial Equipment

The Fluke 8506A Thermal RMS Digital Multimeter (Fluke 8506A) along with the IBM Personal Computer (IBM PC) and Proprinter (printer) are the commercial equipment used in the second measurement system. The Fluke 8506A is used to measure the noise output of the system. The Fluke is a thermal rms meter. It measures waveforms by measuring the heat produced when they pass though a resistor. So it is insensitive to waveform shape. The Fluke 8506A measures the AC signal in units of root-mean-square (rms) Volts. Since the conventional noise units are rms Volts per root Hertz, the signal measured by the Fluke can be easily converted to these units by dividing by the square root of the bandwidth of the measurement system.

The Fluke 8506A makes its rms measurements in a rather different manner than most thermal rms meters. Most meters measure the AC signal by applying the AC signal being measured and a DC voltage to a rms sensor until the two signals are equal. Then the meter displays the DC voltage that was found to be equal to the AC signal. The Fluke 8506A uses a thermal rms sensor and differs from other meters because it does not adjust the equivalent DC signal to be as close to balance as possible. Instead it starts the comparison with a DC signal close to the actual AC signal, by taking the DC equivalent of the sensor's first output as the first approximation to the AC signal, and then making one iteration only.

Figure 3.2.3.1 is taken from the Fluke 8506A Instruction Manual [8] and shows how the rms AC voltage is calculated by the meter. X1

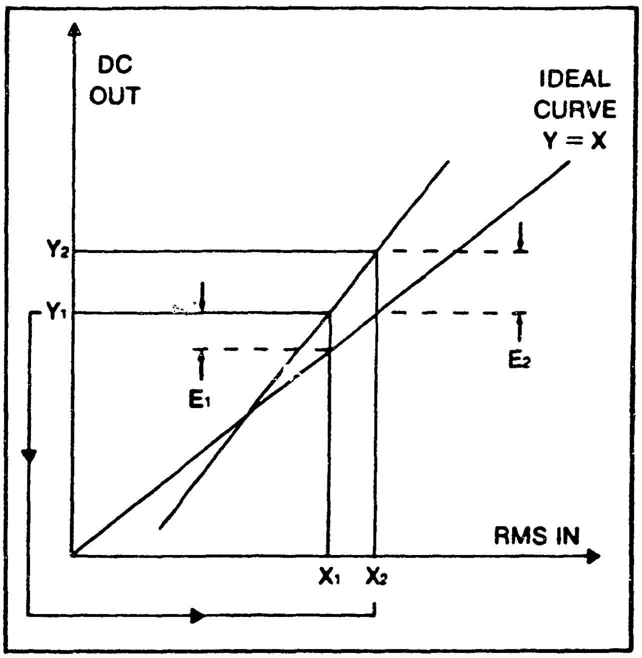


Figure 3.2.3.1 Fluke 8506A Calculation of an AC Signal

is the unknown AC signal and Y1 is the sensor's first output. The difference between the two signals, the error, is represented by E1. The DC signal applied to the sensor for comparison is X2 and is equal to Y1 the sensor's first output. This is the first approximation as mentioned above. The new (second) output of the sensor is Y2. The difference between Y2 and X2 is E2. From the figure and this description one can see that

$$Y1 = X1 + E1 \qquad (3.2.3.1)$$

$$Y2 = X2 + E2 \qquad (3.2.3.2)$$

$$X2 = Y1 \qquad (3.2.3.3)$$

The rms value of the AC signal is then computed by doubling the first output of the sensor, Y1 and subtracting the second output of the sensor, Y2. This procedure is shown in the following formula:

$$\text{rms value} = 2Y1 - Y2 \qquad (3.2.3.4)$$

Substituting equations (3.2.3.1) through 3.2.3.3) into equation (3.2.2.4) results in the following equation:

$$\text{rms value} = X1 + (E1 - E2) \qquad (3.2.3.5)$$

The value that is computed by this equation is the one that is displayed by the meter as the value of the AC signal.

The Fluke 8506A has three different AC voltage settings to choose from; high accuracy, enhanced and normal. All of these AC settings (modes) have eight voltage ranges starting at 100 mV and ending at 500 V. In addition all three have 5½ digits of resolution.

The high accuracy (hi accur) mode produces the most accurate measure of the AC signal and is the mode used for our noise measurements. The mode has a six second reading time for each measurement. Fluke claims that signals may be measured in this mode to within: 0.012% if the frequency is between 40 Hz and 20 kHz; 0.04% if the frequency is between 20 kHz and 50 kHz; and 0,2% if the frequency is between 50 kHz and 100 kHz. The noise signal we are measuring falls within the two lower frequency ranges, so the exact accuracy of the measurements will have to be determined.

The enhanced (enh'd) mode takes an initial high accurate measurement of the signal and uses this to correct all subsequent measurements. Since noise varies from moment to moment, the enhanced mode is not suitable for our measurements.

The normal (normal) mode provides faster reading time than the other two modes but has much less accuracy. Since we need to sample the noise for many seconds to get accurate results, this mode will also not be useful to us.

The IBM Personal Computer and the IBM Proprinter are the other two pieces of commercial equipment used in this system. The IBM PC is used to run computer programs which control the noise measurements made with this system. With the installation of a card (in our computer a Metrabyte card) in the computer, command signals may be sent through a bus (in our case an IEEE-488 bus) to the Fluke 8506A. These commands may be cited within computer programs which in turn tell the Fluke 8506A when to make a measurement and determines the timing of the measurements. The readings made by the Fluke 8506A are transmitted back to the computer via the bus. These readings may be manipulated like any other data within a program. This feature allows for easy storage of data as well as mathematical manipulation of the data. The printer is necessary if a paper copy of the measured data and any manipulation of such data is desired.

3.2.4 Measurement Procedure

The second measurement system also has a calibration procedure as part of its measurement procedure. The calibration procedure will measure the gain and noise of the system for three calibration resistors (10, 100, and 1000 ohms). This data will be used to calculate a set of constants. These calibration constants will be used to determine the dynamic resistance and noise ratio associated with the DUT.

Due to the choice of gains in the stages preceding and following the filter there are eight possible configurations for this system. Calibrations of each of the eight configurations should be performed. Only if you know the rough noise level at the output of the circuit and thus have selected the gains, could you just do a calibration for one of the configurations. If the noise level is unknown all configurations will have to be calibrated.

The calibration procedure for this system is similar to the procedure for the first measurement system for several reasons. Both system use the same bias circuit and amplification stages. The input terminal for the calibration signal (CAL input terminal) are the same for both systems. The quantities measured in both procedures are the gain and noise of the system.

The first step in the calibration procedure is to obtain the frequency response of the system. From this frequency response the gain of the system from the DUT node to the output may be found. This gain must be known so that all noise quantities measured at the output of the system can be divided by this gain so that they are referred band to the DUT itself.

The best way to find the frequency response of this system is to determine the frequency response of the circuit and filter separately and then multiply the two responses together to get the complete system response. Both of these frequency response measurements will be performed with the HP 3562A. This instrument was selected for its ease of measurement, as well as its storage and mathematical capabilities. It will be possible to make the frequency measurement on the circuit, save the display, and later multiply it by the frequency response of the filter to get the complete system response.

The frequency response measurement of the circuit has already been made in the calibration procedure for the first measurement system. The correction waveform measurement was a frequency response measurement for the circuit from the DUT node to the OUTPUT terminal of the circuit. The frequency range of the second system is different from the first system, so a new correction waveform measurement will have to be made. To make this measurement, just follow the procedure described in section 4.1.4.

To make a frequency response measurement of the filter, set the system up in the manner shown in Figure 3.2.4.1. The source of the

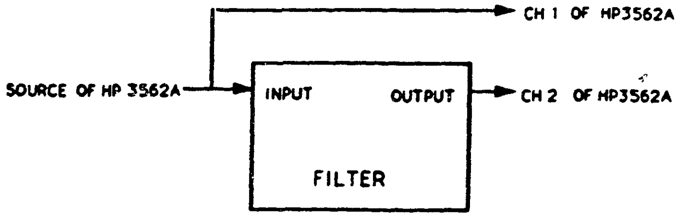


Figure 3.2.4.1 Setup for Frequency Response Measurement of Filter

HP 3562A is attached to the INPUT terminal of the filter and to its own channel one. The OUTPUT terminal of the filter is attached to channel two of the HP 3562A. The HP 3562A is set for a sweep sine frequency response measurement using state 3, see Figure 3.1.4.2. The frequency range of this measurement must be set to observe the passband of the filter, a range from 100 Hz to 100 kHz will be good for this filter. A series of ten sweep sine measurements, with a HP 3562A calibration between each measurements, is taken using the auto sequence "Start w/Cal", seen in Figure 3.1.4.3. After the measurements are complete the result is displayed on the screen of the HP 3562A. The mathematical capabilities of the instrument may be utilized to obtain the absolute magnitude squared gain of the filter.

This frequency response is then multiplied by the response for the circuit, which was saved in the HP 3562A, to get the total response of the system. This total response should be integrated and the largest value recorded. This is the value that will be used to divide all noise measurements made with the system, to refer then to the DUT node. The noise quantities measured by the Fluke 8506A will be in units of Volts. These quantities should be squared before being divided by this system gain factor. Since this system gain factor is in units of Hertz, due to the integration, the noise referred to the DUT node will be in units of Volts^2/Hertz.

The second step in the calibration procedure is to measure the gain of the noise amplifier from the CAL input to the OUTPUT terminal of the filter, at a single frequency. This quantity will be used later, during measurements to find the dynamic resistance of the DUT. A Fluke 5200A AC Programmable Calibrator (Fluke 5200A) along with the Fluke 8506A can be used to make this measurement. Since the signal used in this measurement is a single frequency, the filter is not necessary for this measurement. It may be advisable to remove the filter from the system for this measurement. If this is done, concerns over changes in the gain of the system due to the ripple in the passband of the filter may be eliminated.

The setup for this measurement may be seen in Figure 3.2.4.2. The Fluke 5200A is attached to the CAL input. The signal injected by this calibrator is measured by the Fluke 8506A. The AC signal at the OUTPUT terminal is also measured by the Fluke 8506A.

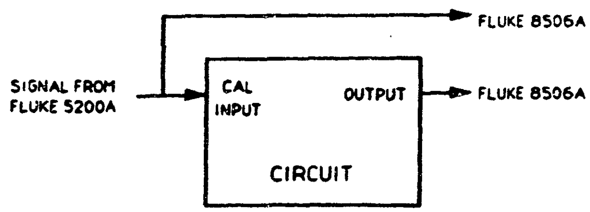


Figure 3.2.4.2 Setup for a Fixed-Frequency Gain Measurement Using the Fluke 5200A and the Fluke 8506A

The 1 kΩ calibration resistor is used as the DUT. The Fluke 5200A is adjusted to produce a sine wave at 5.2 kHz. The amplitude of the sine wave is adjusted so that the signal at the OUTPUT terminal is 1.6 V. The amplitude of the input signal and the amplitude at the OUTPUT terminal of the filter should both be measured by the Fluke 8506A. The gain from CAL input to the OUTPUT terminal of the filter may be found by dividing the voltage at the OUTPUT by the voltage at CAL.

The third step in the calibration is to measure the noise of the 1 kΩ resistor using the Fluke 8506A. The setup for this measurement may be seen in Figure 3.2.4.3. The OUTPUT terminal of the filter is attached to the Fluke 8506A. Since this measurement is computer controlled, the IBM PC is attached to the Fluke 8506A.

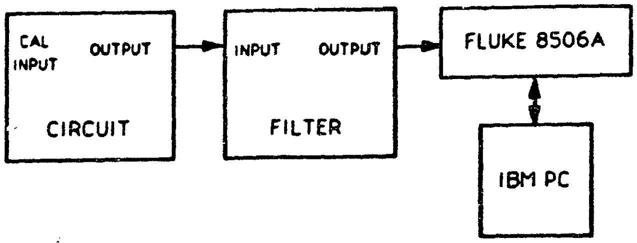


Figure 3.2.4.3 Setup for Noise Measurement

The noise measurement is controlled by the computer using the BASIC program sys22.bas, which may be seen in Appendix C. This program makes the Fluke 8506A take a number, from 1 to 1000, of high accuracy AC voltage readings of the noise signal. The program computes the average value and standard deviation for these readings and displays these quantities along with the individual measurements. For the calibration, 100 readings are selected in the program. The average value for he noise is record. This quantity is in rms Volts, it should be squared and divided by the system gain factor found in the first step of the calibration. This will refer the noise back to the DUT node and convert it to spectral density units of V^2/Hz.

The fourth step in the calibration is to estimate the resistance of the 1 kΩ calibration resistor. This is done by measuring the voltage across the resistor and the current through it, with a meter. These quantities are measured with the Fluke 8506A. The voltage is measured at the VOLTAGE terminal of the circuit. The signal measured at the CURRENT terminal of the circuit is actually a voltage, since the bias current is measured across a resistor. The resistance of the 1 kΩ resistor is estimated using the following formula,

$$R_{DUT} = \frac{V_{VOLTAGE} * R_s}{V_{CURRENT}} \qquad (3.2.4.1)$$

where R_s is the sum of the resistor which the current is measured across, which has a resistance of 100 Ω, and the resistance of the wire, which has a resistance of 0.011Ω.

Steps two through four of the calibration procedure are repeated for the 10 Ω and 100 Ω calibration resistors. After these measurements are complete, four wire resistance measurements of the calibration resistors are made with the Fluke 8506A. The gain and noise data obtained through the calibration along with the calibration resistance are used to find the calibration constants. The program cal.bas, found in Appendix C, can be used to find the calibration constants for this system as well. This is possible because the only difference between this and the first system is the additional gain stages and the filter. The calibration measurements of gain and noise account for these additions to the system. The equations used to find the calibration constants will not be changed by the addition of the gain stages and the filter. Cal.bas solves three linear equations for the calibration constants, C_r, G and K_r. The program also solves three linear equations for the calibration constants A, B, and C.

These six constants are used with noise measurements and gain data for diode DUTs to determine the noise ratio and dynamic resistance of the DUTs.

Now that the calibration procedure is complete, measurements can be made with this system. The constants found in this calibration are valid until things are changed in the system. Then a new calibration should be conducted. Actually you may want to recalibrate after a couple of months, even if no changes are made, because the components in the system could be drifting.

The measurement procedure for the second system is very similar to that of the first system. The procedure begins by installing the DUT, which is usually a diode, and biasing it at a certain current by switching in appropriate bias resistors. The bias voltage and bias current measurements are made in the same manner as the first system. (See section 3.1.4 for more details.) The bias voltage and current data will be used in the study to determine if there is correlation between the noise and radiation characteristics.

The next measurement to be made is of the gain of the system from CAL input to the OUTPUT terminal of the noise amplifier. This measurement is made using a sine wave, from the Fluke 5200A, at a fixed frequency of 5.2 kHz. Following the procedure described in step two of the calibration to make this measurement. Record the gain of the system.

The gain measured in this step is used along with the three calibration constants, G, K_r, and C_r, to find the dynamic admittance, G_d, of the DUT. This admittance is found using equation 3.1.4.2 The dynamic resistance, R_d, is obtained by inverting G_d.

The last quantity to be measured is the noise at the output of the system. This noise signal is measured by the Fluke 8506A. The measurement is controlled by the sys22.bas program. One hundred samples are taken and averaged for each noise measurement. The quantity, V_o, the average value of the measurements made by the Fluke 8506A, is in units of rms Volts. This quantity should be squared and divided by the system gain factor calculated in step one of the calibration. The result is a noise spectral density, S_o, which is referred to the DUT node and is in the appropriate units of $Volts_2$/Hertz.

Noise ratio is the quantity we wish to calculate for each DUT. Equation A.6, found in Appendix A, describes noise ratio. The reverse or bias current along with the current noise spectral density, S_{id}, must be known to calculate noise ratio. The bias current was directly measured. S_{id}, may be obtained from the quantity S_o after some manipulation. The steps that need to be followed to convert S_o to S_{id} have already been described in section 3.1.4. However, now the calibration constants and measurements used in this conversion must all be associated with the second measurement system.

Section 4 EVALUATING THE ACCURACY OF MEASUREMENTS MADE WITH EACH SYSTEM

To determine the limitations to the accuracy of measurements made with each of the noise measurement systems as well as the accuracy itself, various aspects of the systems have to be analyzed. The design of the circuits used in the systems as well as the design of the system itself had to be studied to insure that they are appropriate for measuring noise accurately. Circuit components and portions of the systems that could contribute excess noise to the systems and thus limit accuracy have to be identified. The measurement and calibration procedures have to be studied to insure that they would result in the most accurate measurements. The specifications of any instruments used in the systems has to be known, because the accuracy of measurements would be definitely limited by the instrument accuracy. Other considerations, like sampling time and averaging, also have to be explored.

This section describes in detail the various aspects of the systems that I looked at in determining the limitations to the accuracy and the actual accuracy of noise measurements made with each of the systems. Section 4.1 concentrates on the first measurement system, while section 4.2 concentrates on the second measurement system.

4.1 The First Measurement System

4.1.1 Accuracy of Commercial Equipment

The accuracy of measurements made with this system depends on the accuracy of the HP 3562A. The specifications for the HP 3562A, which may be found in Appendix D, mention several types of amplitude accuracy. It defines accuracy "as Full Scale Accuracy at any of the 801 calculated frequency points." [10] It defines the overall accuracy as "the sum of absolute accuracy, window flatness and noise level." [10] Numbers are given for the absolute accuracy, window flatness and the noise floor is discussed. For evaluating the accuracy of noise measurements, the overall accuracy will be used. Since the window flatness and noise level are small and change the accuracy only slightly, the overall accuracy will be estimated by the absolute accuracy alone.

From the specifications, the absolute accuracy is ± 0.15 dB ± 0.015% of input range, for signals between 24 dBV and -40 dBV, and ± 0.25 dB ± 0.025% of input range, for signals between -41 dBV and -51 dBV. In this thesis accuracy will be described in terms of percentage. Therefore, the accuracy of the HP 3562A must be converted into percentages. This is conversion is described by the following equation.

$$\%\ \text{Accuracy} = 100*(10^{(x/20)} - 1) \qquad (4.1.1.1)$$

where x is either 0.15 dB or 0.25 dB depending on the range. The conversion was made assuming that a voltage was being measured, so a factor of 20, appears in the equation. Using the conversion, the first range has an accuracy of 1.7% and the second range has an accuracy of 2.92%.

The percentage of input range was ignored in the conversion. The major reason for doing this is that the input range is often unknown. The input range depends on the signal being measured. If the range of the signal is known, one can set the input range and thus know its quantity. However, autorange is often used in measurements, in this case several ranges may be used in the measurements and one does not know exactly which ones. Another reason for ignoring the percentage of input range in determining the accuracy is that this contribution is small. Looking at the smallest and largest allowed ranges, -51 dBV and 24 dBV respectively, and calculating the percentage one finds 0.705 μV for the smallest range and 2.38 mV for the largest range. These additions are small compared to the level of signal and therefore can be ignored without much effect on the accuracy.

Actually the specifications for the accuracy of the HP 3562A is an indication of the worst accuracy one can expect. Most measurements will achieve better accuracy. A simple experiment illustrates this point. A 1 V (rms) sine wave at 90 KHz was produced by a Fluke 5200A AC calibrator. The signal was read by both the HP 3562A and the Fluke 8506A. The HP 3562A was set for a standard noise measurement using a 20 kHz bandwidth from 80 kHz to 100 kHz. The Fluke 8506A read 1.00085

V (rms), while the HP 3562A read 49.9282 $\mu V^2/Hz$ (rms). The reading form the HP 3562A must be converted to V^2 by multiplying by the bandwidth of the measurement, 20 kHz. The Fluke 8506A can measure AC signal more accurately than the HP 3562A, so in to calculate the error, assume the Fluke reading is exact. The error, which is also the accuracy, is found using the following relation.

$$\%error = \frac{\text{HP reading - Fluke reading}}{\text{Fluke reading}} * 100 \qquad (4.1.1.2)$$

Substituting the numbers into equation 4.1.1.2 results in

$$\%error = 0.31\%$$

This simple experiment showed that the HP 3562A accuracy to be better than the expected 1.7 %.

The Fluke 8506A Thermal True RMS Multimeter is used in this system to measure the voltage and bias current of the DUT. The bias current will be used in the calculation of the noise ratio. Both the current and the voltage will be later used in determining if there is correlation of noise and radiation characteristics. The current is measured across an 100 Ω resistor, so both measurements are of DC voltages. According to the specifications of the Fluke 8506A, it is capable of measuring signals in the 100 mV range to ±(0.0018% of reading + 15 counts) in the norm (normal) mode and to ±(0.0010% of reading + 8 counts) in the avg (average) mode. In the 1 V range it is capable of measuring signals to ±(0.0008% of reading + 7 counts) in the norm mode and to ±(0.0005% of reading + 4 counts) in the avg mode. The bias voltage and current measurements are usually made in the norm mode, therefore; the accuracy is ±(0.0018% of reading + 15 counts) for the 100 mV range and ±(0.0008% of reading + 7 counts) for the 1 V range.

Before proceeding, the term count will be explained by example. The specifications for the 1 V range is ±(0.0008% of reading + 7 counts). In this range, the display should read 1.000000, for a 1 V input. Seven counts corresponds to an uncertainty of seven in the last digit or to 0.000007, 7 μV, or 0.0007 % error at 1 V. The total accuracy for this range may be written as ±(0.0008% of reading + 7 μV) for clarity.

In determining the accuracy of the measurements made with this system, the worst case accuracy of the HP 3562A will be assumed to be 1.7% for a signal falling between 24 dBV and -40 dBV and 2.92% for a signal falling between -41 dBV and -51 dBV. The accuracy of the Fluke 8506A will be taken as ±(0.0018% of reading + 1.5 μV) in the 100 mV range and ±(0.0008% of reading + 7 μV) in the 1 V range.

4.1.2 Analysis of the Circuit Portion of the System

The purpose of this section is to look at the design of the circuit used in this measurement system and determine if there are any components or sections that add large quantities of noise to the system. These critical components or sections may be identified through a noise model for the circuit. The three sections of the circuit are modeled and the noise is estimated from the models. In calculating the estimated noise, dominant noise sources are easily identified.

Throughout this section there are references to the noise models for certain components. The noise models for the most common components may be seen in Appendix B. Along with each model is a brief explanation of the model and how to use it.

The bias circuit is the first portion of the system to be modeled. The circuit diagram for this section was shown in Figure 3.1.2.1. The noise model for this section can be seen in Figure 4.1.2.1. Looking at Figure 4.1.2.1, one will notice not all components are modeled, while others are not modeled with their typical noise models shown in Appendix B. These differences are a result of considering the operation of the circuit while developing the model.

The National Semiconductor LM399 voltage reference is modeled as a voltage noise generator. The two operational amplifiers, U2 and U3, are modeled using a modified amplifier model. This model differs from the ones shown in Figure B.3, since it only includes the voltage noise generator and the current noise generator of the amplifier. The closed loop gain and not the open loop gain is used with this model. The FET, at the output of the U2 amplifier, is not modeled because it is in a position where it will not significantly effect the noise of the system. All of the resistors are modeled like those in Figure B.1 All capacitors are not modeled, because there is no noise model for

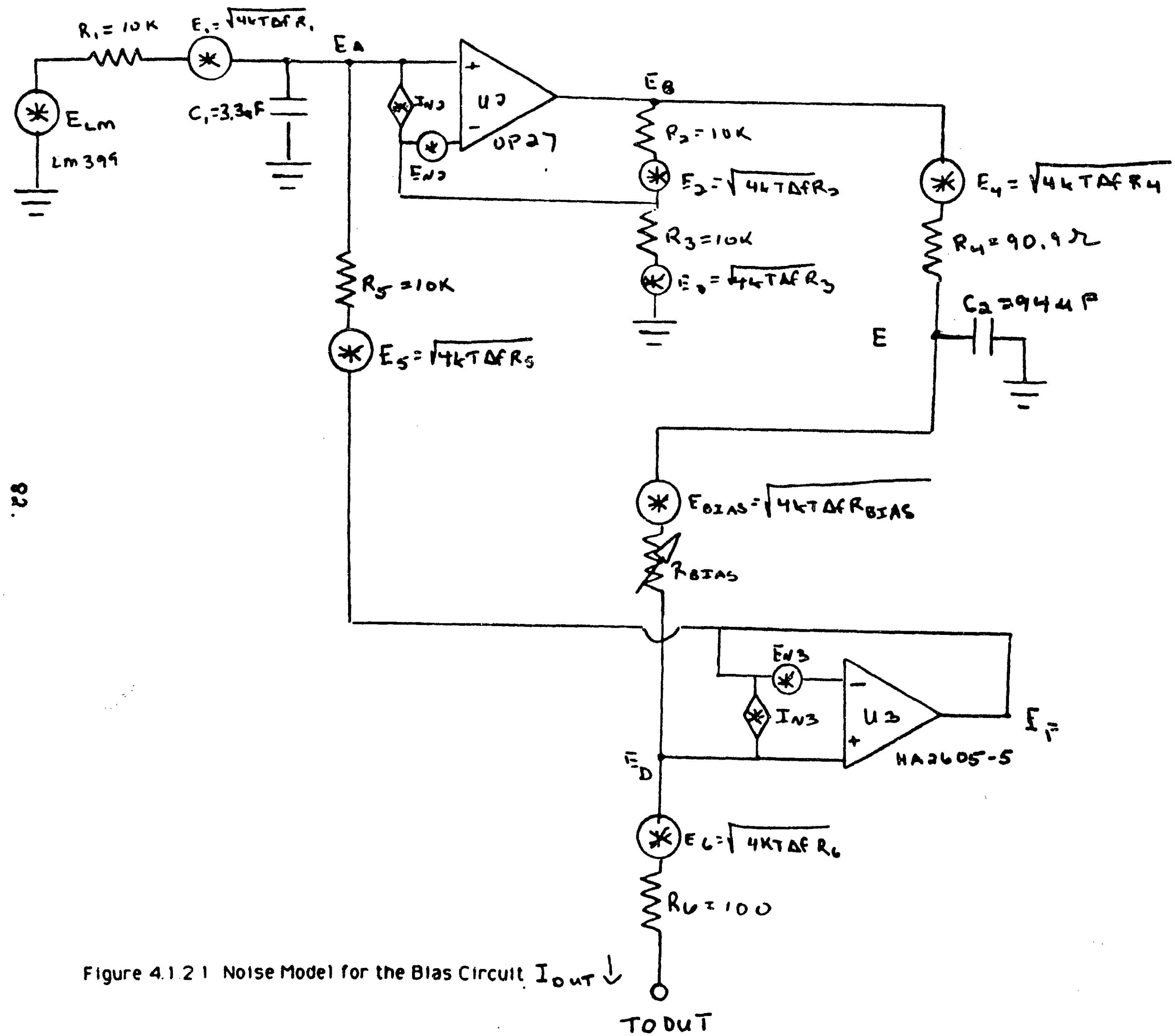


Figure 4.1.2.1 Noise Model for the Bias Circuit I_{OUT} ↓

Bias Circuit Noise Mode"
J. Furlong

capacitors, since they do not produce significant noise. The noise levels for the various components used in this circuit can be found in their specification sheets located in Appendix E.

The noise of the bias circuit may be calculated from this model. This is done so by tracing the noise signal from the input to the output. This method is the same as tracing a signal from the input of a circuit to the output. The noise model is roughly a small signal model and all relevant circuit analysis techniques may be utilized. Please see Appendix F for a step by step analysis of the noise calculation. The noise this portions of the system injects into the node at the top of the DUT is the quantity we wish to know. This signal will be amplified by the amplification stages. If this signal is significantly larger than the DUT noise, it could swamp the DUT noise and seriously affect any noise measurements.

From the step-by-step noise calculation for this portion of the circuit, found in Appendix F, several noisy components have been identified. The National Semiconductor LM399 voltage reference contributes a very significant amount of noise, but this noise is effectively attenuated by the filters following it. The input voltage and current noise of the OP-27, although not terribly large, are amplified by the U2 amplifier. This noise at the output of the U2 amplifier is attenuated by a low pass filter to the point of insignificance. The only significant noise terms are the thermal noise of the bias resistors. Since the bias resistance is varied, this noise is also varied, but in any case it has to be considered.

The noise calculation for this part of the circuit has shown that the current noise injected at the DUT node is several orders of magnitude less than the shot noise of the DUT. This injected noise will be combined with the shot noise and the combined noise will be amplified by the circuit. In later calculations of the noise of the DUT, we must be careful that this injected noise is eliminated from the results so that the measurement will be correct and the accuracy of the measurement is not hurt. This injected noise will be discussed later in this paper, in the section dealing with limitations.

The next portion of the circuit to be modeled is that around the DUT node (socket). A simplified circuit diagram for this section may be seen in Figure 4.1.2.2. The noise model for this section may be seen in Figure 4.1.2.3. At the frequencies of operation the capacitors may be considered shorts. It will also be assumed that no calibration signal is being applied at the CAL input. This input will be shorted to ground. With these considerations in mind, one can see that the resistors, (R_{BIAS} + R_6), R_{10}, R_{11} and R_{12}, are all in parallel. Since they are all in parallel they may be represented as one resistor, R_x, which can be modeled in the usual way. R_{DUT}, is in parallel with R_x, but will not be combined with it since this is the noise source we are trying to measure and we want to compare all other noise sources to it.

The exact calculation of the noise of this section may be found in Appendix F. The calculation shows that the thermal noise of the parallel combination of resistors, although not larger than the DUT noise, is still large enough to be considered. In calculations of the

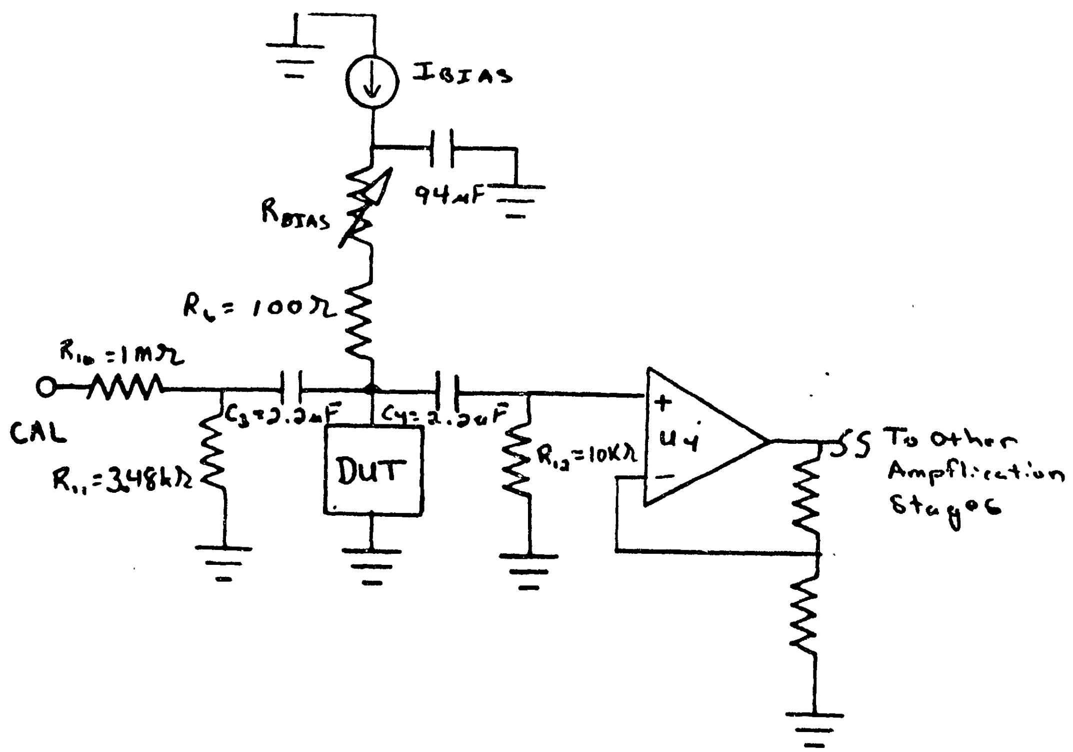


Figure 4.1.2.2 Circuit Diagram for Section Around DUT

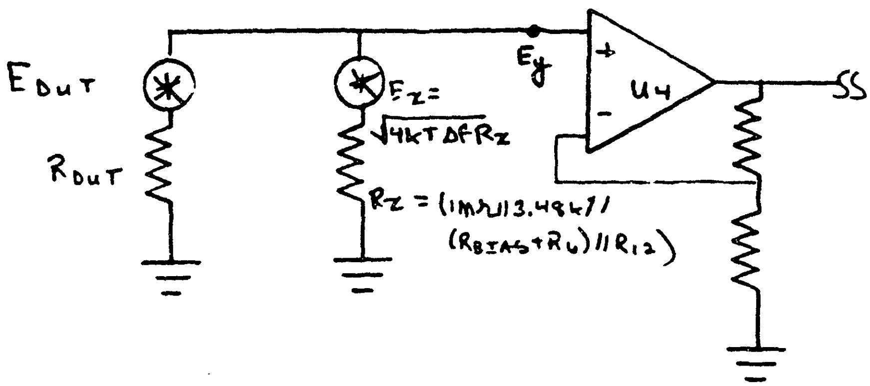


Figure 4.1.2.3 Noise Model for Circuit Around DUT

noise of the DUT, this thermal noise have to be eliminated from the measurements to insure the correct results. This will be discussed later in this paper.

The last part of the system to be modeled is the three amplification stages. The circuit diagram for this section may be seen in Figure 3.1.2.3. The noise model for this section may be seen in Figure 4.1.2.4. There are only three types of components used in this section; resistors, capacitors and op amps. Only the resistors and op amps are modeled. The resistors are modeled with their traditional model shown in Figure B.1. The op amps are modeled with the modified amplifier noise model, discussed above.

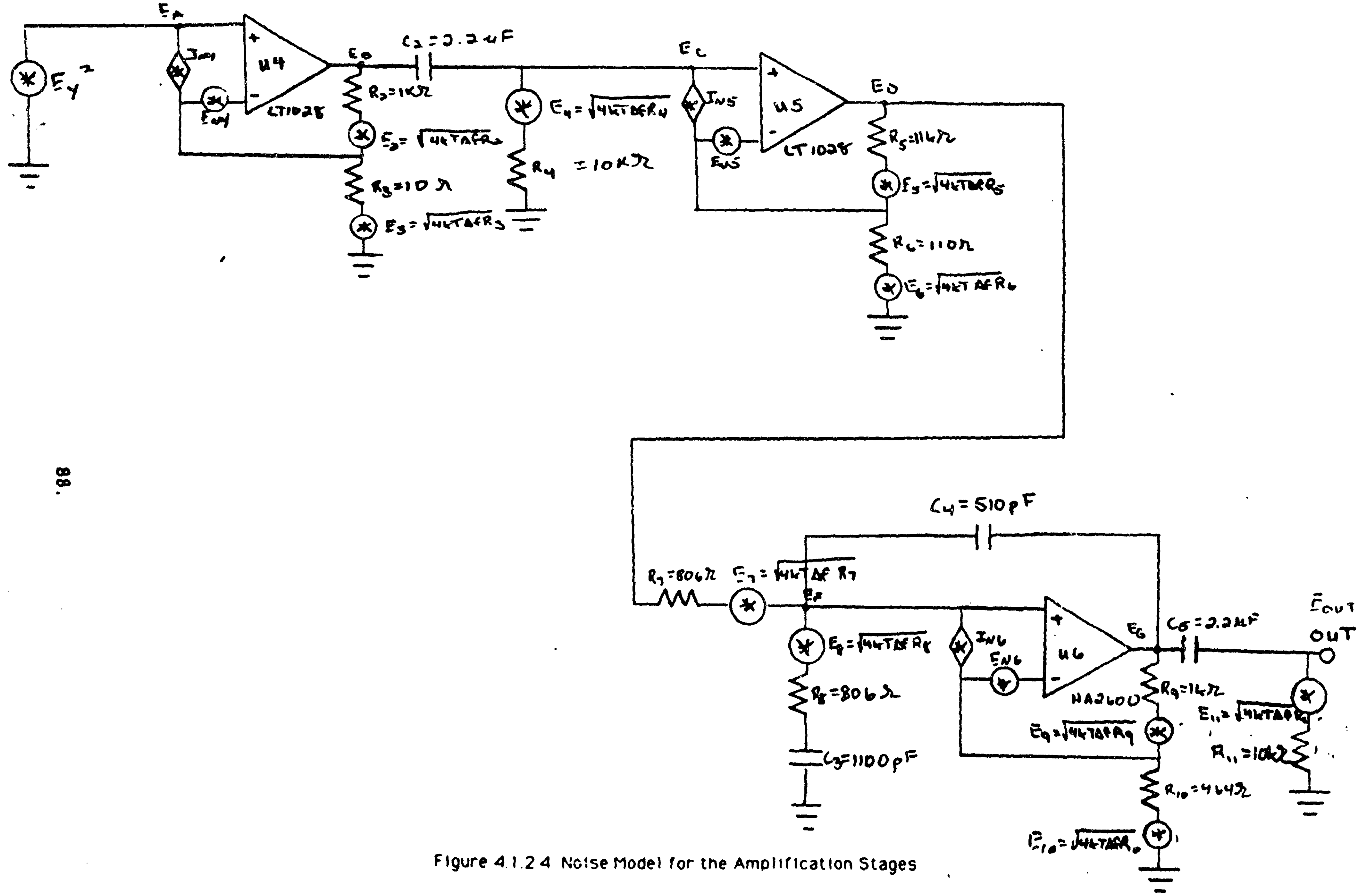


Figure 4.1.2.4 Noise Model for the Amplification Stages

The input noise signal to this part of the circuit is E_Y^2, which was calculated above. This input noise is amplified by the three stages. At the same time, the noise of each of these stages is added to this input noise. This sum of amplified noise and the noise of the three stages is the noise that is measured at the output of the circuit. All this may be seen in the step-by-step calculation of the noise which can be found in Appendix F.

The step-by-step calculation of the noise identified two critical components, the input noise voltage of the first LT1028 and the thermal noise of R_3, a 10 Ω resistor. Of these two critical components, the input noise of the LT1028 is the biggest contributor to the output noise of the two. The noise calculation also showed that the second and third stages contributed no noise to the overall noise, because the noise after the first amplification stage was larger than anything added to it after this point. The input noise to the amplification stages, which contains the DUT noise and the thermal noise of the parallel resistance combination, appears at the output of these stages, significantly amplified but added to an additional noise term. This additional noise term consists of the noise of the two critical components, mentioned above, all amplified. The calculation showed that the amplified input noise was smaller than the additional noise term, but was large enough not to be swamped by it. So it appears that these amplification stages have amplified the noise as they had been designed to do. They have done this without adding significant noise themselves. Two of the significant noise sources, the thermal noise of the parallel combination, R_X, and the thermal noise of R_3, can be calculated precisely if the resistance and temperature are known. The other significant term, the input noise of the LT1028 is also a well known quantity. Therefore, the excess noise should be easy to eliminate from any calculations of the noise of the DUT itself.

4.1.3 Analysis of the Calibration and Measurement Procedures

The calibration and measurement procedures used with the first measurement system were described in section 3.1.4. In the first part of this section, both procedures will be discussed in light of the common noise measurement methods discussed in section 2.3. Both of these procedures have to be reviewed to determine if they lead to accurate measurements. The results of these reviews are presented in the second part of this section.

Neither the calibration procedure nor the measurement procedure can be described exactly as a sine-wave, noise generator or correlation method of noise measurement. The calibration procedure resemble the sine-wave method the most, but it also resembles the noise generator method. The measurement procedure has the closest resemblance to the sine-wave method, but also resemble the noise generator method in one case.

There are several parts of the calibration procedure that are similar to the sine-wave method. The sweep sine frequency response measurement used to find the correction waveform also uses sine waves as the calibration signal. This correction waveform is equal to the magnitude of the gain squared. The output noise is measured with the calibration source set to zero. This is equivalent to having the CAL input shorted to ground. In the sine-wave method, a shorting bar is put at the input when the output noise is measured. Finally, the noise measured at the output is referred back to the DUT by dividing by the correction waveform. In the sine-wave method, the equivalent input noise is found by dividing the output noise by the gain squared.

Noise is measured at the output of the circuit. Then this noise is divided by the correction waveform, found in the calibration procedure, to refer the noise to the DUT node. Since the correction waveform equals the magnitude of the gain squared, this procedure is like the sine-wave method where the equivalent input noise is found by dividing the output noise by the gain squared. The gain from the CAL input to the OUT terminal is measured in the same way as in the calibration procedure, with the source amplitude adjusted to make the output 1.6 V.

It should also be noted that in both the calibration and measurements procedures, voltage and current measurements are made along with the noise and gain measurements. These voltage and current measurements will be used in the correlation of noise characteristics to the radiation characteristics of the diodes.

In the previous section, several different components and sections of the circuit were identified as noisy and potential problems. One of the things a calibration procedure should do in a noise measurement system is to characterize the system's noise. This characterization may eliminate the concerns about added noise if the noise of the system is a well known quantity and can easily be subtracted from measurements made with the system.

Part of the calibration procedure for this first measurement system consists of taking a gain and noise measurements for three resistors, 10 Ω, 100 Ω and 1 kΩ. The data from these measurements is used to find six calibration constants. It is through these measurements and calculations that the system is characterized.

Three of the constants, A, B, and C, found in the procedure, are used to calculate the output noise given the resistance for the DUT. In this equation the extra noise injected at the DUT node from the bias circuit is accounted for. This calculated output noise is subtracted from the measured output noise in the calculation of noise ratio. In this manner, the extra noise sources are eliminated from the measured noise. Therefore, the calibration procedure achieves its goal of characterizing the noise of the circuit so that concerns over noisy components can be alleviated.

The other part of the calibration procedure is to calculate a correction-waveform. The system was designed to have a flat frequency response within the bandwidth from 5 kHz to 55 kHz, where it was expected to operate. However, the system has been used in other frequency bandwidth to avoid 1/f noise in some DUTs. In these other ranges the frequency response is not necessarily flat. So the correction-waveform compensates measurements made in other frequency bandwidths for the non-flat nature of the response. Although this part of the calibration does not directly involve noise measurements, it is necessary for producing accurate noise measurements with this system. If the measurements were not corrected the non-flat frequency response could seriously affect the accuracy of the measurements.

A measurement procedure will be declared appropriate for a system as long as it is consistent with any related calibration procedure. It should also make all the measurements required by the project. The measurement procedure for the first system meets both of these criteria.

There are four quantities that are measured, the gain from the CAL input to the OUT terminal, the noise at the OUT terminal. the voltage and current at the DUT. The first two quantities coordinate with the calibration procedure and are used with the calibration constants to find the dynamic resistance of the DUT and predict the noise at the output of the circuit, assuming the device is just a resistor. This output noise calculation takes care of the non-DUT noise in the system. In this case, the measurement procedure is consistent with the calibration procedure and utilizes all constants calculated in that procedure. All four of the quantities measured in this procedure will later be used directly or indirectly in the project to determine correlation between noise and temperature and radiation characteristics of the DUTs. In this case, the measurement procedure makes all the necessary measurements for the overall project.

4.1.4 Discussion of the Effect of Sampling Time and Averaging on Accuracy

In this section the relationship between sampling time, averaging and accuracy of noise measurements will be discussed. The first half of the section will discuss accuracy equations which relate these quantities. The second half will discuss the sampling time and averaging considerations of the first measurement system.

Due to its nature, it is only possible to measure the instantaneous value of a noise signal. However, if the power of the signal is averaged over time it becomes possible to measure an average value for the noise. If the meter used to measure noise has a short time constant, it will try to measure the instantaneous value of the noise signal instead of an average value. The result is fluctuating readings. These fluctuations affect the accuracy of the noise measurement.

An equation relating accuracy to the noise equivalent bandwidth a system and the time constant of the meter used in the system has been developed. [2][3] The equation for a square law detector, one that measures power, is

$$\sqrt{\alpha} = \frac{1}{\sqrt{2\tau\Delta f}} \qquad (4.1.4.1)$$

$\sqrt{\alpha}$ = relative accuracy

τ = time constant of meter

Δf = noise equivalent bandwidth

This equation was published before digital instruments and computers were common. For a measurement system that averages power for time T,

$$\sqrt{\alpha} = \frac{1}{\sqrt{T\Delta f}} \tag{4.1.4.2}$$

According to Motchenbacher and Fitchen, the accuracy, $\sqrt{\alpha}$, is equal to "the ratio of the rms value of the meter fluctuation to the average meter reading."[2:296]

After studying this equation, one will realize that better accuracies at narrow bandwidth are achieved if long time constants or integration times are used. At higher bandwidths, a short sample time may be used to achieve relatively good accuracy. To achieve better accuracy you can change either the bandwidth or the time constant. Usually the bandwidth of system can be changed. However, in noise measurements one must be careful that the low end of the bandwidth used does not contain any 1/f noise, and the high end is limited by equipment, device capacitance, or something. The time constant of a meter is usually fixed. But the number of readings taken by a modern instrument under computer control is limited only by the time available to make a measurement.

The HP 3562A is an integrating instrument, so equation 4.1.4.2 should be used with the first measurement system to find its accuracy. To calculate the accuracy of the noise measurements made with the first measurement system, we must know the noise equivalent bandwidth of the system and the integrating time of the HP 3562A. The noise equivalent bandwidth is the frequency span or bandwidth that the operator selects for the measurement. The number of averages, which is need to determine the integrating time is also selected by the operator. The only quantity that is left is the time per measurement of the HP 3562A. This quantity can be obtained from the operating manual [10]. Before presenting the value of the time, it will be useful to discuss how the HP 3562A makes a measurement, because the time per measurement was determined by the way the measurement is made.

The HP 3562A samples the signal being measured and converts it to digital. Any necessary digital filtering is performed and then an FFT is performed on the digital signal. After that averaging is done. The result is prepared for the display by converting it to the correct units, etc.

The HP3562A displays a maximum frequency of 100 kHz. To avoid aliasing, its A/D samples at a frequency of 256 kHz. This value meets the Nyquist criteria, since it samples at twice the highest frequency. Sampling at a frequency of 256 kHz is equivalent to saying 256,000 samples are taken per second or that a sample is taken every 3.91 μs.

The HP 3562A always takes enough data to compute a display that will fill the 100 kHz span, even when a smaller span is selected. If a smaller span is selected, this is called a zoom measurement if the lowest frequency is larger than 0 Hz, the HP 3562A, digitally filters the data to fit the selected span. Although the data is filtered, the sample rate for each point is not altered.

To make its measurements, the HP 3562A works with a time record. This time record consist of 2048 points, if the full span of 100 kHz is selected, and 1024 points, if a smaller span is selected. Each of these points are either sampled at 3.91 μs or are selected from a larger number of points coming out of a digital filter, that eliminated high frequencies. Therefore, the total measurement time is equal to the number of points per record multiplied by the time for one point. For a 2048 point record that goes to 100 kHz, the total measurement time is 8 ms, while for a 1024 point record that goes to 100 kHz, the total measurement time is 4 ms.

When averaging is selected, a series of records, converted to the frequency domain by FFTs, are averaged together. For example, consider the case where two is selected as the number of averages. Two time records are consecutively made, filtered and converted to the frequency domain. These records are then added together and the sum is divided by two to get the average which will be displayed. In light of this information, the effective integration time for the HP 3562A will equal the number of averages multiplied by the individual measurement time, 4 ms for a 50 to 100 kHz span for example.

4.1.5 Summary of the Limitations on the Accuracy of the System

The purpose of the preceding four sections was to identify possible limitations to the accuracy of this noise measurement system. These limitations will be summarized in this section and some additional constraints will be discussed. All the limitations will be ranked to show which affects the accuracy the most.

In most cases, the presence of noisy components and/or circuit sections, would limit the accuracy of the measurements. In section 4.1.2, several noisy components and circuit sections were identified.

However, in section 4.1.3, the evaluation of the calibration procedure showed that the stationary noise produced by these components were eliminated from the measured noise, through the calibration procedure. Thus, noisy components are not a limitation to the accuracy of the first system's measurements to the extent that the noise they produce is predictable.

The specifications for the HP 3562A present several limitations to the measurement accuracy. First, this analyzer may only measure signals to 1.7% accuracy, if the signal falls between 24 dBV and -40 dBV, and 2.92% accuracy, if the signal falls between -40 dBV and -51 dBV. These numbers are a worst case, so the actual accuracy should be better, but there still will be some limit on the systems accuracy produced by the HP 3562A accuracy. Second, the HP 3562A also has a fixed sampling rate,of one sample every 3.91 μs and a fixed record length of 1024 points for frequency spans less than 100 kHz. This is the type of span that will be used in these measurements. Together these quantities fix the time record length at 4 ms. Third this time can be extended by using the averaging capabilities of the HP 3562A, but there is a limit to the number of averages that can be selected, 32,767. Fourth, the maximum bandwidth of the HP 3562A and thus of the whole measurement system is restricted to 100 kHz. Recall that the two variables in the accuracy equation are bandwidth and the time constant, the specifications of the HP 3562A put a lower limit on the accuracy of the system.

The bandwidth of the measurement system is further constrained by the concerns for 1/f noise. To eliminate the 1/f noise from the measurements, a low frequency cutoff must be set. Depending on the device, these cutoffs range from 5 kHz to 80 kHz. Considering that measurements can only be made to 100 kHz, the bandwidth of the system can range from 20 kHz to 95 kHz. This reduction in bandwidth will further constrain the accuracy of the system.

This section has shown there are three limitations to the accuracy of the system. These limitations are the accuracy of the HP 3562A, the bandwidth of the system and the fixed intergration time of the HP 3562A. It is inappropriate to rank these three limitations because they should all affect the accuracy equally.

4.1.6 The Accuracy of the First Measurement System

The accuracy for the first measurement system can be determined both theoretically and experimentally. The theoretical prediction of the accuracy may be obtained by using the accuracy equation discussed in section 4.1.2.5. The experimental prediction may be obtained by using a resistor, or several different resistors, as the DUT. The noise of these resistor DUTs will be measured in the same manner as any diode DUT. The thermal noise or even the resistance can be calculated from the measured noise and compared to the expected value for the device. Thus, the accuracy of the system can be obtained. This is the procedure W. Lukaszek used to determine the accuracy of this system. (See section 2.5.2.) Both methods of obtaining the accuracy will described in more detail in this section. A comparison between the two methods will be made and the results will be presented.

The accuracy of a noise measurement can be theoretically predicted by the accuracy equation discussed in section 4.1.4, as long as the sampling time and bandwidth are known. As shown in section 4.1.4, both these quantities are known for the first measurement system. The only thing that must be kept in mind is that the bandwidth of the measurements depend on the device being tested, due to avoidance of 1/f noise. Variations in bandwidth will cause variations in the accuracy.

All noise measurements made with the HP 3562A use frequency spans less than the full 100 kHz span and with the low frequency cutoff larger than 0 Hz. Therefore, all these measurements and have 1024 points in their time record. That means the sample length is 4 ms is the measurement goes to 100 kHz. The typical noise measurement uses state 1 with the "Start w/cal" autosequence (see section 3.1.4). That means ten, 1000 average measurements are taken and averaged together, for a total of 10,000 averages per measurement. Thus the sample time, the number of averages times the time per single measurement, equals 40 s. Using this time along with the two most common bandwidths, 20 kHz and 50 kHz, the accuracy of the noise measurements are calculated below.

For the 20 kHz bandwidth

$$\sqrt{\alpha} = \frac{1}{\sqrt{(40s)(20\ \text{kHz})}}$$

$$\sqrt{\alpha} = 1.12e\text{-}3 * 100\% = 0.112\% \qquad (4.1.6.1)$$

For the 50 kHz bandwidth

$$\sqrt{\alpha} = \frac{1}{\sqrt{(40s)(50\ kHz)}}$$

$$\sqrt{\alpha} = 7.07e\text{-}4 * 100\% = 0.071\% \qquad (4.1.6.2)$$

The experimental determination of the accuracy is obtained by making measurements with a set of resistors. Five resistor were selected with the following values, 10 Ω, 49 Ω, 100 Ω, 499 Ω, and 1 kΩ. These resistor DUTs are measured just like they were a diode, with the exception that the bias current does not have to be adjusted to the breakdown value, as in the diode measurements. This measurement procedure was described in detail in section 3.1.4. An additional measurement, of the temperature inside the box containing the circuit, is also made. This measurement is achieved by measuring the resistance of a thermistor located inside the box.

The value of the gain measured for the DUT is substituted into equation 3.1.4.2 to find the admittance of the DUT. The resistance of the DUT, R_d may be found by inverting the admittance. The total resistance seen at the input of the amplification stages, R_t may be found using equation 3.1.4.3.

The total resistance, R_t, calculated above, is substituted in equation 3.1.4.3 to predict the noise at the output of the circuit. In normal measurements, this noise would be subtracted from the output noise to eliminate the excess noise in the system. This equation was determined assuming that the DUT was a resistor. In this measurement the DUT is a resistor, so subtracting it from the output noise should result in zero. It is not necessary to calculate the current spectral density for this DUT to determine the accuracy, because accuracy can be determined from the voltage spectral density just as easily. The error of the measurement may be determined from how much the noise measured at the output differs from the calculated output noise. Accuracy is equal to the error. This may be illustrated by the following equations.

$$\text{error} = \frac{\text{calculated - measured}}{\text{measured}} \tag{4.1.6.3}$$

$$\text{error} = \frac{e_o^2 - S_V}{e_o^2} \tag{4.1.6.4}$$

The measurement for the five resistors were made in the frequency span of 80 kHz to 100 kHz. This was the frequency range that the system was set up for and I did not wish to change it to make resistance measurements. Making the measurements in this range, gives an opportunity to study the effect of the correction-waveform on the measurements, since it is needed in this frequency span. The calibration constants used with this span may be seen in Table 2. The experimental data for the five resistors may be seen in Table 3. The calculated values, of G_d, R_d, e_o^2, and the error are given in Table 4.

Table 2 Calibration Constants

Constant	Value
C_r	-6.80612e-5
G	3.88728e-4
K_r	3.19929e-2
A	2.69192e-18
B	1.63908e-20
C	1.45490e-23

Table 3 Resistor Measurement Data

Resistor	V (mV)	I (mA)	Gain	Noise (V^2/Hz)	Ω	Temp. (°C)
10 Ω	3.054	0.307	0.3163	2.84078e-18	9.923	27
49 Ω	15.324	0.307	1.5608	3.51476e-18	49.841	26
100 Ω	30.721	0.307	3.0624	4.37363e-18	100.015	24
499 Ω	152.753	0.307	12.9999	1.17740e-17	497.378	27
1 kΩ	3070.612	0.307	22.0575	2.09466e-17	999.850	27

Table 4 Calculated Values

Resistor	G_d Ω^{-1}	R_d Ω	e_o^2 V^2/Hz	error %
10 Ω	1.007e-1	9.93	2.85545e-18	+ 0.52
49 Ω	2.011e-2	49.73	3.52626e-18	+ 0.33
100 Ω	1.006e-2	99.42	4.39415e-18	+ 0.47
499 Ω	2.072e-3	482.56	1.17544e-17	+ 0.17
1 kΩ	1.062e-3	941.89	2.09083e-17	+ 0.18

Looking at the error column in Table 4, it appears that we are getting respectable accuracy for our measurement. Of course, we must consider the fact that two of the resistors, the 100 Ω and the 1 kΩ, are two of the resistors used to calculate the calibration constants. Thus, the error for the noise measurements should be zero. The error is not zero, but these calculation may test the stability of the system instead of the accuracy, since the calibration was made several months before these resistor measurements were made. However, the measurements on the other three resistors should give some indication of the accuracy of the system. The diode noise measurements, which we are really concerned with, will be dealing with noise quantities larger than just thermal noise of resistors. The system noise will affect these measurements less, so the accuracy of the measurements should be better than the approximately 0.5 % indicated by these measurements.

The theoretical and experimental prediction of the accuracy of the noise measurements are really not much different form one another. The major reason for the discrepancy is that in the theoretical calculation, the accuracy of the HP 3562A was not considered. The specifications for this device shows that it can only measure signals to a certain degree of accuracy. Although the percentages given are worst case values, one must assume that the HP 3562A reading will have some limit on the accuracy. This limit will probably be higher that the theoretical prediction of the noise.

The theoretical and experimental calculations of the accuracy of the noise set the range for where the true accuracy of the system will fall. Although the experimental prediction will be closer to the actual value because they inherently include the HP 3562A accuracy limit and follow all the measurement procedures. With all the data and predictions in mind, I expect the true accuracy of the measurement to be between about 0.5 %.

4.1.7 Recommendations for Improving Accuracy

Now that the limitations of the system have been identified and the accuracy of the system determined, are there any ways to improve the accuracy of the measurements? In this section an attempt will be made to answer this question. The answer will take the form of recommendations for improving the accuracy of the first system. Four recommendations will be presented.

First, some of the noisy components in the system could be replaced, reducing the extra noise injected to the system. Although the calibration procedure took care of this excess noise in the first measurement system, better performance could be achieved if some of the components were replaced with low noise components. Most of the noisy components are resistors. Since the noise produced by these devices is a thermal noise, the only way to reduce it is to operate at a lower temperature.

Second, it may be desirable to retune the circuit, especially the amplifier stages, to make the frequency response flat over different measurement bandwidths. This would eliminated the necessity of the correction-waveform, since the noise measurements would not be affected by the flat frequency response. So the noise measurement should be cleaner and maybe more accurate. Of course, the correction-waveform serves a dual role in the measurement system, correcting for the non-flat nature of the response and dividing the measurement by the gain of the circuit, so that the noise is referred back to the DUT node. So some type of gain measurement would still be needed to refer the signal back to the DUT node.

Third, it may be desirable to temperature control the entire circuit portion of the system. Temperature controls should keep all thermal noise sources, especially resistors, at a constant value and make it easier to predict a value for them and eliminated the value from all calculations, if the component is particularly troublesome. Temperature control should also cut down on the drift of the circuit components. That means there will be much more consistence in measurements.

Fourth, it may be desirable to take more averages in each measurement so that the time constant could be made longer. There is a limit to the number of averages one can select with the HP 3562A, but this value can be extended by averaging several multi-averaged measurements together. This is the method followed in the noise measurements for this system, when a series of ten, 10,000 average measurements were taken and averaged together resulting in a total of 1000,000 averages. This may not improve the accuracy that much because the measurements are still constrained by the HP 3562A measurement accuracy.

4.2 The Second Measurement System

4.2.1 Accuracy of Commercial Equipment

The Fluke 8506A Thermal True RMS Multimeter will be used in this system to make the noise, the bias current and bias voltage measurements for the system. Knowing the accuracy to which the Fluke 8506A can measure these quantities will be quite useful, especially in the case of the AC noise measurements.

All noise measurements made with the Fluke 8506A will be made with the high accuracy (hi accur) mode for AC voltage measurements. (For a discussion of the three AC measurement modes of the Fluke 8506A see section 3.2.3.) From the specifications of the Fluke 8506A, which can be found in Appendix D, the accuracy of the these measurements are dependent on the voltage and frequency range of the input signal. For an AC voltage in the 100 mV range the accuracy is ±(0.02% of reading + 5 counts) for a frequency between 40 Hz and 20 kHz, ±(0.04% of reading + 5 counts) for a frequency between 20 kHz and 50 kHz and ±(0.2% of reading + 0 counts) for a frequency between 50 kHz and 100 kHz. For an AC signal in the 300 mV range to 10 V range the accuracy is ±(0.012% of reading + 0 counts) for a frequency between 40 Hz and 20 kHz, ±(0.04% of reading + 0 counts) for a frequency between 20 kHz and 50 kHz and ±(0.2% + 0 counts) for a frequency between 50 kHz and 100 kHz. The term count was described in section 4.1.1.

The Fluke 8506A will also be used to make the bias voltage and current measurements necessary for the correlation part of the project. The current is measured across an 100 Ω resistor, so both the current and voltage measurements are of DC voltages. According to the specifications of the Fluke 8506A, it is capable of measuring signals in the 100 mV range to ±(0.0018% of reading + 15 counts) in the norm (normal) mode and to ±(0.0010% of reading + 8 counts) in the avg (average) mode. In the 1 V range it is capable of measuring signals to ±(0.0008% of reading + 7 counts) in the norm mode and to ±(0.0005% of reading + 4 counts) in the avg mode. The bias voltage and current measurements are usually made in the norm mode, therefore; the accuracy is ±(0.0018% of reading + 15 counts) for the 100 mV range and ±(0.0008% of reading + 7 counts) for the 1 V range.

The noise signal will be bandlimited by a filter with a bandpass from 1 kHz and 30 kHz, this range falls into two of the frequency ranges for the Fluke, so I will use the worst accuracy of the two ranges in determining the accuracy of the measurements made with this system. Thus the accuracy of the noise measurements will be ±(0.04% of reading + 0.5 μV) if the voltage is in the 100 mV range and ±(0.04% of reading) if the voltage is in the 300 mV to 10 V range. The accuracy of the bias current and voltage measurements will be taken as ±(0.0018% of reading + 1.5 μV) in the 100 mV range and ±(0.0008% of reading + 7 μV) in the 1 V range.

4.2.2 Analysis of the Circuit Portion of the System

The circuit portion of the second measurements system consists of the circuit of the first system, containing the bias circuit and the amplification stages, and a filter. The circuit of the first system was modeled and analyzed earlier in this section, see section 4.1.2, and will not be discussed again. The filter will be modeled and analyzed in this section, to identify any noisy components or sections.

The filter circuit actually contains two gain stages along with the actual bandpass filter. To determine the noise of the whole circuit, each stage could be modeled separately or the whole circuit could be modeled. However, neither the gain stage nor the filter will be modeled in this section, because the noise signal that will reach them will be significantly larger than any noise they themselves can produce. Tracing the signal through a model will just confirm this.

Even though we know the filter will not add a significant amount of noise to the signal, it still will add some amount. To determine this noise contribution of the filter, we will actually measure the noise of the filter using the HP 3562A. This filter noise measurement, with the input of the filter shorted and the second gain stage set at 6, may be seen in Figure 4.2.2.1. Looking at this noise spectrum one will see two peaks. These peaks are produced by two stages of the filter. The peak at 1 kHz corresponds to stage 2 of the filter, while the peak at 30 kHz corresponds to stage 1 of the filter. recalling that the stages are implemented in reverse order, these are the last two stages of the filter. The value of the noise may be obtained from the integral of the spectral density. The result is 133.959 e-9 V^2. This value would be approximately nine times as large

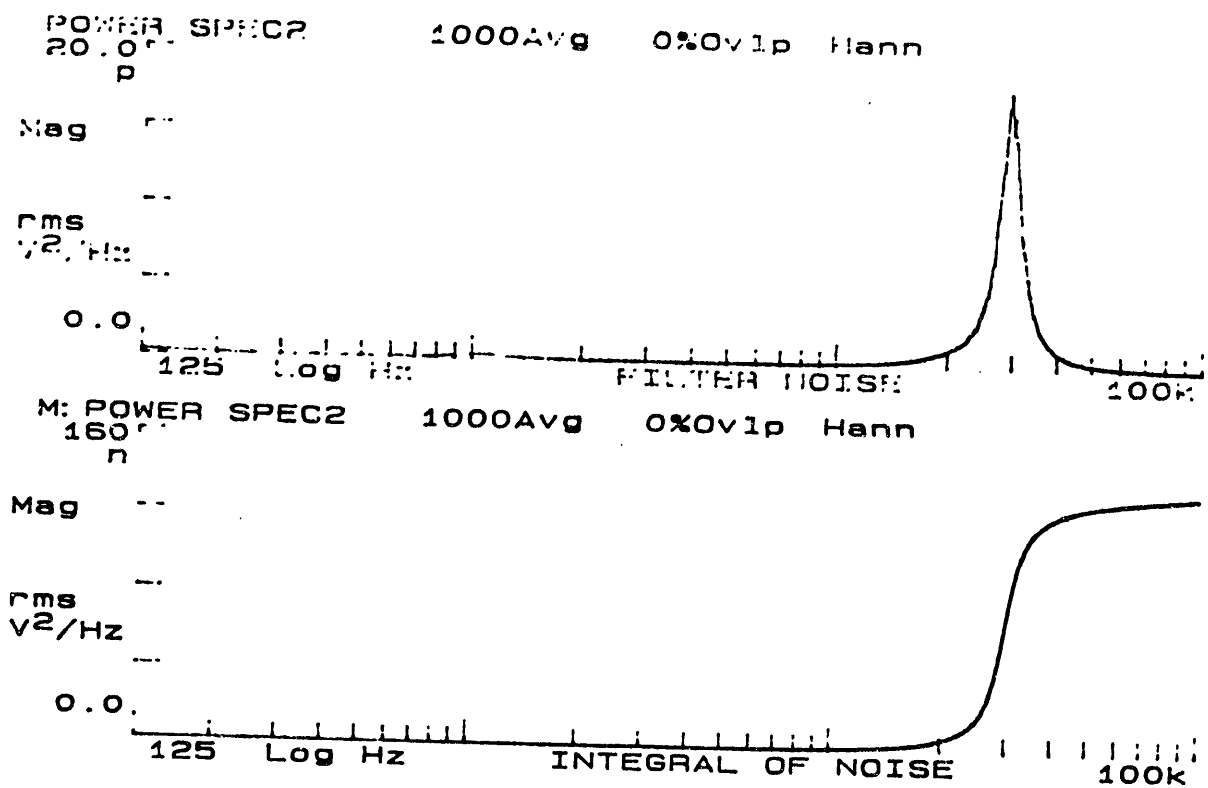


Figure 4.2.2.1 Noise Spectral Density of Filter

if the last gain stage was set for a gain of 18.

Since the filter will add some noise to the noise signal to be measured by the Fluke 8506A, it will be a concern. To insure that the calculated noise values contain only data for the DUT, all extra noise contributed by the system will have to be removed. Thus the noise of the filter will have to be eliminated from all measured noise quantities. The way in which this is done is described in the next section.

4.2.3 Analysis of the Calibration and the Measurement Procedures

The calibration and measurement procedures used with the second measurement system were described in section 3.2.4. These procedures can be discussed in light of the common noise measurement methods, discussed in section 2.3. However, due to the similarities of the procedures for this system and the first system, the discussion in section 4.1.3 applies to the second system as well, and will not be repeated here. This section will review the calibration and measurements procedures for this second system and determine if they lead to accurate measurements.

In the previous section, it was determined that the filter added some additional noise to the system. From section 4.1.2, we know that the bias/amplification circuit added some noise to the system. The major purpose of the calibration procedure is to characterize the noise of the system, so it can be eliminated from the measurements made with the system.

The calibration procedure for this system, like that of the first system, makes gain and noise measurements on a set of calibration resistors. The data from these measurements are used to calculate six calibration constants. Three of the constants, A, B, and C, are used to calculate the noise of the system. The data used in calculating these constants took into consideration the noise and gain of both the bias/amplification circuit and the filter. So, when this calculated noise is subtracted from the measured noise, the system noise is eliminated from this measured noise and the calibration procedure achieves its goal.

The measurement procedure for this system will be appropriate for the system if, it coordinates with the calibration procedure and it makes all measurements that are required for the project. The first criterion is met by the procedure, since the noise and gain measurements are made in the same manner as those in the calibration procedure and appropriate calibration constants are used with the data. The second criterion is met because bias voltage, bias current, noise and gain from CAL input to the OUTPUT terminal are all measured. These quantities are needed directly or indirectly in the project to determine correlation between noise and radiation characteristics of the DUT.

4.2.4 Discussion of the Effect of Sampling Time and Averaging on Accuracy

In section 4.1.4, it was shown that sampling time and averaging effect the accuracy of noise measurements. Equations 4.1.4.1 and 4.1.4.2, describe the relationship between the noise equivalent bandwidth of a system,the time constant of the meter used to measure the noise,and measurement accuracy. The time constant for our measurements is replaced by total sampling time per measurement. The effective time constant equals the number of samples to be averaged multiplied by the time it takes for one sample.

To calculate the accuracy of the second system, we first must know which accuracy equation to use. The Fluke 8506A, like the HP 3562A, is an averaging meter. Therefore, equation 4.1.4.2, will be used to calculate the accuracy.

To calculate the accuracy of the second measurement system, the noise equivalent bandwidth, the time per sample of the Fluke 8506A and the number of averages selected must be known. The noise equivalent bandwidth of the second system is set by the bandwidth of the filter. The filter ranges from 1 kHz to 30 kHz, for a bandwidth of 29 kHz. Thus the noise equivalent bandwidth of the system is 29 kHz. The Fluke 8506A will be used to make noise measurements. These measurement will be conducted in the AC high accuracy (hi accur) mode, where the signal being measured is sampled for 3.5 s. So the time constant for one measurement is 3.5 s. The number of measurements to be averaged is selected by the operator in the sys22.bas program, see Appendix C, and can range from 1 to 1000. The effective time constant for the system can range from 3.5 s to 3500 s, depending on the number of measurements to be averaged that is selected.

4.2.5 Summary of the Limitations on the Accuracy of the System

In the preceding four sections, possible limitations to the accuracy of noise measurements make with this system were identified. These limitations will be summarized and additional constraints to the accuracy will be discussed in the section. The limitations will be ranked to show which affects the accuracy of the noise the most.

One of the major limitations to the accuracy of the second measurement system is the frequency range of the filter. The filter had cutoff frequencies at 1 kHz and 30 kHz, for a bandwidth of 29 kHz. The real problem with the filter is its low cutoff frequency of 1 kHz. Most of the devices being measured by this system still have 1/f noise at 1 kHz, so the filter allows the 1/f noise to reach the meter reading the noise. That means that the large 1/f noise is averaged with the smaller broadband noise. The value of the noise measured by the meter will be dominated by the 1/f noise and not reflect the average value of the broadband noise, which is desired.

Even if the filter's frequency range did not allow 1/f noise into the measurement, the bandwidth of the filter would still limit the accuracy of the measurements. The bandwidth of the filter sets the noise equivalent bandwidth of the system as a whole. This quantity is one of the variables in the accuracy equation. If this quantity is fixed, one of the means of improving accuracy is lost.

Another concern involving the filter used in this system is its stability. Drift in the components used to build the filter is the cause of instability. This drift may cause changes in the bandwidth and in the gain of the filter. These changes could easily cause inaccuracies in the measurements because the calibration constants and gain that was measured before are not valid now.

In section 4.2.2, it was determined that the filter contributed some noise to the system. The noisy components and sections were those found in the circuit portion of this system also add to the system noise. Section 4.2.3 showed that the calibration procedure adequately accounts for those noise sources. The noise contributed by these sources is subtracted from all measurements though the predicted output noise term. So, the noisy components and system sections are not a limitation to the accuracy of the system, if the noise is stationary.

The specifications for the Fluke 8506A present several limitations to the accuracy of this system. First, the Fluke 8506A can only measure the noise signal to ±(0.04 % of reading + 0.5 μV), if the voltage is in the 100 mV range, and ±(0.04 % of reading), if the voltage is in the 300 mV to 10 V range. These are worst case errors, so actually measurements will have less error. The Fluke 8506A has a fixed sampling time of 3.5 s in the high accuracy mode. This time can be extended by averaging a number of readings. This can be easily done by using the sys22.bas program to control the measurements. The total sampling time will still be fixed at some level. Since the time is the other variable that constrains the accuracy, the accuracy of this system is definitely limited.

After considering all these limitations, I have found that some are more important than others. The most important limitation is the low frequency cutoff of the bandpass filter. This frequency is low enough that 1/f noise is entering into the measurements and causing errors. If this problem was eliminated by changing the frequency range of the filter, the noise equivalent bandwidth and effective time constant of the system, the stability of the system and the accuracy of the Fluke 8506A measurements would all limit the accuracy.

4.2.6 The Accuracy of the Second Measurement System

The accuracy of the second measurement system could be determined both theoretically and experimentally. For this system, the accuracy will only be determined theoretically. The experimental determination will not be feasible for this system until the filter is changed and 1/f noise is eliminated from the measurements. Even with the 1/f noise concerns taken care of, the experimental determination may not be a true test of the accuracy for several reasons. Thermal noise of a calibration resistor is smaller than the noise of the diode that the system will be usually used to measure. The calibration constants are determined using some of the resistors used to determine the accuracy, so the accuracy experiment may be more of a test of stability of the system than the actual accuracy. However, the 49 Ω and 499 Ω resistors were not used in the calibration and may give some indication of the accuracy.

A theoretical prediction of the noise may be obtained by using the accuracy equation 4.1.4.2. The time constant of the meter and the noise equivalent bandwidth of the system are the two quantities needed to calculate the accuracy of measurements. As shown in section 4.2.4, both these quantities are known for this system. The noise measurements made with the system, average 100 Fluke 8506A measurements of the noise. So the effective time constant is 350 s. The bandwidth of the second system is dependent on the bandwidth of the filter and is 29 kHz. Using these quantities in this equation 4.1.4.2, the accuracy of the system is

$$\sqrt{\alpha} = \frac{1}{\sqrt{(350\ \mathrm{s})(29\ \mathrm{kHz})}}$$

$$\sqrt{\alpha} = 3.14\ \mathrm{e}\text{-}4 * 100 = 0.032\% \qquad (4.1.6.1)$$

The actual accuracy of the system will probably be larger than 0.032% because the measurement accuracy of the Fluke 8506A is not considered. If a comparison is make between this accuracy and that of the first system, one will note that this theoretical accuracy is better than that found for the first system. The Fluke 8506A can also measure an AC signal more accurately than the HP 3562A. In light of these facts, I estimate that the second system will have the same if not slightly better accuracy than the first system. The estimated accuracy of this system then should be between one-half to one percent.

4.2.7 Recommendations for Improving Accuracy

The limitations to the accuracy of this second system, along with the accuracy of the second system have been determined in the preceding sections. This section will look at ways some of these limitations may be overcome or minimized.

The first way the accuracy of the second system can be improved is to change the filter used with the system. The filter bandwidth should be moved to higher frequencies. The lower cutoff frequency must be moved from 1 kHz to a higher frequency, so that the 1/f noise of the devices will be eliminated from the signal being measured. The bandwidth of the old filter is reasonable at 29 kHz, so a change is not necessary. However, changing the bandwidth slightly, as long as amplifiers operate well in the range, will not affect the noise measurement accuracy. It may be desirable to change the gain of the passband of the filter to unity. The old filter attenuates signals so that extra gain stages are needed to boost the signal back up to a reasonable level.

Actually, at this moment a new filter, is being built for this system. It is a fifth order Chebyshev bandpass filter with frequency cutoffs at 50 kHz and 80 kHz, for a bandwidth of 30 kHz. The gain of the passband is approximately unity, but with ± 0.3 dB ripple. The gain stages used with the old filter will have to be changed to account for the change in the gain of the filter.

Two of the recommendations mentioned in section 4.1.7 could also apply to the second system, since this system utilizes a portion of the first system. First, noisy components could be replaced in both the bias/amplification circuit and the filter. Second, temperature control of the circuit and filter should fix thermal noise sources at constant values and cut down on drift of the components. This suggestion is particularly good for improving the stability of the filter and keeping the gain and bandwidth of it constant.

Section 5 GENERALIZED DISCUSSION OF THE ACCURACY OF NOISE MEASUREMENTS

5.1 Common Limitations to Accuracy and Ways to Improve Accuracy

In the analysis for the two noise measurement systems, many limitations to accuracy were identified. Several of these limitations would be encountered in any noise measurement system. These include 1/f noise, bandlimiting, sampling time, and added noise from components and equipment. A designer of a noise measurement system should be aware of these limits so he will know what to look at if problems arise and where care must be taken in designing the system.

In the last section, several recommendations were made for improving the accuracy of the two systems analyzed in this thesis. These recommendations address the specific limits of these systems, but the recommendations could be applied to another system if similar limits exist. Temperature control, additional averaging to extend sampling time and some expansion of the noise equivalent bandwidth will be the most effective of the recommendations mentioned.

5.2 Estimate of How Accurately Noise May Be Measured

After evaluating the accuracy of the two noise measurements systems, I have concluded that one may not make general statements about the accuracy of noise measurements. While there are common limits to accuracy, shared by most noise measurement systems, there are still many other limits to accuracy that are inherent to a particular system. The way a system makes a noise measurement, the way it is calibrated, the type of equipment used in the system, are just several things which are unique to system and which may limit the accuracy of that particular system.

Considering the number of limitations to the accuracy of measurements that are specific to a measurement system, it is impossible to predict a numerical answer for the best accuracy that can be achieved. The best answer that may be given, is that one may expect to achieve noise measurements, that are accurate to several percent, as long as certain steps are taken to achieve good accuracy. These step include designing the system to avoid or minimize common limits to accuracy, identifying limitations, specific to a measurement system, which degrade the accuracy and minimizing these limitations, that are specific to a measurement system, as much as possible.

This thesis has described two noise measurement systems where steps like these were taken and where reasonable accuracy was achieved. Anyone trying to make accurate noise measurements could use similar measurement systems and/or follow the recommendations made in this thesis for improving measurement accuracy.

Section 6 CONCLUSIONS AND RECOMMENDATIONS FOR FURTHER STUDY

6.1 Conclusions

This thesis has carefully analyzed two different systems for measuring noise in voltage reference diodes. The limitations to the accuracy of each of these systems were identified and recommendations were made for improving the accuracy.

The accuracy of the two systems was found theoretically and the accuracy of the first system was also found experimentally. The accuracy of the first system, which used the HP 3562A Dynamic Signal Analyzer to measure the noise, was determined to be about 0.5%. The accuracy of the second system which used the Fluke 8506A Thermal RMS Multimeter to measure the noise, was determined to be 0.5% or better. The accuracy of this second system is not as well known as the first, because measurements to predict accuracy could not be conducted because the filter used in the system allowed 1/f noise into the signal to be measured. Measurements of this signal would be suspect, because of the inclusion of 1/f noise.

Some measurements on commercial diodes were conducted with the first system to determine whether there is correlation between their noise and radiation characteristics. The noise measurements were accurate enough to distinguish from one another. No conclusions about the correlation with radiation characteristics were possible, because of problems with voltage measurements due to temperature deviations. Another set of diodes is about to be examined. Much more care will be taken with these measurements, including temperature control of the diodes. Hopefully correlation will be established from the data from these measurements.

It appears from these experiments that the first system has sufficient accuracy for the purposes of this project. The second system has the potential to have better accuracy than the first system, so it too will have sufficient accuracy for this project.

Also included in this thesis, was a more generalized discussion of noise measurements. Common limitations and possible ways to improve accuracy were presented. The accuracy of an arbitrary noise measurements could not be estimated, because accuracy is dependent upon the system being used to measure the noise. If one is careful in developing the system and making measurements, they should expect accuracy of several percent.

6.2 Recommendations for Further Study

An immediate continuation of the work begun by this thesis is a further study of the second measurement system. Several factors, like the Fluke 8506A having better accuracy than the HP 3562A and the longer effective time constant for the second system, make me believe that the second system will be more accurate. Experiments, even a rough one which measures the noise of resistors, must be conducted to see if this assumption is correct. These tests cannot be conducted until the new filter is completed. This new filter is necessary to eliminate 1/f noise from the noise signal to be measured.

Further exploration of the topic of accurate noise measurements could take the form of developing more accurate noise measurement systems. This topic could be further divided into systems to measure noise in diodes and systems to measure noise in other devices. Some of the recommendations made in this thesis for improving accuracy could be followed to develop a more accurate measurement system for voltage reference diodes. The technology discussed in this thesis could possibly be modified to measure other types of devices. Transistors are the likely choice for extension of this existing measurements system. Further work would be necessary for building a system to measure noise in ICs. Measurements systems for ICs would probably be device orientated, considering the wide variety of ICs.

The two systems discussed in this system were analog in nature, with the exception of the meters used to measure the noise. It is possible to build a digital system to measure noise. Actually the system would be a mixed analog and digital system, since you still need analog circuitry for biasing the DUT, amplification of the noise signal and anti-alias filtering. According to a paper by D. Rod White [15], the anti-aliasing filter used with this system can be quite simple, even single pole, because aliasing does not bias noise power measurements, it just increases the variance of the measurements. After the signal passes through the anti-aliasing filter, it is sampled by an A/D. Then it can be processed by a digital signal processing circuit of your design. For examples of such DSP systems, see the paper by W. Rod White [15]. The major limitations of this new system will be the gain of the noise amplifier, the accuracy of the A/D and bandwidth limitations caused by limitations to the computational power of the computer.

Appendix A. GLOSSARY OF NOISE RELATED TERMS

1/f NOISE OR LOW FREQUENCY NOISE

A type of noise whose spectral density increases without limit as frequency decreases. The causes of 1/f noise are matters of controversy, one cause of 1/f noise in semiconductor devices is traceable to properties of the surface of the material.
The generation and recombination of carriers in surface energy states and the density of surface states are important factors. [2]

NOISE

"Any unwanted disturbance that obscures or interferes with a desired signal." [2:7] But for out purposes, noise is a random signal that we believe can provide insight into the operation of reference diodes.

NOISE BANDWIDTH or NOISE EQUIVALENT BANDWIDTH

"The frequency span of a rectangularly shaped power gain curve equal in area to the area of the actual power gain versus frequency curve." [2:302] (See Figure A.1.)

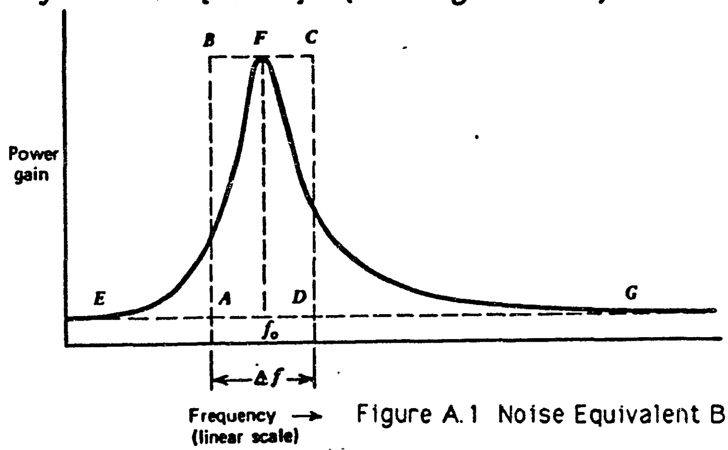


Figure A.1 Noise Equivalent Bandwidth

Noise bandwidth in equation form is represented in the following manner:

$$\Delta f = \frac{1}{G_o}\int_{-\infty}^{\infty} G(f)\, df \qquad (A.1)$$

$G(f)$ = power gain as a function of frequency

G_o = peak power gain

Alternatively, noise bandwidth may be represented in terms of voltage again squared instead of power gain.

$$\Delta f = \frac{1}{A_{vo}^2}\int_{-\infty}^{\infty} [A_v^2(f)]\, df \qquad (A.2)$$

$A_v(f)$ = voltage gain as a function of frequency

A_{vo} = midband voltage gain

NOISE RATIO

A normalization of the diode noise current spectral density, S_{id}, with respect to the noise current spectral density of a saturated themionic diode conducting a DC current equal to the p-n junction reverse current. [1] In equation form noise ratio is:

$$NR = \frac{S_{id}}{2qI_r} \qquad (A.3)$$

S_{id} = noise current spectral density

I_r = diode reverse current

SHOT NOISE

Noise found in tubes, transistors, and diodes caused by the fact that current flowing in this devices is not smooth and continuous but rather it is the sum of pulses of current caused by the flow of carriers, each carrying one electronic charge. [2] Shot noise is described by the following equation:

$$I_{sh} = \sqrt{2qI_r\Delta f} \qquad (A.4)$$

q = electronic charge 1.6 e-19

I_r = reverse current

Δf = noise bandwidth in Hz

SPECTRAL DENSITY

"Term used to describe the noise content in a unit of bandwidth." [2]

THERMAL NOISE

"Noise caused by the random thermally excited vibration of the charge carriers in a conductor." [2] Thermal rms (root-mean-square) noise voltage has the following value:

$$E_n = \sqrt{4kTR\Delta f} \qquad (A.5)$$

k = Boltzmann's constant = 1.38e-23 W-sec/°K

T = temperature of conductor in degrees Kelvin (°K)

R = resistance or the real part of the conductor's impedance

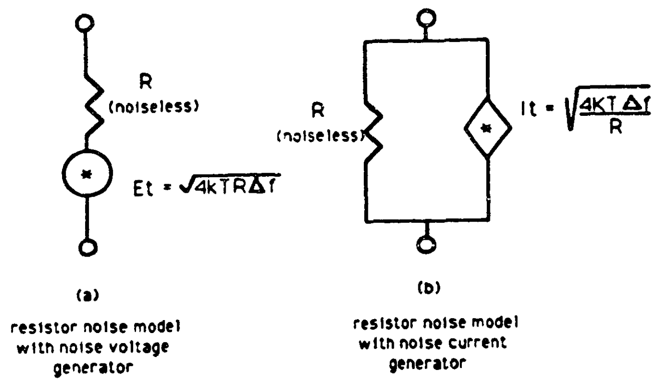
Thermal rms current voltage has the following value:

$$I_n = \sqrt{\frac{4kT\Delta f}{R}} \qquad (A.6)$$

where k, T and R are the same as (A.5)

Appendix B. NOISE MODELS

Resistor: A resistor can exhibit both thermal and 1/f noise. However, the noise model for the resistor only shows the thermal noise. If one is working at very low frequencies, 1/f noise must be added to the model. The model for a resistor consists of a noiseless resistor, with the same value as the resistor being modeled, in series with a thermal voltage noise generator (See Figure B.1 (a)) or in parallel with a thermal current noise generator (See Figure B.1 (b)).



Source: [2:16]

Figure B.1 Resistor Noise Models

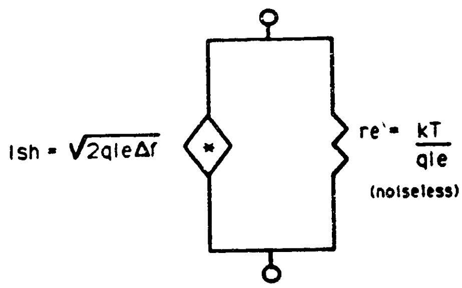
Diode: A diode can exhibit both shot and 1/f noise. As with the resistor, the noise model does not contain 1/f noise. So, if one is working at low frequencies, 1/f noise must be added to the model. There are two diode noise models, one for forward-biased diodes and one for reverse-biased diodes. The forward-biased diode noise model may be seen in Figure B.2 (a). The model consists of a shot noise generator in parallel with a resistor. This resistor is sometimes referred to as the Schockley emitter resistance. It is equal to kT/qI_e and is the reciprocal of the conductance obtained by differentiating the diode equation

$$I_e = I_o(\exp(qV_{BE}/kT) - 1) \qquad \text{(B.1)}$$

I_e = diode current

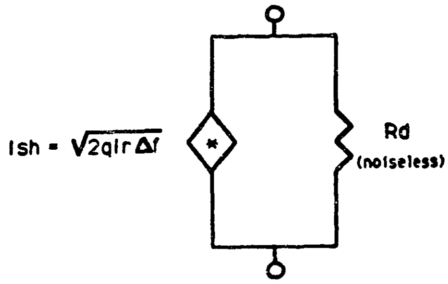
I_o = saturated value of reverse current

V_{BE} = potential across the p-n junction of the diode with respect to V_{BE}.



Source (2:22)
Figure B.2 (a) Noise Model For A Forward-Biased Diode

The reverse-biased diode noise model may be seen in Figure B.2 (b). This model consists of a shot noise generator and a measured resistance, R_d. The current used to calculate this shot noise is the reverse bias current of the diode.



Source: [2:22]
Figure B.2 (b) Noise Model For A Reverse-Biased Diode

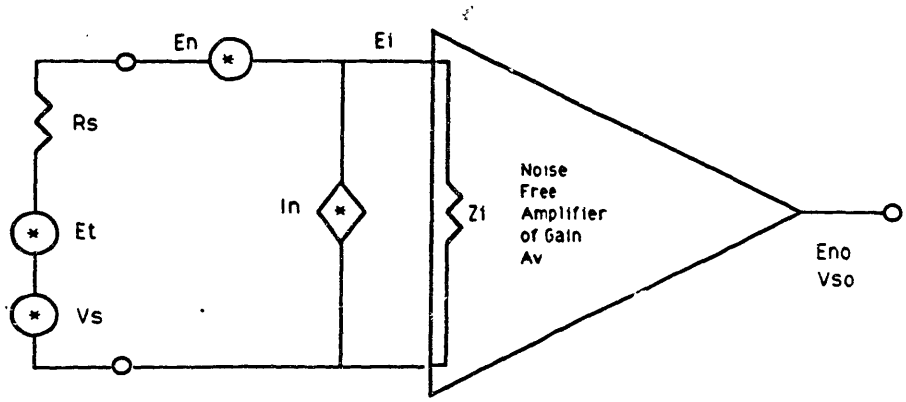
Amplifier: The amplifier model is usually used to model the noise of an operational amplifier, but it could be used for other types of amplifiers. The model for the amplifier consists of a current noise generator, a voltage noise generator and a noise free amplifier with a gain equal to the gain of the amplifier being modeled. This model may be seen in Figure B.3. This model has some additional parts: R_s, the impedance of the input to the amplifier; E_t, the total noise of the circuitry that comes before the amplifier; and V_s, the input of the amplifier. The actual noise model is the section to the right of the open circles. The input impedance of the amplifier, Z_i, is also included in the model and is used to determine the system gain. The value of the current and voltage generators are usually given in the specification sheets for the operational amplifier. If such data is not included, there are methods and test circuits to measure the value of these generators. Using this model the equivalent input noise, E_{ni}^2, of the amplifier is equal to:

$$E_{ni}^2 = E_t^2 + E_n^2 + I_n^2 R_s^2 \qquad (B.2)$$

The equivalent output noise, E_{no}^2 is equal to:

$$E_{no}^2 = E_{ni}^2 K_t^2 \qquad (B.3)$$

where $K_t = A_v Z_i/(R_s + Z_i)$, the system gain.



Source: [2.30]

Figure B.3 Amplifier Noise Model

Appendix C. COMPUTER PROGRAMS

This appendix contains the computer programs that were used with the two noise measurement systems described in this thesis. The programs appear in the same order that they are mentioned in the thesis.

Cal.bas

```
1 'The following program was written by Sumner Brown and calculates
   the calibration constants to be used in the analysis of data
   obtained by the noise measurement system.
2 'The first three constants, Kr, Cr, and G are related by the
   following equation:
   Gi * Cr - GAINi * G + Kr + G * Cr = Gi * GAINi.  Where Gi
   is the admittance of the DUT and bias resistors and GAINi is the
   measured gain of the system, for i = 1, 2, and 3.
3 'The second set of constants, A, B, and C are related by the
   following equation:  eo^2 = A + B + RTi + C * (RTi)^2.  Where eo is
   the noise measured at the output of the system but referred to the
   DUT node and RTi is the equivalent single resistance in the DUT
   position, for i = 1, 2, and 3.
4 'For both sets of constants, three sets of data are used to form
   three equations.  These three equations are solved simultaneously
   using Cramer's Rule for the constants.
5 '
99 'Line 100 initializes all arrays to be used in this program.
    Arrays R#, GN#, and ESI# are used to store the input data.  All
    other arrays are intermediate arrays used in calculations.
100 DIM L1#(2), L2#(2), L3#(2), R#(2), GN#(2), ES#(2), RI#(2)
298 '
299 'Lines 300-380 ask for three sets of input data.
300 INPUT "R1";R#(0)
```

```
310 INPUT "GAIN 1";GN#(0)

320 INPUT "NOISE 1";ES#(0)

330 INPUT "R2";R#(1)

340 INPUT "GAIN 2";GN#(1)

350 INPUT "NOISE 2";ES#(1)

360 INPUT "R3";R#(2)

370 INPUT "GAIN 3";GN#(2)

380 INPUT "NOISE 3";ES#(2)

498 '

499 'Lines 500-660 calculate Kr, Cr, and G.

500 FOR J=0 TO 2: RI#(J)=1/R#(J): NEXT

550 FOR J=0 TO 2: L1#(J)=RI#(J): L2#(J)=GN#(J): L3#(J)=1: NEXT

560 GOSUB 1000

570 D#=DET#

580 FOR J=0 TO 2: L1#(J)=RI#(J)*GN#(J): NEXT

590 GOSUB 1000

600 CR#=DET#/D#

610 FOR J=0 TO 2: L2#(J)=L1#(J): L1#(J)=RI#(J): NEXT

620 GOSUB 1000

630 G#=-DET#/D#

640 FOR J=0 TO 2: L3#(J)=L2#(J): L2#(J)=GN#(J): NEXT

650 GOSUB 1000

660 KR#=DET#/D#-G#*KR#

698 '

699 'Lines 700-810 calculate A, B, and C.

700 FOR J=0 TO 2: L1#(J)=1: L2#(J)=(GN#(J)-CR#)/KR#:
```

```
    L3#(J)=L2#(J)*L2#(J):NEXT
710 GOSUB 1000
720 D#=DET#
730 FOR J=0 TO 2: L1#(J)=ES#(J): NEXT
740 GOSUB 1000
750 A#=DET#/D#
760 FOR J=0 TO 2: L1#(J)=1: L2#(J)=ES#(J): NEXT
770 GOSUB 1000
780 B#=DET#/D#
790 FOR J=0 TO 2: L2#(J)=(GN#(J)-CR#)/KR#: L3#(J)=ES#(J): NEXT
798 '
799 'The values of the constants are returned.
800 GOSUB 1000
810 C#=DET#/D#
900 PRINT "CR =" CR#
910 PRINT "G ="G#
920 PRINT "KR =" KR#
930 PRINT "A =" A#
940 PRINT "B =" B#
950 PRINT "C =" C#
990 END
998 '
999 'The following subroutine calculates the determinant of a three by
     three matrix
1000 DET#=L1#(0)*(L2#(1)*L3#(2)-L2#(2)*L3#(1))-L2#(0)*(L1#(1)*L3#(2)
     -L3#(1)*L1#(2))+L3#(0)*(L1#(1)*L2#(2)-L1#(2)*L2#(1))
```

```
1010 RETURN
```

Sys22.bas

```
10 'SYS22.BAS
20 'JUDY FURLONG
30 'DECEMBER 20, 1989
40 REM- THIS PROGRAM TAKES NOISE MEASUREMENTS FROM THE FLUKE 8506A
50 CLS
60 KEY OFF
70 DIM VRMS(1000)
80 N=0:SUM = 0
90 GOSUB 2000 'Initialization of GRIB Board
100 GOSUB 2500 'Triggering
110 GOSUB 3000 'Initialization of Fluke
120 FOR T= 1 TO 3000: NEXT T
130 CLS
140 GOSUB 3500 'Initial Input
150 GOSUB 4000 'Taking a Reading
160 GOSUB 6500 'PRINT AVE AND STD
180 END
200 '
2000 'Initialization of the GRIB Board
2010 DEF SEG =&HC000
2020 REM-IBM Address = 3, Controller in Charge = 1, Number of GRIB
         Boards = 1, Base Address = &H300
2030 CMD$ = "SYSCON MAD = 3, CIC = 1, NOB = 1, BA0 = &H300"
2040 IE488 = 0
```

```
2050 A% = 0: FLG% = 0: BRD% = 0: RD$ = SPACE$ (20)

2060 CALL IE488 (CMD$, A%, FLG%, BRD%)

2070 IF NOT FLG% THEN 2090

2080 PRINT "MetroByte Error:"; HEX$ (FLG%): END

2090 CLS

2100 PRINT "SYSTEM GRIB SETUP COMPLETED"

2110 FOR T = 1 TO 1000: NEXT T

2120 CLS

2130 RETURN

2140 '

2500 'Triggering

2510 CMD$ = "REMOTE 17": TRG$ = "TRIGGER 17"

2520 CALL IE488 (CMD$, A%, FLG%, BRD%)

2530 CMD$ = "TIMEOUT 15"

2540 CALL IE488 (CMD$, A%, FLG%, BRD%)

2550 FOR T= 1 TO 500:NEXT T

2560 RETURN

2570 '

3000 'Initialization of Fluke

3010 FLKIN$ = "ENTER 17[$]": FLKOUT$ = "OUTPUT 17[$+]": MSF$ = "?":
     RDF$ = SPACE$(20)

3020 X$ = "*"

3030 CALL IE488 (FLKOUT$, X$, FLG%, BRD%)

3040 FOR I = 1 TO 3000: NEXT I: BEEP

3050 X$ = "DOLOBOVA2R," 'DO LO BO = DISPLAY ON, PANEL ON, ASCII OUTPUT
     VA2 = AC HIGH ACCURACY VOLTAGE, AUTO RANGE
```

```
3060 CALL IE488 (FLKOUT$, X$, FLG%, BRD%)
3070 IF FLG% THEN PRINT "Fluke Error, MetroByte Code: ";HEX$(FLG%):
     GOTO 3020 ELSE PRINT "Fluke DVM Initialized."
3080 RETURN
3090 '
3500 'Initial Input
3520 INPUT "Please enter today's date. (In mo/day/yr form): ", DA$:
     PRINT
3530 INPUT "Please enter the DUT's ID number. (8 chars. max):",
     DUTID$: PRINT
3540 INPUT "Please enter a filename. (8 chars. max):" , FLNAME$: PRINT
3550 ON ERROR GOTO 3580: OPEN FLNAME$+ ".A1" FOR INPUT AS #1
3560 CLOSE #1
3570 BEEP: PRINT "Do you want to overwrite the file "; FLNAME$+ ".A1";
     " ? (0/1):"             ;: INPUT "", X%: IF X% THEN KILL FLNAME$+
     ".A1":PRINT:GOTO 3590 ELSE PRINT: GOTO 3540
3580 RESUME 3590
3590 OPEN FLNAME$+ ".A1" FOR APPEND AS #1:PRINT #1, "*":PRINT #1, " N
     VRMS"
3600 CLOSE #1
3610 INPUT "Please enter the number of readings (1000 max):" ,K:PRINT
3620 CLS
3630 RETURN
3640 '
4000 'TAKING A READING
4010 CLS
```

```
4020 PRINT DA$; SPC(61); "ID "; DUTID$
4030 PRINT:PRINT:PRINT "N    VRMS":PRINT
4040 LPRINT DA$; SPC(61); "ID "; DUTID$
4050 LPRINT:LPRINT:LPRINT "N    VRMS":LPRINT
4060 CALL IE488 (TRG$, X%, FLG%, BRD%)
4070 FOR L = 1 TO 8000:NEXT L
4080 IF N < K THEN N = N + 1 ELSE RETURN
4090 CALL IE488 (TRG$, X%, FLG%, BRD%)
4100 FOR M = 1 TO 4000:NEXT M
4110 CALL IE488 (FLKIN$, RDF$, FLG%, BRD%)
4120 CALL IE488 (FLKIN$, RDF$, FLG%, BRD%)
4130 VRMS(N) = VAL(RDF$)
4140 GOSUB 4500 'SAVING DATA
4150 GOSUB 5000 'SCREEN PRINTOUT
4155 GOSUB 6000 'AVE
4160 GOTO 4080
4170 '
4500 'Saving Data
4510 F$ = " ##.######^^^^  "
4520 OPEN FLNAME$+ ".A1" FOR APPEND AS #1
4530 PRINT #1, N;:PRINT #1, USING F$; VRMS(N)
4550 CLOSE #1
4560 RETURN
4570 '
5000 'SCREEN PRINTOUT
5060 PRINT N;:PRINT USING F$; VRMS(N)
```

```
5090 'Output on Printer

5140 LPRINT N;:LPRINT USING F$; VRMS(N)

5170 RETURN

5180 '

6000 'AVE

6010 SUM = SUM + VRMS(N)

6020 RETURN

6030 '

6500 'PRINT AVE AND STD

6510 AVE = SUM / K

6520 PRINT "AVERAGE: ";:PRINT USING F$; AVE

6525 LPRINT "AVERAGE: ";:LPRINT USING F$; AVE

6530 T = 0

6540 FOR Z = 1 TO K

6550 T = T + (VRMS(Z) - AVE)^2

6560 NEXT Z

6570 B = K - 1

6580 D = T / B

6590 SIGMA = SQR(D)

6600 PRINT "STANDARD DEVIATION: ";:PRINT USING F$; SIGMA

6605 LPRINT "STANDARD DEVIATION: ";:LPRINT USING F$; SIGMA

6610 RETURN
```

Appendix D. HP 3562A AND FLUKE 8506A SPECIFICATIONS

This appendix contains the specifications for the Hewlett-Packard 3562A Dynamic Signal Analyzer. These specifications were found in the Operating Manual [10] for this piece of equipment. Also included in this appendix is the specifications for the Fluke 8506A Thermal True RMS Multimeter. These specification were found in the Instruction Manual [8] for this piece of equipment.

HP 3562A SPECIFICATIONS

Specifications describe the instrument's warranted performance. Supplemental characteristics are intended to provide information useful in applying the instrument by giving typical, but non-warranted, performance specifications. Supplemental characteristics are denoted as 'typical,' 'nominal,' or 'approximately.'

Frequency

Measurement Range: 64 µHz to 100 kHz, both channels, single- or dual-channel operation

Accuracy: ±0.004% of frequency reading

Resolution: Span/800, both channels, single- or dual-channel operation, Linear Resolution mode

Spans:	Baseband	Zoom
# of spans	66	65
min span	10.24 mHz	20.48 mHz
max span	100 kHz	100 kHz
time record (Sec)	800/span	800/span

Window Functions: Flat Top, Hann, Uniform, Force, Exponential and User-Defined

Window Parameters:	Flat Top	Hann	Uniform
Noise Equiv BW (% of span)	0.478	0.188	0.125
3 dB BW (% of span)	0.45	0.185	0.125
Shape factor (60 dB BW/3 dB BW)	2.6	9.1	716

Typical Real Time Bandwidths:

Single-channel, single display	2.5 kHz
Single-channel, Fast Averaging	10 kHz
Dual-channel, single display	2 kHz
Dual-channel, Fast Averaging	5 kHz
Throughput to CS/80 disc	
Single-channel	10 kHz
Dual-channel	5 kHz

Amplitude

Accuracy: Defined as Full Scale Accuracy at any of the 801 calculated frequency points. Overall accuracy is the sum of absolute accuracy, window flatness and noise level.

Absolute Accuracy:
Single Channel (Channel 1 or Channel 2)
±0.15 dB ±0.015% of input range (+27 dBV to −40 dBV, input connections as specified in Cases 1 and 2 below)
±0.25 dB ±0.025% of input range (−41 dBV to −51 dBV, input connections as specified in Cases 1 and 2 below)

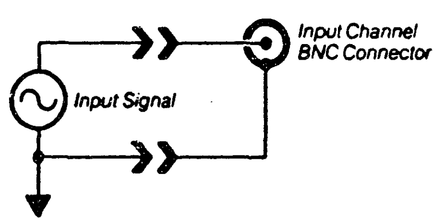
DC Response: Auto-Cal and Auto-Zero on

Input Range (dBV rms)	dc Level
+27 to −35	>30 dB below full scale
−36 to −51	>20 dB below ful scale

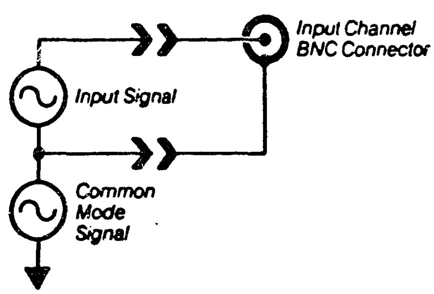
Frequency Response Channel Match:
±0.1 dB, ±0.5 degree (input connections as specified in Cases 1 and 2 below)

Input Connections:
Cases 1 and 2 are the recommended input connections. For these cases, the amplitude accuracy specified above is applicable.

Case 1



Case 2



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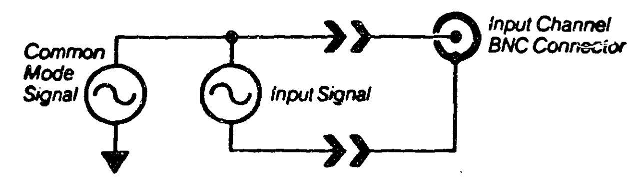
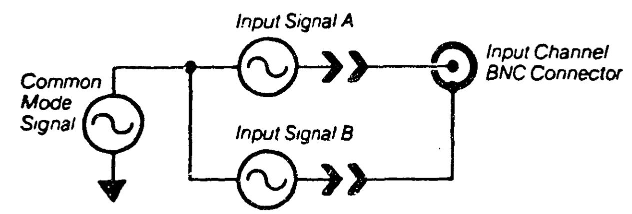
Cases 3 and 4 are input connections which degrade amplitude accuracy. For these cases, the amplitude accuracy specified above must be modified with the accuracy adders stated below.

Case 3



Case 4



Accuracy Adder: Single-channel, inputs connected as shown in Cases 3 and 4 above.
Add ±0.35 dB and ±4.0 degrees to the absolute accuracy.

Accuracy Adder: Dual-channel measurements
Add ±0.35 dB and ±4.0 degrees once for each input connected as shown in Cases 3 and 4 above.

Window Flatness:

Flat Top:	+0, −0.01 dB
Hann:	+0, −1.5 dB
Uniform:	+0, −4.0

Noise Floor: Flat top window, 50 Ω source impedance, −51 dBV range
20 Hz to 1 kHz (1 kHz span) < −126 dBV (−134 dBV/√Hz)
1 kHz to 100 kHz (100 kHz span) < −116 dBV (−144 dBV/√Hz)

Dynamic Range: All distortion (intermodulation and harmonic), spurious and alias products ≥80 dB below full scale input range (16 averages) < 10KΩ termination

Phase

Accuracy: Single channel, input connections as specified above in Cases 1 and 2, referenced to trigger point.

<10 kHz	±2.5 degrees
10 kHz to 100 kHz	±12.0 degrees

Inputs

Input Impedance: 1 MΩ ±5% shunted by <100 pF

Input Coupling: The inputs may be ac or dc coupled; ac rolloff is <3 dB at 1 Hz

Crosstalk: −140 dB (50 Ω source, 50 Ω input termination, input connectors shielded)

Common Mode Rejection:

0 Hz to 66 Hz	80 dB
66 Hz to 500 Hz	65 dB

Common Mode Voltage: dc to 500 Hz

Input Range (dBV rms)	Maximum (ac + dc)
+27 to −12	±42.0 Vpeak
−13 to −51	±18.0 Vpeak*

*For the −43 to −51 dBV input ranges, common mode signal levels cannot exceed ±18 Vpeak or (Input Range) + (Common Mode Rejection), whichever is the lesser level.

Common Mode Voltage: 500 Hz to 100 kHz. The ac part of the signal is limited to 42 Vpeak or (Input Range) + (10 dB), whichever is the lesser level.

Common Mode Distortion: For the levels specified, distortion of common mode signals will be less than the level of the rejected common mode signal.

External Trigger Input Impedance: Typically 50 kΩ ±5%

External Sampling Input: TTL compatible input for signals ≤256 kHz (nominal maximum sample rate).

External Reference Input:
Input Frequencies: 1, 2, 5 or 10 MHz ±0.01%
Amplitude Range: 0 dBm to +20 dBm (50 Ω)

Trigger

Trigger Modes: Free Run, Input Channel 1, Input Channel 2, Source and External Trigger. Free Run applies to all Measurement Modes; Input Channel 1, Input Channel 2, Source and External Trigger apply to the Linear Resolution, Time Capture and Time Throughput measurement modes.

Trigger Conditions:
Free Run: A new measurement is initiated by the completion of the previous measurement.

Input: A new measurement is initiated when the input signal to either Channel 1 or Channel 2 meets the specified trigger conditions. Trigger Level range is ± 100% of Full Scale Input Range; Trigger Level is user-selected in steps of (Input Range in volts)/128.

Source: Measurements are synchronized with the periodic signal types (burst random, sine chirp and burst chirp).

External: A new measurement is initiated by a signal applied to the front panel External Trigger input. Trigger Level range is ± 10 Vpeak; Trigger Level is user selected in 80 mV steps.

Trigger Delay:
Pre-Trigger: The measurement can be based on data from 1 to 4096 samples (1/2048 to 2 time records) prior to trigger conditions being met. Resolution is 1 sample (1/2048 of a time record).

Post-Trigger: The measurement is initiated from 1 to 65,536 samples (1/2048 to 32 time records) after the trigger conditions are met. Resolution is 1 sample (1/2048 of a time record).

Source

Band limited, band translated random noise, burst random, sine chirp, burst chirp, as well as fixed sine and swept sine signals are available from the front panel Source output. DC Offset is also user-selectable.

Output Impedance: 50 Ω (nominal)

Output Level: ≤ ± 10 Vpeak (ac + dc) into a ≥ 10 kΩ, < 1000 pF load. Maximum current = 50 mA.

AC Level: ± 5 Vpeak (≥ 10kΩ, < 1000 pF load)

DC Offset: ± 10 Vpeak in 100 mV steps. Residual offset at 0 V offset ≤ 10 mV.

% In-Band Energy: (1 kHz span, 5 kHz center frequency)
Random Noise: 70%
Sine Chirp: 85%

Accuracy and Purity: Fixed or Swept Sine
Flatness: ± 1 dB (0 to 65 kHz),
+ 1, − 1.5 dB (65 kHz to 100 kHz)

Distortion: (including subharmonics)

dc to 10 kHz	−60 dB
10 kHz to 100 kHz	−40 dB

General

Specifications apply when AUTO CAL is enabled, or within 5°C and 2 hrs of last internal calibration (except for transient environmental changes).

Ambient temperature: 0° to 55° C.
Relative Humidity: ≤ 95% at 40°C.
Altitude: 4,572 m (15,000 ft)

Storage:
Temperature: − 40° to + 75°C.
Altitude: ≤ 15,240m (50,000 ft)

Power:
115 VAC + 10% — 25%, 48 to 440 Hz
230 VAC + 10% — 25%, 48 to 66 Hz
450 VA maximum

Weight:
26 kg (56 lbs) net
35 kg (77 lbs) shipping

Dimensions:
222 mm (8.75 in) high
426 mm (16.75 in) wide
578 mm (22.75 in) deep

HP-IB:
Implementation of IEEE Std 488-1978
SH1 AH1 T5 TE0 L4 LE0 SR1 RL1 PP0 DC1 DT1 C0
Supports the 91XX and 794X families of HP disc drives as well as Hewlett-Packard Graphics Language (HP-GL) digital plotters.

Section 1
Introduction & Specifications

1-1. INTRODUCTION

1-2. This eight-section manual provides comprehensive information for installing, operating and maintaining your Fluke digital multimeter. Complete descriptions and instructions are included for the instrument mainframe, for all modules necessary in making thermal true-rms and dc volts measurements, and for any optional modules ordered with the instrument. Appropriate sectionalized information is included with any optional modules subsequently ordered and may be inserted in Section 6.

1-3. DESCRIPTION

1-4. The multimeter features 6-1/2 digit resolution, full annunciation and simplicity of operation. Modular construction, microprocessor control, and a bus structure provide excellent flexibility. Memory programming from the front panel (or through a remote interface) controls all measurement parameters, mathematical operations and special operations. The standard hardware configuration allows for measurement of thermal true-rms volts on eight ranges and dc volts in five ranges. An averaging mode is available to automatically optimize display resolution and stability for each range in dc volts, resistance, and dc current functions. Extended resolution is also available in the ac volts function. Optional modules are available for dc current (five ranges), and resistance (eight ranges) in two-wire or four-wire arrangements.

1-5. Thermal True-RMS Conversion

1-6. The thermal true-rms feature allows the operator to measure the true-rms value of an ac signal at accuracies of up to .012% with a reading rate of one every six seconds. This response time compares favorably with that of existing thermal transfer standards which can take up to five minutes to complete a measurement.

1-7. Modular Construction

1-8. Considerable versatility is realized through unique modular construction. All active components are contained in modules which plug into a mainframe motherboard. This module-to-motherboard mating, combined with bus architecture and microprocessor control, yields ease of option selection.

1-9. Microprocessor Control

1-10. All modules function under direct control of a microprocessor based controller. Each module is addressed by the controller as virtual memory. Scaling factors and offset values can be applied separately, stored in memory, and automatically used as factors in all subsequent readings. Digital filtering utilizes averaged samples for each reading.

1-11. Software Calibration

1-12. The 8506A features microprocessor-controlled calibration of all ranges and functions. Any range can be calibrated using a reference input of any known value from 60% of range to full scale. Software calibration can be performed using front-panel or remote control, allowing recertification without opening the case or removing the multimeter from the system.

1-13. Recirculating Remainder A/D Conversion

1-14. The multimeter adapts Fluke's patented recirculating remainder (R^2) A/D conversion technique to microprocessor control. This combination provides fast, accurate, linear measurements and long-term stability.

1-15. Options and Accessories

1-16. Remote interfaces, a dc current converter, and an ohms converter are among the options and accessories available for use with the multimeter. Refer to Tables 1-1 and 1-2 for complete listings. Any one of the three Remote Interface modules (Option 05, 06, or 07) may be installed at one time.

1-17. SPECIFICATIONS

1-18. Mainframe specifications for ac volts, dc volts and dc ratio measurement capability are presented in Table 1-3. Optional function specifications are supplied with the respective option modules and included in Section 6.

Table 1-1. Options

OPTION NO.	NAME	NOTES
02A	Ohms Converter	1
03	Current Shunts	1
05	IEEE Standard 488-1975 Interface	2
06	Bit Serial Asynchronous Interface	2
07	Parallel Interface	2

1) Either Option 02A or Option 03 can be installed at one time.

2) Only one of Options 05, 06, and 07 can be installed at any time.

Table 1-2. Accessories

MODEL OR PART NO.	NAME
M04-205-600	5¼-inch Rack Adapter
M00-260-610	18-inch Rack Slides
M00-280-610	24-inch Rack Slides
80K-6	High Voltage Probe
80K-40	High Voltage Probe
83RF	High Frequency Probe
85RF	High Frequency Probe
Y8021	IEEE Std. Cable, 1 Meter Length
Y8022	IEEE Std. Cable, 2 Meter Length
Y8023	IEEE Std. Cable, 4 Meter Length
629170	TRMS Extender Card
MIS-7190K*	Static Controller
MIS-7013K*	Bus Interconnect and Monitor

*For use during service or repair.

Table 1-3. Specifications

GENERAL SPECIFICATIONS

Dimensions	10.8 cm High x 43.2 cm Wide x 42.5 cm Long (4.25 in High x 17 in Wide x 16.75 in Long) (See Figure 1-1)
Weight	
BASIC	10 kg (22 lbs)
FULLY LOADED	12 kg (26 lbs)
Operating Power	
VOLTAGE	100V ac, 120V ac, 220V ac, or 240V ac (±10%)
BASIC INSTRUMENT POWER	12 watts
FULLY LOADED POWER	24 watts
FREQUENCY	47 Hz to 63 Hz (400 Hz available on request)
Warm-Up	2 hours to rated accuracy
Shock and Vibration	Meets requirements of MIL-T-28800 for type III, class 5, style E equipment.
Temperature Range	
OPERATING	0°C to 50°C
NON-OPERATING	-40°C to 70°C
Humidity Range	
0°C TO 18°C	80% RH
18°C TO 40°C	75% RH
40°C TO 50°C	45% RH
Maximum Terminal Voltage	
LO TO GUARD	127V rms
GUARD TO CHASSIS	500V rms
HI SENSE TO HI SOURCE	127V rms
LO SENSE TO LO SOURCE	127V rms
HI SENSE TO LO SENSE	1000V rms or 1200V dc
HI SOURCE TO LO SOURCE	280V rms

Table 1-3. Specifications (cont)

AC VOLTAGE

Input Characteristics

RANGE	FULL SCALE 5½ DIGITS	RESOLUTION 6½ DIGITS*	RESOLUTION 5½ DIGITS	INPUT IMPEDANCE
100 mV	125.000 mV	—	1 µV	1 MΩ
300 mV	400.000 mV	—	1 µV	
1V	1.25000V	1 µV	10 µV	±1%
3V	4.00000V	1 µV	10 µV	
10V	12.5000V	10 µV	100 µV	Shunted by
30V	40.0000V	10 µV	100 µV	
100V	125.000V	100 µV	1 mV	<180 pF
500 V	600.000V	100 µV	1 mV	

*In AVG operating mode.

Accuracy

HIGH ACCURACY MODE ±(% of Reading + Number of Counts)[1]

24 HOUR: 23°C ±1°C[2]

RANGE	FREQUENCY IN HERTZ 10 TO 40*	40 TO 20k	20k TO 50k	50k TO 100k	100k TO 200k	200k TO 500k	500k TO 1M
100 mV	0.08 + 0	0.02 + 5	0.04 + 5	0.2 + 0	0.6 + 0	1.5 + 0	3.5 + 0
300 mV to 10V	0.08 + 0	0.012 + 0	0.04 + 0	0.2 + 0	0.5 + 0	1.5 + 0	3.5 + 0
30V	0.08 + 0	0.012 + 0	0.04 + 0	0.2 + 0	0.5 + 0	3.5 + 0	12 + 0
100V	0.08 + 0	0.012 + 0	0.04 + 0	0.2 + 0	1.0 + 0	3.5 + 0	—
500V[3]	0.08 + 0	0.012 + 0	0.04 + 0	0.2 + 0	—	—	—

90 DAY: 23°C ±5°C

RANGE	FREQUENCY IN HERTZ 10 TO 40*	40 TO 20k	20k TO 50k	50k TO 100k	100k TO 200k	200k TO 500k	500k TO 1M
100 mV	0.08 + 0	0.026 + 5	0.06 + 0	0.2 + 0	0.6 + 0	1.5 + 0	3.5 + 0
300 mV to 10V	0.08 + 0	0.016 + 0	0.06 + 0	0.2 + 0	0.5 + 0	1.5 + 0	3.5 + 0
30V	0.08 + 0	0.016 + 0	0.06 + 0	0.2 + 0	0.5 + 0	3.5 + 0	12 + 0
100V	0.08 + 0	0.016 + 0	0.06 + 0	0.2 + 0	1.0 + 0	3.5 + 0	—
500V[3]	0.08 + 0	0.016 + 0	0.06 + 0	0.2 + 0	—	—	—

*With slow filter

Table 1-3. Specifications (cont)

AC VOLTAGE (cont)

Input Characteristics (cont)

>90 DAY: 23°C ±5°C ADD TO THE 90 DAY SPECIFICATION PER MONTH THE FOLLOWING % OF READING							
ALL RANGES	FREQUENCY IN HERTZ						
	10 TO 40	40 TO 20k	20k TO 50k	50k TO 100k	100k TO 200k	200k TO 500k	500k TO 1M
	0.008	0.001	0.0025	0.012	0.021	0.06	0.11

NOTES:

[1] AC coupled, 5½ digits, input level >0.25 x full scale. For 6½ digits multiply Number of Counts by 10. For input levels between 0.1 x and 0.25 x full scale, add 5 counts for the 100 mV, 1V, 10V, and 100V ranges, add 15 counts for the 300 mV, 3V, 30V ranges, and add 25 counts for the 500V range.

[2] Relative to calibration standards, within 1 hour of dc zero.

[3] Add 0.02 x $(\text{Input voltage} / 600)^2$ % of Reading to the specification.

ENHANCED MODE: Add the following (% of Reading + Number of Counts) to the High Accuracy Mode Specifications.

RANGE	TIME SINCE FIRST READING	
	<5 MINUTES	<30 MINUTES
100 mV, 1V, 10V, 100V	0 + 0	0.003 + 4
300 mV, 3V, 30V	0 + 0	0.003 + 4
500V	0 + 0	0.003 + 6

*AC-coupled, 5½ digits, temperature change <1° C, input level >0.25 x full scale. For input levels between 0.1x and 0.25x full scale, multiply % of Reading adder by 10.

NORMAL MODE: Add the following % of Reading to the High Accuracy Mode Specification.

SEGMENT OF SCALE	24 HOUR, 90 DAY	>90 DAY ADD PER MONTH
0.25x to 1x full scale	0.4	0.044
0.1x to 0.25x full scale	0.6	0.055

AC+DC COUPLED MODES: ±(1.1 times the ac specification for the appropriate mode + the result (Adder) from the following table).

RANGE	ADDER
100 mV to 1V	±(150 μV x (dc volts / total rms volts))
3V and 10V	±(1 mV x (dc volts / total rms volts))
30V and 100V	±(10 mV x (dc volts / total rms volts))
500V	±(50 mV x (dc volts / total rms volts))

Table 1-3. Specifications (cont)

Operating Characteristics

STABILITY: ±(1% of Reading + Number of Counts)*

RANGE	24 HOUR	90 DAY
100 mV, 1V, 10V, 100V	0.0025 + 1	0.004 + 1
300 mV, 3V, 30V	0.0025 + 3	0.004 + 4
500V	0.0025 + 5	0.004 + 6

*High Accuracy Mode, ac coupled, 5½ digits, input level >0.25x full scale, 40 Hz to 20 kHz, temperature change <1°C. For 6½ digits, multiply Number of Counts by 10. For input levels between 0.1x and 0.25x full scale, add to the Number of Counts specification 2 counts for the 100 mV, 1V, 10V, and 100V ranges, 6 counts for the 300 mV, 3V, and 30V ranges, and 10 counts for the 500V range.

CREST FACTOR .... Up to 8:1 at full 90 day (or greater) accuracy for input signals with peaks less than two times full scale, and highest frequency components within the 3 dB bandwidth. Up to 4:1 for signals with peaks less than four times full scale, with an addition of 0.03 to the % of Reading.

3 dB BANDWIDTH .... 3 MHz for the 100 mV range and 10 MHz for the 300 mV, 1V, 3V and 10V ranges (typical).

MAXIMUM INPUT VOLTAGE .... ±600V rms or dc, 840V peak, or 1x 10^7 volts-hertz product.

TEMPERATURE COEFFICIENT ...... 0°C to 18°C and 28°C to 50°C
1/10 of 90 day Specification per °C

COMMON MODE REJECTION .... >120 dB, dc to 60 hertz, with 100Ω in series with either lead.

SETTLING TIME

High Accuracy Mode .... Sample time = 3.5 seconds
Hold time = 2.5 seconds
Measurement time = 6 seconds

If the state of the instrument is unknown, two complete measurement times will be required to guarantee a correct reading. Use of the external trigger mode will always allow a 6 second measurement time.

Enhanced Mode .... The first reading requires the same time as the High Accuracy Mode. Subsequent readings occur every 500 milliseconds. If the input changes 1% the analog settling time to 90 Day mid-band accuracy is 1.5 seconds.

Normal Mode .... Settling times for large changes are non-linear. Zero to Full Scale changes require 2.0 seconds to settle to 90 Day, mid-band specifications. Full scale to 1/10th full scale changes require 3.0 seconds to settle to 1/10th full scale, mid-band; 90 day specifications. Small changes (<1%) settle to mid-band specifications in <1.5 seconds.

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Table 1-3. Specifications (cont)

AC VOLTAGE (cont)

Operating Characteristics (cont)

AUTORANGE POINTS

RANGE	UPRANGE	DOWNRANGE
100 mV	125.000 mV	None
300 mV	400.000 mV	110 mV
1V	1.25000V	0.352V
3V	4.00000V	1.1V
10V	12.5000V	3.52V
30V	40.0000V	11V
100V	125.000V	35.2V
500V	None	110V

OPERATING RANGE

RANGE	UNDERRANGE DISPLAY LLLLL	MINIMUM SPECIFIED LEVEL	OVERRANGE DISPLAY HHHHH
100 mV	None	12.5 mV	125.000 mV
300 mV	20 mV	40 mV	400.000 mV
1V	62.5 mV	125 mV	1.25000V
3V	200 mV	400 mV	4.00000V
10V	625 mV	1.25V	12.5000V
30V	2V	4V	40.0000V
100V	6.25V	12.5V	125.000V
500V	30V	60V	600.000V

DC VOLTAGE

Input Characteristics

RANGE	FULL SCALE 6½ DIGITS	RESOLUTION		INPUT RESISTANCE
		7½ DIGITS*	6½ DIGITS	
100 mV	200.0000 mV	—	100 nV	>10,000MΩ
1V	2.000000V	—	1 μV	>10,000MΩ
10V	20.00000V	1 μV	10 μV	>10,000MΩ
100V	128.0000V	—	100 μV	10MΩ
1000V	1200.000V	—	1 mV	10MΩ

*7½-digit resolution: In AVG operating mode.

Table 1-3. Specifications (cont)

Accuracy

DC VOLTS: ±(% of Reading + Number of Counts)				
RANGE	24-HOUR 23°C ±1°C[1]		90-DAY 23°C ±5°C	
	OPERATING MODE		OPERATING MODE	
	NORM	AVG	NORM	AVG[3]
100 mV	0.0018 + 15	0.0010 + 8	0.0025 + 40	0.0020 + 8
1V	0.0008 + 7	0.0005 + 4	0.0015 + 8	0.0012 + 6
10V	0.0006 or 6*	0.0005 or 50[2]*	0.0010 + 8	0.0008 + 60[2]
100V	0.0010 + 6	0.0005 + 5	0.0018 + 8	0.0015 + 6
1000V	0.0008 + 6	0.0005 + 5	0.0018 + 8	0.0015 + 6
*Whichever is greater				

>90-Day: 23°C ±5°C
Add to the 90-day specification per month the following % of Reading and Number of Counts.

RANGE	OPERATING MODE	
	NORM	AVG[3]
100 mV	0.00017 + 5.6	0.0001 + 0.1
1V	0.0001 + 0.1	0.0001 + 0.1
10V	0.0001 + 0.1	0.00008 + 1[2]
100V	0.00013 + 0.1	0.0001 + 0.1
1000V	0.00013 + 0.1	0.0001 + 0.1

NOTES:

[1]Relative to calibration standards, 4-hour warm-up, within 1 hour of dc zero. After software calibration, add the following to the 24 hour accuracy specification:

TIME SINCE INTERNAL (HARDWARE) CALIBRATION	NUMBER OF COUNTS*
<30 Days	0
<90 Days	1
<1 Year	2
>1 Year	3
*With 6½-digit display. For 7½-digits, multiply Number of Counts by 10.	

[2]7½-digit mode of operation.

[3]After 4-hour warm-up, within 1 hour of dc zero.

Table 1-3. Specifications (cont)

Operating Characteristics

TEMPERATURE COEFFICIENT: ±(% of Reading + Number of Counts)/°C

RANGE	0°C TO 18°C AND 28°C TO 50°C
100 mV	0.0003 + 5
1V	0.0003 + 1
10V	0.0002 + 0.5*
100V	0.0003 + 1
1000V	0.0003 + 0.5

*Multiply Number of Counts by 10 for AVG operating mode (7½-digit).

INPUT BIAS CURRENT

AT TIME OF ADJUSTMENT	1-YEAR 23°C ±1°C	TEMPERATURE COEFFICIENT
<±5 pA	<±30 pA	<±1 pA/°C

ZERO STABILITY .... Less than 5 μV for 90 days after a 4-hour warm-up. Front panel ZERO push button stores a zero correction factor for each range.

MAXIMUM INPUT VOLTAGE .... ±1200V dc or 1000V rms ac to 60 Hz, or 1400V peak above 60 Hz may be applied continuously to any dc range without permanent damage. Maximum common mode rate of voltage rise is 1000V / μsec.

ANALOG SETTLING TIME

FILTER MODE	FILTER COMMAND	TO 0.01% OF STEP CHANGE	TO 0.001% OF STEP CHANGE
Bypassed	F1	2 ms	20 ms
Fast	F0 or F3	40 ms	50 ms
Slow	F or F2	400 ms	500 ms

DIGITIZING TIME

Line Synchronous .... For 2^0 to 2^{17} samples per reading digitizing time is from 4 ms to 9 minutes 6 seconds using a 60 Hz ac line with times increasing 20% using a 50 Hz ac line. Selectable in 18 binary steps.

Line Asynchronous .... 2 ms. (In 3 byte binary mode with dc zero, offset, limits and calibration factors turned off.)

NOISE REJECTION

Normal Mode Rejection

LINE FREQUENCY	FILTER MODE	4 SAMPLES/ READING	32 SAMPLES/ READING	128 SAMPLES/ READING
50 hertz	Fast	60 dB	70 dB	75 dB
50 hertz	Slow	85 dB	90 dB	95 dB
60 hertz	Fast	60 dB	70 dB	75 dB
60 hertz	Slow	90 dB	95 dB	100 dB

Common Mode Rejection .... 160 dB at 60 hertz with 1 kΩ in series with either lead, and 4 samples or more per reading. Greater than or equal to 100 dB with less than 4 samples per reading.

Table 1-3. Specifications (cont)

DC RATIO

Accuracy

EXTERNAL REFERENCE VOLTAGE*	ACCURACY[1]
±20V to ±40V	±(A + B + 0.001%)
$\pm V_{min}$ to ±20V	±(A + B + (0.02% / $\lvert V_{xref} \rvert$))

*Maximum External Reference Voltage = ±40V between External Reference HI and LO terminals, providing neither terminal is greater than ±20V relative to the Sense LO or Ohms Guard[2] terminals.

Operating Characteristics

INPUT IMPEDANCE .... External Reference HI or LO >10,000 MΩ relative to Ohms Guard[2] or Sense LO.

BIAS CURRENT .... External Reference HI or LO relative to Ohms Guard[2] or Sense LO <5 nA.

SOURCE IMPEDANCE .... Resistive Unbalance (External Reference HI to LO) <4 kΩ. Total Resistance to Sense LO from either External Reference HI or LO <20 kΩ.

MAXIMUM OVERLOAD VOLTAGE ... ±180V dc or peak ac (relative to Ohms Guard[2] or Sense LO). ±300V dc or peak ac (External Reference HI to LO).

NOISE REJECTION

INPUT TERMINALS	NORMAL MODE	COMMON MODE
Sense	Same as dc volts	Same as dc volts
External Reference	line frequency and 2x line frequency >100 dB	line frequency and 2x line frequency >75 dB

RESPONSE TIME

Analog Settling Time

FILTER MODE	FILTER COMMAND	TO 0.01% OF STEP CHANGE	TO 0.001% OF STEP CHANGE
Bypassed	F1	2 ms	20 ms
Fast	F0 or F3	40 ms	50 ms
Slow	F or F2	400 ms	500 ms

NOTES: (DC Ratio)

[1]A = 10V dc range accuracy for the appropriate period of time.

B = Input signal function and range accuracy for the appropriate period of time.

V_{min} = Minimum allowable External Reference Voltage = ±0.0001V, or $V_{input}/10^{8}$ (whichever is greater).
$\lvert V_{xref} \rvert$ = Absolute value of the External Reference Voltage

[2]Ohms Guard is available through the rear input.

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Table 1-3. Specifications (cont)

DC RATIO (cont)	
Operating Characteristics (cont)	
Digitizing Time	For 2^0 to 2^{17} samples per reading digitizing time is from 196 ms to 9 minutes 6 seconds using a 60 Hz ac line with times increasing 20% using a 50 Hz ac line. Selectable in 18 binary steps.
MAXIMUM RATIO DISPLAY	+1.00000 E±9
EXTERNAL TRIGGER INPUT	
Polarity	May be wired internally for either rising or falling edge. Factory wired for falling edge.
High Level	+4.3V (minimum)
Low Level	+0.7V (maximum)
Pulse Width	10 μs (minimum)
Connector	BNC with the outer shell at interface common
Maximum Input	±30V
Maximum Shell to Ground Voltage	±30V
SCAN ADVANCE OUTPUT	
Polarity	Positive
High Level	>+4V (TTL High)
Low Level	<+0.7V (TTL Low)
Pulse Width	3 μs (minimum)
Connector	BNC with the outer shell at interface common
Maximum Shell to Ground Voltage	±30V

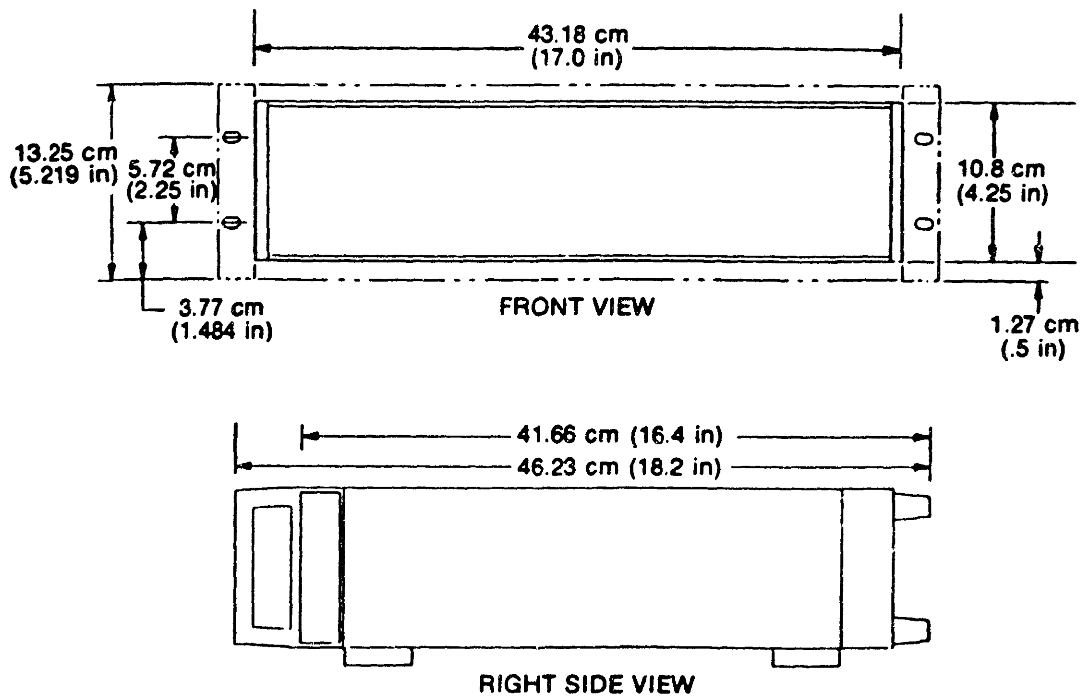

Figure 1-1. Dimension Drawing

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Appendix E. COMPONENT SPECIFICATIONS

This appendix contains the specification sheets for selected components used in the circuit portion of the two systems. In particular, they show the noise specifications for the components. The noise data is necessary in developing the noise model and in calculating the noise in the circuits.

Linear Databook 2 National Semiconductor Corp. 1988 Edition

National Semiconductor

Voltage References

LM199/LM299/LM399 Precision Reference

General Description

The LM199/LM299/LM399 are precision, temperature-stabilized monolithic zeners offering temperature coefficients a factor of ten better than high quality reference zeners. Constructed on a single monolithic chip is a temperature stabilizer circuit and an active reference zener. The active circuitry reduces the dynamic impedance of the zener to about 0.5Ω and allows the zener to operate over 0.5 mA to 10 mA current range with essentially no change in voltage or temperature coefficient. Further, a new subsurface zener structure gives low noise and excellent long term stability compared to ordinary monolithic zeners. The package is supplied with a thermal shield to minimize heater power and improve temperature regulation.

The LM199 series references are exceptionally easy to use and free of the problems that are often experienced with ordinary zeners. There is virtually no hysteresis in reference voltage with temperature cycling. Also, the LM199 is free of voltage shifts due to stress on the leads. Finally, since the unit is temperature stabilized, warm up time is fast.

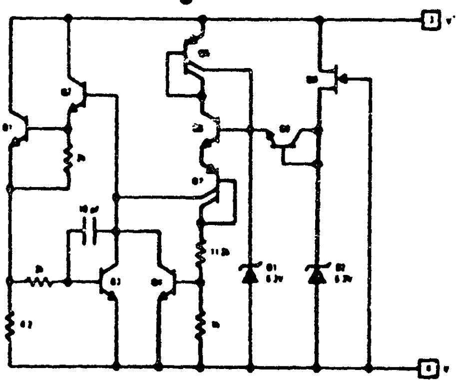
The LM199 can be used in almost any application in place of ordinary zeners with improved performance. Some ideal applications are analog to digital converters, calibration standards, precision voltage or current sources or precision power supplies. Further in many cases the LM199 can replace references in existing equipment with a minimum of wiring changes.

The LM199 series devices are packaged in a standard hermetic TO-46 package inside a thermal shield. The LM199 is rated for operation from −55°C to +125°C while the LM299 is rated for operation from −25°C to +85°C and the LM399 is rated from 0°C to +70°C.

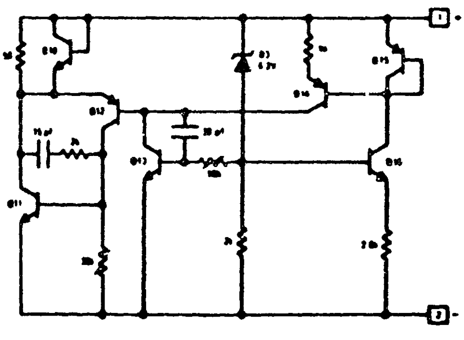
Features

- Guaranteed 0.0001%/°C temperature coefficient
- Low dynamic impedance – 0.5Ω
- Initial tolerance on breakdown voltage – 2%
- Sharp breakdown at 400μA
- Wide operating current – 500μA to 10 mA
- Wide supply range for temperature stabilizer
- Guaranteed low noise
- Low power for stabilization – 300 mW at 25°C
- Long term stability – 20 ppm

Schematic Diagrams



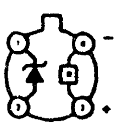
Temperature Stabilizer



Reference

Connection Diagram

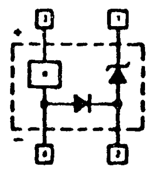
Metal Can Package



TOP VIEW

Order Number LM199H, LM299H or LM399H
See Package H04A

Functional Block Diagram



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Absolute Maximum Ratings

Temperature Stabilizer Voltage	40V
Reverse Breakdown Current	20 mA
Forward Current	1 mA
Reference to Substrate Voltage $V_{(RS)}$ (Note 1)	40V
	−0.1V
Operating Temperature Range	
LM199	−55°C to +125°C
LM299	−25°C to +85°C
LM399	0°C to +70°C
Storage Temperature Range	−55°C to +150°C
Lead Temperature (Soldering, 10 seconds)	300°C

Electrical Characteristics (Note 2)

PARAMETER	CONDITIONS		LM199/LM299			LM399			UNITS
			MIN	TYP	MAX	MIN	TYP	MAX	
Reverse Breakdown Voltage	$0.5\ mA \leq I_R \leq 10\ mA$		6.8	6.95	7.1	6.6	6.95	7.3	V
Reverse Breakdown Voltage Change With Current	$0.5\ mA \leq I \leq 10\ mA$			6	9		6	12	mV
Reverse Dynamic Impedance	$I_R = 1\ mA$			0.5	1		0.5	1.5	Ω
Reverse Breakdown Temperature Coefficient	$-55°C \leq T_A \leq 85°C$	LM199		0.00003	0.0001				%/°C
	$85°C \leq T_A \leq 125°C$	LM199		0.0005	0.0015				%/°C
	$-25°C \leq T_A \leq 85°C$	LM299		0.00003	0.0001				%/°C
	$0°C \leq T_A \leq 70°C$	LM399					0.00003	0.0002	%/°C
RMS Noise	$10\ Hz \leq f \leq 10\ kHz$			7	20		7	50	µV
Long Term Stability	Stabilized, $22°C \leq T_A \leq 28°C$, 1000 Hours, $I_R = 1\ mA \pm 0.1\%$			20			20		ppm
Temperature Stabilizer Supply Current	$T_A = 25°C$, Still Air, $V_S = 30V$			8.5	14		8.5	15	mA
	$T_A = -55°C$			22	28				mA
Temperature Stabilizer Supply Voltage	(Note 3)		9		40	9		40	V
Warm-Up Time to 0.05%	$V_S = 30V$, $T_A = 25°C$			3			3		Seconds
Initial Turn-on Current	$9 \leq V_S \leq 40$, $T_A = 25°C$, (Note 3)			140	200		140	200	mA

Note 1: The substrate is electrically connected to the negative terminal of the temperature stabilizer. The voltage that can be applied to either terminal of the reference is 40V more positive or 0.1V more negative than the substrate.

Note 2: These specifications apply for 30V applied to the temperature stabilizer and $-55°C \leq T_A \leq +125°C$ for the LM199, $-25°C \leq T_A \leq +85°C$ for the LM299 and $0°C \leq T_A \leq +70°C$ for the LM399.

Note 3: This initial current can be reduced by adding an appropriate resistor and capacitor to the heater circuit. See the performance characteristic graphs to determine values.

Analog IC Data Book PMI 1988

PMI

OP-27

LOW-NOISE PRECISION OPERATIONAL AMPLIFIER

Precision Monolithics Inc.

FEATURES

- **Low Noise** .... 80nV$_{p-p}$ (0.1Hz to 10Hz) .... 3nV/$\sqrt{Hz}$
- **Low Drift** .... 0.2μV/°C
- **High Speed** .... 2.8V/μs Slew Rate .... 8MHz Gain Bandwidth
- **Low V_{OS}** 10μV
- **Excellent CMRR** .... 126dB at V_{CM} of ±11V
- **High Open-Loop Gain** .... 1.8 Million
- **Fits 725, OP-07, OP-05, AD510, AD517, 5534A sockets**

ORDERING INFORMATION†

T_A = 25°C V_{OS} MAX (µV)	PACKAGE HERMETIC TO-99 8-PIN	HERMETIC DIP 8-PIN	PLASTIC DIP 8-PIN	LCC	OPERATING TEMPERATURE RANGE
25	OP27AJ*	OP27AZ*	—	—	MIL
25	OP27EJ	OP27EZ	OP27EP	—	IND/COM
60	OP27BJ*	OP27BZ*	—	OP27BRC/883	MIL
60	OP27FJ	OP27FZ	OP27FP	—	IND/COM
100	OP27CJ	OP27CZ	—	—	MIL
100	OP27GJ	OP27GZ	OP27GP	—	IND/COM

*For devices processed in total compliance to MIL-STD-883, add /883 after part number. Consult factory for 883 data sheet.

†Burn-in is available on commercial and industrial temperature range parts in cerdip, plastic dip, and TO-can packages. For ordering information, see 1988 Data Book, Section 2.

GENERAL DESCRIPTION

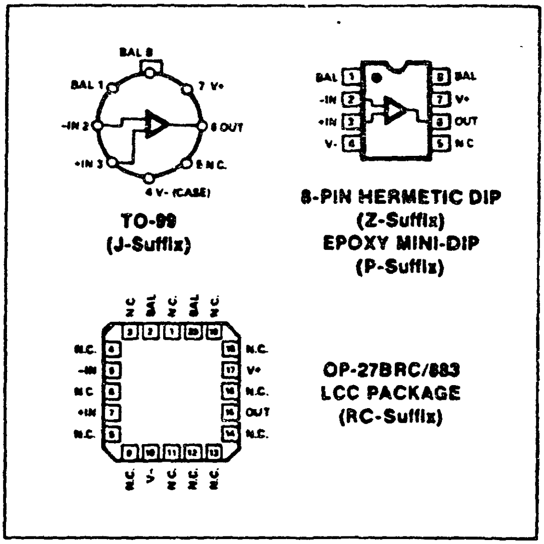
The OP-27 precision operational amplifier combines the low offset and drift of the OP-07 with both high-speed and low-noise. Offsets down to 25μV and drift of 0.6μV/°C maximum make the OP-27 ideal for precision instrumentation applications. Exceptionally low noise, e_n = 3.5nV/$\sqrt{Hz}$, at 10Hz, a low 1/f noise corner frequency of 2.7Hz, and high gain (1.8 million), allow accurate high-gain amplification of low-level signals. A gain-bandwidth product of 8MHz and a 2.8V/μsec slew rate provides excellent dynamic accuracy in high-speed data-acquisition systems.

A low input bias current of ±10nA is achieved by use of a bias-current-cancellation circuit. Over the military temperature range, this circuit typically holds I_B and I_{OS} to ±20nA and 15nA respectively.

The output stage has good load driving capability. A guaranteed swing of ±10V into 600Ω and low output distortion make the OP-27 an excellent choice for professional audio applications.

PSRR and CMRR exceed 120dB. These characteristics, coupled with long-term drift of 0.2μV/month, allow the circuit designer to achieve performance levels previously attained only by discrete designs.

PIN CONNECTIONS

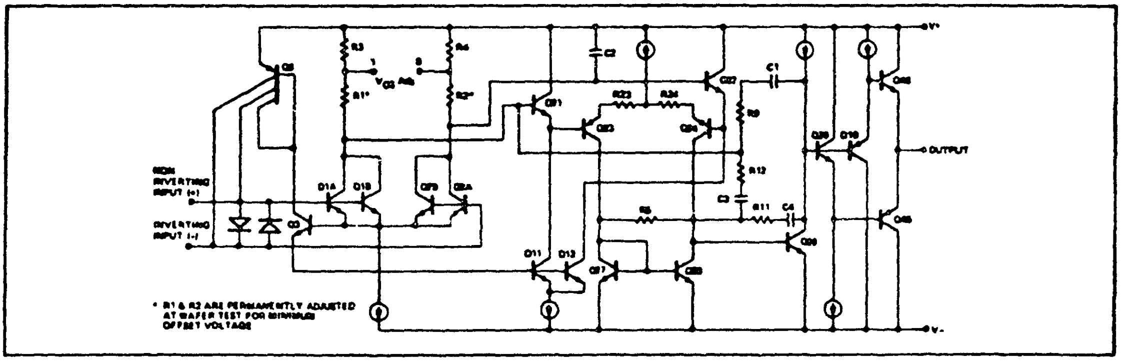


TO-99 (J-Suffix)

8-PIN HERMETIC DIP (Z-Suffix) EPOXY MINI-DIP (P-Suffix)

OP-27BRC/883 LCC PACKAGE (RC-Suffix)

SIMPLIFIED SCHEMATIC



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Low cost, high-volume production of OP-27 is achieved by using an on-chip zener-zap trimming network. This reliable and stable offset trimming scheme has proved its effectiveness over many years of production history.

The OP-27 provides excellent performance in low-noise high-accuracy amplification of low-level signals. Applications include stable integrators, precision summing amplifiers, precision voltage-threshold detectors, comparators, and professional audio circuits such as tape-head and microphone preamplifiers.

The OP-27 is a direct replacement for 725, OP-06, OP-07 and OP-05 amplifiers; 741 types may be directly replaced by removing the 741's nulling potentiometer.

ABSOLUTE MAXIMUM RATINGS (Note 4)

Supply Voltage .... ±22V
Internal Power Dissipation (Note 1) 500mW
Input Voltage (Note 3) .... ±22V
Output Short-Circuit Duration .... Indefinite
Differential Input Voltage (Note 2) .... ±0.7V
Differential Input Current (Note 2) .... ±25mA
Storage Temperature Range .... −65°C to +150°C
Operating Temperature Range
OP-27A, OP-27B, OP-27C (J, Z, RC) .... −55°C to +125°C
OP-27E, OP-27F, OP-27G (J, Z) .... −25°C to +85°C
OP-27E, OP-27F, OP-27G (P) .... 0°C to +70°C
Lead Temperature Range (Soldering, 60 sec) ...... 300°C
DICE Junction Temperature .... −65°C to +150°C

NOTES:

1. See table for maximum ambient temperature rating and derating factor.

PACKAGE TYPE	MAXIMUM AMBIENT TEMPERATURE FOR RATING	DERATE ABOVE MAXIMUM AMBIENT TEMPERATURE
TO-99 (J)	80°C	7.1mW/°C
8-Pin Hermetic DIP (Z)	75°C	6.7mW/°C
8-Pin Plastic DIP (P)	62°C	5.6mW/°C
LCC	80°C	7.8mW/°C

2. The OP-27's inputs are protected by back-to-back diodes. Current limiting resistors are not used in order to achieve low noise. If differential input voltage exceeds ±0.7V, the input current should be limited to 25mA.
3. For supply voltages less than ±22V, the absolute maximum input voltage is equal to the supply voltage.
4. Absolute maximum ratings apply to both DICE and packaged parts, unless otherwise noted.

ELECTRICAL CHARACTERISTICS at $V_S = \pm15V$, $T_A = 25°C$, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	OP-27A/E MIN	OP-27A/E TYP	OP-27A/E MAX	OP-27B/F MIN	OP-27B/F TYP	OP-27B/F MAX	OP-27C/G MIN	OP-27C/G TYP	OP-27C/G MAX	UNITS
Input Offset Voltage	V_{OS}	(Note 1)	—	10	25	—	20	60	—	30	100	µV
Long-Term V_{OS} Stability	V_{OS}/Time	(Notes 2, 3)	—	0.2	1.0	—	0.3	1.5	—	0.4	2.0	µV/Mo
Input Offset Current	I_{OS}		—	7	35	—	9	50	—	12	75	nA
Input Bias Current	I_B		—	±10	±40	—	±12	±55	—	±15	±80	nA
Input Noise Voltage	$e_{np\text{-}p}$	0.1Hz to 10Hz (Notes 3, 5)	—	0.08	0.18	—	0.08	0.18	—	0.09	0.25	µVp-p
Input Noise Voltage Density	e_n	f_O = 10Hz (Note 3)	—	3.5	5.5	—	3.5	5.5	—	3.8	8.0	nV/$\sqrt{Hz}$
		f_O = 30Hz (Note 3)	—	3.1	4.5	—	3.1	4.5	—	3.3	5.6	
		f_O = 1000Hz (Note 3)	—	3.0	3.8	—	3.0	3.8	—	3.2	4.5	
Input Noise Current Density	i_n	f_O = 10Hz (Notes 3, 6)	—	1.7	4.0	—	1.7	4.0	—	1.7	—	pA/$\sqrt{Hz}$
		f_O = 30Hz (Notes 3, 6)	—	1.0	2.3	—	1.0	2.3	—	1.0	—	
		f_O = 1000Hz (Notes 3, 6)	—	0.4	0.6	—	0.4	0.6	—	0.4	0.6	
Input Resistance — Differential-Mode	R_{IN}	(Note 7)	1.3	6	—	0.94	5	—	0.7	4	—	MΩ
Input Resistance — Common-Mode	R_{INCM}		—	3	—	—	2.5	—	—	2	—	GΩ
Input Voltage Range	IVR		±11.0	±12.3	—	±11.0	±12.3	—	±11.0	±12.3	—	V
Common-Mode Rejection Ratio	CMRR	V_{CM} = ±11V	114	126	—	106	123	—	100	120	—	dB
Power Supply Rejection Ratio	PSRR	V_S = ±4V to ±18V	—	1	10	—	1	10	—	2	20	µV/V
Large-Signal Voltage Gain	A_{VO}	$R_L \geq 2k\Omega$, V_O = ±10V	1000	1800	—	1000	1800	—	700	1500	—	V/mV
		$R_L \geq 600\Omega$, V_O = ±10V	800	1500	—	800	1500	—	600	1500	—	
Output Voltage Swing	V_O	$R_L \geq 2k\Omega$	±12.0	±13.8	—	±12.0	±13.8	—	±11.5	±13.5	—	V
		$R_L \geq 600\Omega$	±10.0	±11.5	—	±10.0	±11.5	—	±10.0	±11.5	—	
Slew Rate	SR	$R_L \geq 2k\Omega$ (Note 4)	1.7	2.8	—	1.7	2.8	—	1.7	2.8	—	V/µs

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LT1028 Pamplet Linear Technology 1987

$e_n i_n = .85 \times 10^{-21} \frac{W}{Hz}$

LINEAR TECHNOLOGY

LT1028

Ultra-Low Noise Precision High Speed Op Amp

FEATURES

- Voltage Noise 1.1nV/√Hz Max. at 1kHz
  - 0.85nV/√Hz Typ. at 1kHz
  - 1.0nV/√Hz Typ. at 10Hz
  - 35nVp-p Typ., 0.1Hz to 10Hz
- Voltage and Current Noise 100% Tested
- Gain-Bandwidth Product 50MHz Min.
- Slew Rate 11V/μs Min.
- Offset Voltage 40μV Max.
- Voltage Gain 7 Million Min.
- Drift with Temperature 0.8μV/°C Max.

APPLICATIONS

- Low Noise Frequency Synthesizers
- High Quality Audio
- Infrared Detectors
- Accelerometer and Gyro Amplifiers
- 350Ω Bridge Signal Conditioning
- Magnetic Search Coil Amplifiers
- Hydrophone Amplifiers

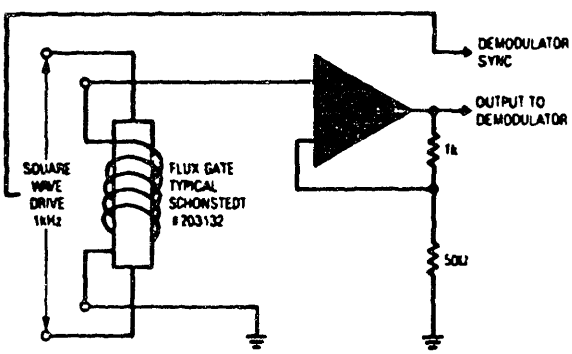
DESCRIPTION

The LT1028 achieves a new standard of excellence in noise performance with 0.85nV/√Hz 1kHz noise, 1.0nV/√Hz 10Hz noise. This ultra low noise is combined with excellent high speed specifications (gain-bandwidth product is 75MHz), distortion free output, and true precision parameters (0.1μV/°C drift, 10μV offset voltage, 30 million voltage gain). Although the LT1028 input stage operates at nearly 1mA of collector currents to achieve low voltage noise, input bias current is only 25nA.

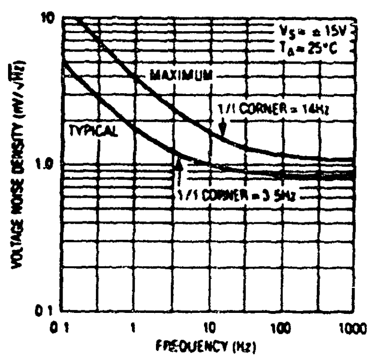
The LT1028's voltage noise is less than the noise of a 50Ω resistor. Therefore, even in very low source impedance transducer or audio amplifier applications, the LT1028's contribution to total system noise will be negligible.

$e_n i_n =$

Flux Gate Amplifier



Voltage Noise vs Frequency



ABSOLUTE MAXIMUM RATINGS

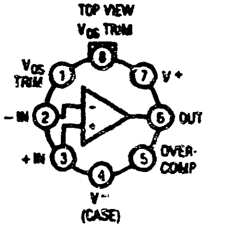
Supply Voltage
- −55°C to 105°C .... ±22V
- 105°C to 125°C .... ±16V

Differential Input Current (Note 8) .... ±25mA
Input Voltage .... Equal to Supply Voltage
Output Short Circuit Duration .... Indefinite
Operating Temperature Range
- LT1028AM, M .... −55°C to 125°C
- LT1028AC, C .... 0°C to 70°C

Storage Temperature Range
- All Devices .... −65°C to 150°C

Lead Temperature (Soldering, 10 sec.) .... 300°C

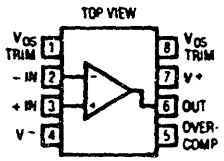
PACKAGE/ORDER INFORMATION



H8 PACKAGE TO-5 METAL CAN

ORDER PART NUMBER

LT1028AMH
LT1028MH
LT1028ACH
LT1028CH



J8 PACKAGE HERMETIC DIP
N8 PACKAGE PLASTIC DIP

LT1028AMJ8
LT1028MJ8
LT1028ACJ8
LT1028CJ8
LT1028ACN8
LT1028CN8

ELECTRICAL CHARACTERISTICS $V_S = \pm 15V$, $T_A = 25°C$, unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS	LT1028AM/AC MIN	LT1028AM/AC TYP	LT1028AM/AC MAX	LT1028M/C MIN	LT1028M/C TYP	LT1028M/C MAX	UNITS
V_{OS}	Input Offset Voltage	(Note 1)		10	40		20	80	μV
$\frac{\Delta V_{OS}}{\Delta Time}$	Long Term Input Offset Voltage Stability	(Note 2)		0.3			0.3		μV/Mo
I_{OS}	Input Offset Current	$V_{CM} = 0V$		12	50		18	100	nA
I_B	Input Bias Current	$V_{CM} = 0V$		±25	±90		±30	±180	nA
e_n	Input Noise Voltage	0.1Hz to 10Hz (Note 3)		35	75		35	90	nVp-p
	Input Noise Voltage Density	$f_o = 10Hz$ (Note 4) $f_o = 1000Hz$, 100% tested		1.0 0.85	1.7 1.1		1.0 0.9	1.9 1.2	nV/√Hz nV/√Hz
i_n	Input Noise Current Density	$f_o = 10Hz$ (Notes 3 and 5) $f_o = 1000Hz$, 100% tested		4.7 1.0	10.0 1.6		4.7 1.0	12.0 1.8	pA/√Hz pA/√Hz
	Input Resistance Common-Mode Differential Mode			 300 20			 300 20		 MΩ kΩ
	Input Capacitance			5			5		pF
	Input Voltage Range		±11.0	±12.2		±11.0	±12.2		V
CMRR	Common-Mode Rejection Ratio	$V_{CM} = \pm 11V$	114	126		110	126		dB
PSRR	Power Supply Rejection Ratio	$V_S = \pm 4V$ to $\pm 18V$	117	133		110	132		dB
A_{VOL}	Large Signal Voltage Gain	$R_L \ge 2k\Omega$, $V_o = \pm 12V$ $R_L \ge 1k\Omega$, $V_o = \pm 10V$ $R_L \ge 600\Omega$, $V_o = \pm 10V$	7.0 5.0 3.0	30.0 20.0 15.0		5.0 3.5 2.0	30.0 20.0 15.0		V/μV V/μV V/μV
V_{OUT}	Maximum Output Voltage Swing	$R_L \ge 2k\Omega$ $R_L \ge 600\Omega$	±12.3 ±11.0	±13.0 ±12.2		±12.0 ±10.5	±13.0 ±12.2		V V
SR	Slew Rate	$A_{VCL} = -1$	11	15		11	15		V/μs
GBW	Gain-Bandwidth Product	$f_o = 20kHz$ (Note 6)	50	75		50	75		MHz
Z_o	Open Loop Output Impedance	$V_o = 0$, $I_o = 0$		80			80		Ω
I_S	Supply Current			7.4	9.5		7.6	10.5	mA

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ELECTRICAL CHARACTERISTICS $V_S = \pm 15V$, $-55°C \le T_A \le 125°C$, unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS		LT1028AM MIN	LT1028AM TYP	LT1028AM MAX	LT1028M MIN	LT1028M TYP	LT1028M MAX	UNITS
V_{OS}	Input Offset Voltage	(Note 1)	●		30	120		45	180	μV
$\frac{\Delta V_{OS}}{\Delta Temp}$	Average Input Offset Drift	(Note 7)	●		0.2	0.8		0.25	1.0	μV/°C
I_{OS}	Input Offset Current	$V_{CM} = 0V$	●		25	90		30	180	nA
I_B	Input Bias Current	$V_{CM} = 0V$	●		±40	±150		±50	±300	nA
	Input Voltage Range		●	±10.3	±11.7		±10.3	±11.7		V
CMRR	Common-Mode Rejection Ratio	$V_{CM} = \pm 10.3V$	●	106	122		100	120		dB
PSRR	Power Supply Rejection Ratio	$V_S = \pm 4.5V$ to $\pm 18V$	●	110	130		104	130		dB
A_{VOL}	Large Signal Voltage Gain	$R_L \ge 2k\Omega$, $V_O = \pm 10V$ $R_L \ge 1k\Omega$, $V_O = \pm 10V$	●	3.0 2.0	14.0 10.0		2.0 1.5	14.0 10.0		V/μV V/μV
V_{OUT}	Maximum Output Voltage Swing	$R_L \ge 2k\Omega$	●	±10.3	±11.6		±10.3	±11.6		V
I_S	Supply Current		●		8.7	11.5		9.0	13.0	mA

ELECTRICAL CHARACTERISTICS $V_S = \pm 15V$, $0°C \le T_A \le 70°C$, unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS		LT1028AC MIN	LT1028AC TYP	LT1028AC MAX	LT1028C MIN	LT1028C TYP	LT1028C MAX	UNITS
V_{OS}	Input Offset Voltage	(Note 1)	●		15	80		30	125	μV
$\frac{\Delta V_{OS}}{\Delta Temp}$	Average Input Offset Drift	(Note 7)	●		0.1	0.8		0.2	1.0	μV/°C
I_{OS}	Input Offset Current	$V_{CM} = 0V$	●		15	65		22	130	nA
I_B	Input Bias Current	$V_{CM} = 0V$	●		±30	±120		±40	±240	nA
	Input Voltage Range		●	±10.5	±12.0		±10.5	±12.0		V
CMRR	Common-Mode Rejection Ratio	$V_{CM} = \pm 10.5V$	●	110	124		106	124		dB
PSRR	Power Supply Rejection Ratio	$V_S = \pm 4.5V$ to $\pm 18V$	●	114	132		107	132		dB
A_{VOL}	Large Signal Voltage Gain	$R_L \ge 2k\Omega$, $V_O = \pm 10V$ $R_L \ge 1k\Omega$, $V_O = \pm 10V$	●	5.0 4.0	25.0 18.0		3.0 2.5	25.0 18.0		V/μV V/μV
V_{OUT}	Maximum Output Voltage Swing	$R_L \ge 2k\Omega$ $R_L \ge 600\Omega$ (Note 9)	●	±11.5 ±9.5	±12.7 ±11.0		±11.5 ±9.0	±12.7 ±10.5		V V
I_S	Supply Current		●		8.0	10.5		8.2	11.5	mA

The ● denotes the specifications which apply over the full operating temperature range.

Note 1: Input Offset Voltage measurements are performed by automatic test equipment approximately 0.5 sec. after application of power. In addition, at $T_A = 25°C$, offset voltage is measured with the chip heated to approximately 55°C to account for the chip temperature rise when the device is fully warmed up.

Note 2: Long Term Input Offset Voltage Stability refers to the average trend line of Offset Voltage vs. Time over extended periods after the first 30 days of operation. Excluding the initial hour of operation, changes in V_{OS} during the first 30 days are typically 2.5μV.

Note 3: This parameter is tested on a sample basis only.

Note 4: 10Hz noise voltage density is sample tested on every lot. Devices 100% tested at 10Hz are available on request.

Note 5: Current noise is defined and measured with balanced source resistors. The resultant voltage noise (after subtracting the resistor noise on an RMS basis) is divided by the sum of the two source resistors to obtain current noise. Maximum 10Hz current noise can be inferred from 100% testing at 1kHz.

Note 6: Gain-bandwidth product is not tested. It is guaranteed by design and by inference from the slew rate measurement.

Note 7: This parameter is not 100% tested.

Note 8: The inputs are protected by back-to-back diodes. Current limiting resistors are not used in order to achieve low noise. If differential input voltage exceeds ±1.8V, the input current should be limited to 25mA.

Note 9: This parameter guaranteed by design, fully warmed up at $T_A = 70°C$. It includes chip temperature increase due to supply and load currents.

Analog Product Data Book, Harris, 1986

HARRIS

HA-2600/02/05

Wideband, High Impedance Operational Amplifiers

FEATURES

- WIDE BANDWIDTH 12MHz
- HIGH INPUT IMPEDANCE 500MΩ
- LOW INPUT BIAS CURRENT 1nA
- LOW INPUT OFFSET CURRENT 1nA
- LOW INPUT OFFSET VOLTAGE 0.5mV
- HIGH GAIN 150K V/V
- HIGH SLEW RATE 7V/μs
- OUTPUT SHORT CIRCUIT PROTECTION

APPLICATIONS

- VIDEO AMPLIFIER
- PULSE AMPLIFIER
- AUDIO AMPLIFIERS AND FILTERS
- HIGH-Q ACTIVE FILTERS
- HIGH-SPEED COMPARATORS
- LOW DISTORTION OSCILLATORS

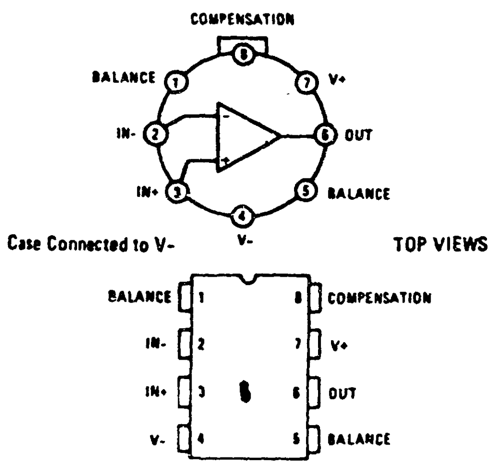
DESCRIPTION

HA-2600/2602/2605 are internally compensated bipolar operational amplifiers that feature very high input impedance (500 MΩ, HA-2600) coupled with wideband AC performance. The high resistance of the input stage is complemented by low offset voltage (0.5mV, HA-2600) and low bias and offset current (1nA, HA-2600) to facilitate accurate signal processing. Input offset can be reduced further by means of an external nulling potentiometer. 12MHz unity gain-bandwidth product, 7V/μs slew rate and 150,000V/V open-loop gain enables HA-2600/2602/2605 to perform high-gain amplification of fast, wideband signals. These dynamic characterisitics, coupled with fast settling times, make these amplifiers ideally suited to pulse amplification designs as well as high frequency (e.g. video) applications. The frequency response of the amplifier can be tailored to exact design requirements by means of an external bandwidth control capacitor.

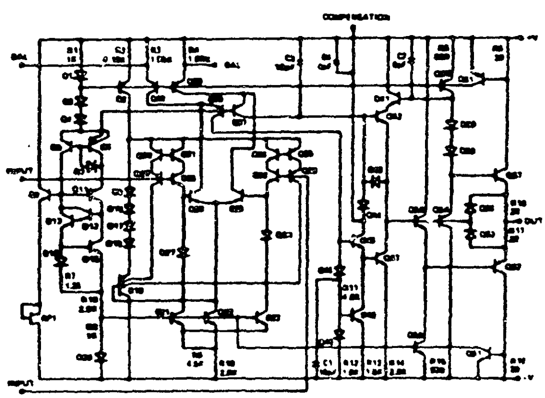
In addition to its application in pulse and video amplifier designs, HA-2600/2602/2605 is particularly suited to other high performance designs such as high-gain low distortion audio amplifiers, high-Q and wideband active filters and high-speed comparators.

The HA-2600 and HA-2602 have guaranteed operation from -55°C to +125°C and are available in metal can and ceramic mini DIP packages. Both are offered as a military grade part. The HA-2605 has guaranteed operation from 0°C to +75°C and is available in plastic and ceramic mini DIP and metal can packages.

PINOUTS



SCHEMATIC



ABSOLUTE MAXIMUM RATINGS

Voltage Between V^+ and V^- Terminals	45.0V	Operating Temperature Ranges:	
Differential Input Voltage	±12.0V	HA-2600/HA-2602	$-55^oC \leq T_A \leq +125^oC$
Peak Output Current	Full Short Circuit Protection	HA-2605	$0^oC \leq T_A \leq +75^oC$
Internal Power Dissipation	300mW	Storage Temperature Range:	$-65^oC \leq T_A \leq +150^oC$

ELECTRICAL CHARACTERISTICS $V+ = +15V$ D. C., $V- = -15V$ D. C.

PARAMETER	TEMP	HA-2600 -55°C to +125°C			HA-2602 -55°C to +125°C			HA-2605 0°C to +75°C			UNITS
		MIN	TYP	MAX	MIN	TYP	MAX	MIN	TYP	MAX	
INPUT CHARACTERISTICS											
Offset Voltage	+25°C		0.5	4		3	5		3	5	mV
	Full		2	6			7			7	mV
Offset Voltage Average Drift	Full		5								μV/°C
Bias Current	+25°C		1	10		15	25		5	25	nA
	Full		10	30			60			40	nA
Offset Current	+25°C		1	10		5	25		5	25	nA
	Full		5	30			60			40	nA
Input Resistance (Note 10)	+25°C	100	500		40	300		40	300		MΩ
Common Mode Range	Full	±11.0			±11.0			±11.0			V
TRANSFER CHARACTERISTICS											
Large Signal Voltage Gain (Note 1, 4)	+25°C	100K	150K		80K	150K		80K	150K		V/V
	Full	70K			60K			70K			V/V
Common Mode Rejection Ratio (Note 2)	Full	80	100		74	100		74	100		dB
Unity Gain Bandwidth Product (Note 3)	+25°C		12			12			12		MHz
OUTPUT CHARACTERISTICS											
Output Voltage Swing (Note 1)	Full	±10.0	±12.0		±10.0	±12.0		±10.0	±12.0		V
Output Current (Note 4)	+25°C	±15	±22		±10	±18		±10	±18		mA
Full Power Bandwidth (Notes 4, 11)	+25°C	50	75		50	75		50	75		kHz
TRANSIENT RESPONSE											
Rise Time (Notes 1, 5, 6 & 7)	+25°C		30	60		30	60		30	60	ns
Overshoot (Notes 1, 5, 6 & 7)	+25°C		25	40		25	40		25	40	%
Slew Rate (Notes 1, 5, 7 & 12)	+25°C	±4	±7		±4	±7		±4	±7		V/μs
Settling Time (Notes 1, 5, 7 & 12)	+25°C		1.5			1.5			1.5		μs
POWER SUPPLY CHARACTERISTICS											
Supply Current	+25°C		3.0	3.7		3.0	4.0		3.0	4.0	mA
Power Supply Rejection Ratio (Note 9)	Full	80	90		74	90		74	90		dB

NOTES:
1. $R_L = 2K\Omega$
2. $V_{CM} = \pm 10V$
3. $V_O < 90mV$
4. $V_O = \pm 10V$
5. $C_L = 100pF$
6. $V_O = \pm 200mV$
7. $V_O = \pm 200mV$
8. See Transient Response Test Circuits & Waveforms Page 2-57.
9. $\Delta VS = \pm 5V$
10. This parameter value guaranteed by design calculations.
11. Full power bandwidth guaranteed by slew rate measurement: $FPBW = S. R./2 \pi V_{peak}$.
12. $V_{OUT} = \pm 5V$

Analog Product Data Book, Harris, 1986

APP NOTE

FOR YOUR INFORMATION

No. 519

Harris Analog

OPERATIONAL AMPLIFIER NOISE PREDICTION

By Richard Whitehead

INTRODUCTION

When working with op amp circuits an engineer is frequently required to predict the total RMS output noise in a given bandwidth for a certain feedback configuration. While op amp noise can be expressed in a number of ways, "spot noise" (RMS input voltage noise or current noise which would pass through 1Hz wide bandpass filters centered at various discrete frequencies), affords a universal method of predicting output noise in any op amp configuration.

THE NOISE MODEL

Figure 1 is a typical noise model depicting the noise voltage and noise current sources that are added together in the form of root mean square to give the total equivalent input voltage noise (RMS), therefore:

$$E_{ni} = \sqrt{e_{ni}^2 + I_{ni}^2 R_g^2 + 4KTR_g} \quad \text{where,}$$

E_{ni} is the total equivalent input voltage noise of the circuit.

e_{ni} is the equivalent input voltage noise of the amplifier.

$I_{ni}^2 R_g^2$ is the voltage noise generated by the current noise.

$4KTR_g$ expresses the thermal noise generated by the external resistors in the circuit where $K = 1.23 \times 10^{-23}$ joules/°K; $T = 300°K$ (27°C) and $R_g = \left(\frac{R_1 R_3}{R_1 + R_3}\right) + R2$

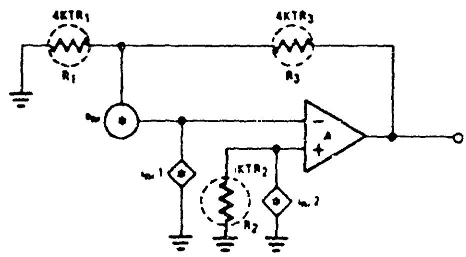


Figure 1

The total RMS output noise (E_{no}) of an amplifier stage with gain = G in the bandwidth between f_1 and f_2 is:

$$E_{no} = G\left(\int_{f_1}^{f_2} E_{ni}^2 df^{1/2}\right)$$

Note that in the amplifier stage shown, G is the non-inverting gain $\left(G = 1 + \frac{R_2}{R_1}\right)$ regardless of which input is normally driven.

PROCEDURE FOR COMPUTING TOTAL OUTPUT NOISE

1. Refer to the voltage noise curves for the amplifier to be used. If the R_g value in the application is close to the R_g value in one of the curves, skip directly to step 6, using that curve for values of E_{ni}^2. If not, go to step 2.

2. Enter values of e_{ni}^2 in line (a) of the table below from the curve labeled "$R_g = 0\,\Omega$".

3. From the current noise curves for the

amplifier, obtain the values of i_{ni}^2 for each of the frequencies in the table, and multiply each by R_g^2, entering the products in line (b) of the table.

4. Obtain the value of $4KTR_g$ from Figure 14, and enter it on line (c) of the table. This is constant for all frequencies. The $4KTR_g$ value must be adjusted for temperatures other than normal room temperature.

5. Total each column in the table on line (d). This total is E_{ni}^2.

	10Hz	100Hz	1KHz	10KHz	100KHz
(a) e_{ni}^2					
(b) $i_{ni}^2R_g^2$					
(c) $4KTR_g$					
(d) E_{ni}^2					

6. On linear scale graph paper enter each of the values for E_{ni}^2 vs. frequency. In most cases, sufficient accuracy can be obtained simply by joining the points on the graph with straight line segments.

7. For the bandwidth of interest, calculate the area under the curve by adding the areas of trapezoidal segments. This procedure assumes a perfectly square bandpass condition; to allow for the more normal -6db/octave bandpass skirts, multiply the upper (-3db) frequency by 1.57 to obtain the effective bandwidth of the circuit, before computing the area. The total area obtained is equivalent to the square of the total input noise over the given bandwidth.

8. Take the square root of the area found above and multiply by the gain (G) of the circuit to find the total Output RMS noise.

A TYPICAL EXAMPLE

It is necessary to find the output noise of the circuit shown below between 1KHz and 24KHz.

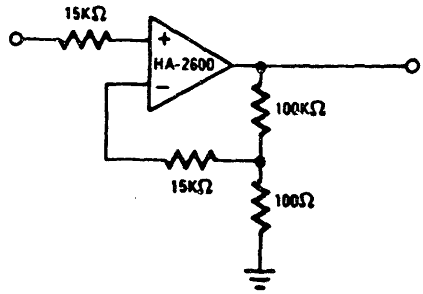
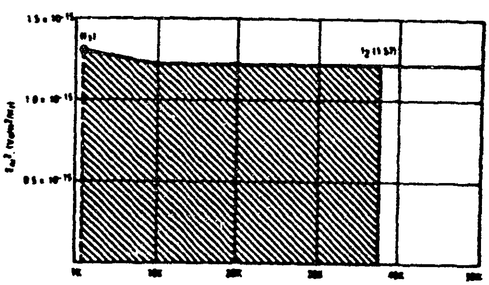


Figure 2
The HA-2600 In a Typical G = 1000 Circuit

Values are selected from Figures 5, 5a and 14 to fill in the table as shown below. An R_g of 30KΩ was selected.

	10Hz	100Hz	1KHz	10KHz	100KHz
(a) e_{ni}^2	3.6×10^{-15}	1.156×10^{-15}	7.84×10^{-16}	7.29×10^{-16}	7.29×10^{-16}
(b) $i_{ni}^2R_g^2$	9.9×10^{-18}	1.89×10^{-16}	3.15×10^{-17}	7.2×10^{-18}	7.2×10^{-18}
(c) $4KTR_g$	4.968×10^{-16}	4.968×10^{-16}	4.968×10^{-16}	4.968×10^{-16}	4.968×10^{-16}
(d) E_{ni}^2	5.08×10^{-15}	1.86×10^{-15}	1.31×10^{-15}	1.23×10^{-15}	1.23×10^{-15}

The totals of the selected values for each frequency is in the form of E_{ni}^2. This should be plotted on linear graph paper as shown below:

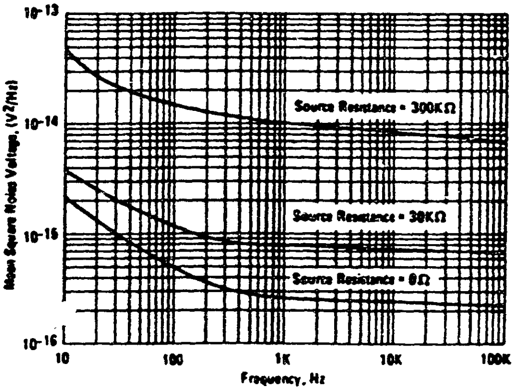


HA-2600 Total Equivalent Input Noise Squared

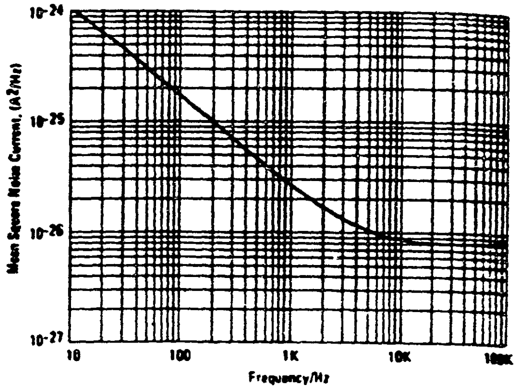
Since a noise figure is needed for the frequency of 1KHz to 24KHz, it is necessary to calculate the effective bandwidth of the circuit. With AV = 60db the upper 3db point is approximately 24KHz. The product of 1.57 (24KHz) is 37.7KHz and is the effective bandwidth of the circuit.

TYPICAL SPOT NOISE CURVES (continued)

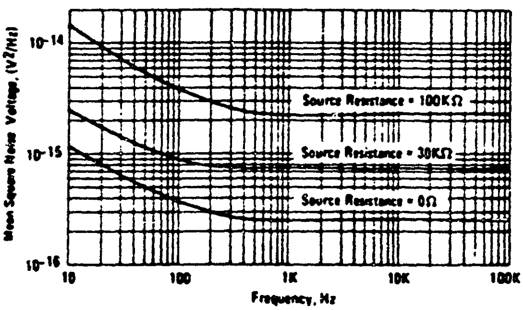
Curve 4
HA-2600/2620 INPUT NOISE VOLTAGE



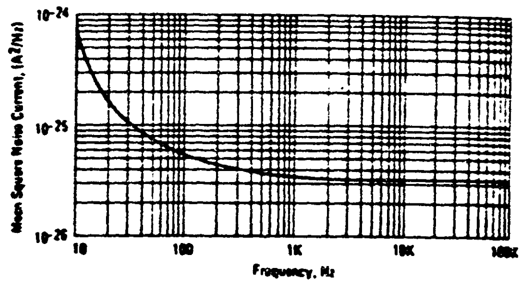
Curve 4A
HA-2600/2620 INPUT NOISE CURRENT



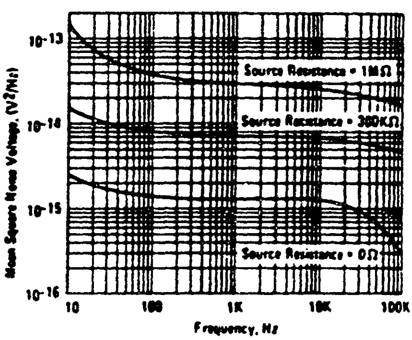
Curve 5
HA-2640/2645 INPUT VOLTAGE NOISE ($V_S = \pm 30V$)



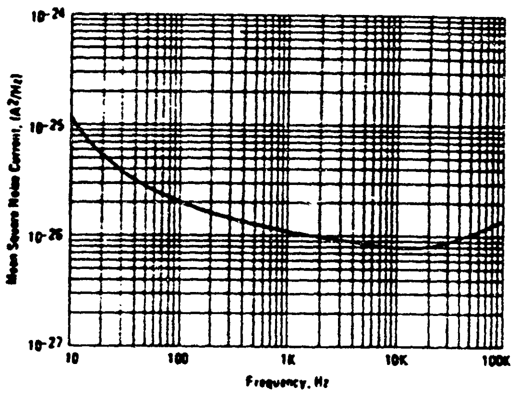
Curve 5A
HA-2640/45 INPUT NOISE CURRENT ($V_S = \pm 30V$)



Curve 6
HA-2700 INPUT NOISE VOLTAGE



Curve 6A
HA-2700 INPUT NOISE CURRENT



Appendix F. ANALYSIS OF CIRCUIT NOISE MODELS

This appendix contains the step-by-step calculation of the noise in all the circuits used in the two systems. The noise models are included once again for easy reference. In addition, nodes have been labeled so that the calculation is easier to follow.

F.1 Noise Model for the Bias Circuit

The noise model for the bias circuit may be seen in Figure F.1. This is the same model that appears in Figure 4.1.2.1. From this model, the noise being injected into the DUT node can be calculated. This noise, by the circuit configuration, is a current noise. To obtain this value, I need to calculate the noise of the circuit from input to output. The noise of the LM399 is filtered with a low pass filter, before it reaches the U2 amplifier. At the input of the amplifier, the noise is

$$E_A^2 = \left|\frac{1}{1 + R_1C_1s}\right|^2 (E_{LM}^2 + E_1^2)$$

$$= \frac{1}{1 + R_1^2C_1^2w^2} (E_{LM}^2 + E_1^2) \qquad \text{(F.1)}$$

The noise then passes through the U2 amplifier, where the input voltage and current noise of the amplifier is added.

$$E_B^2 = E_2^2 + \left(\frac{R_2 + R_3}{R_3}\right)^2 (E_{N2}^2 + E_A^2) + I_{N2}^2R_2^2 + \left(\frac{R_2}{R_1}\right)^2 E_3^2 \qquad \text{(F.2)}$$

The noise is then attenuated by a low pass filter.

$$E_C^2 = \left|\frac{1}{1 + R_4C_2s}\right|^2 (E_B^2 + E_4^2)$$

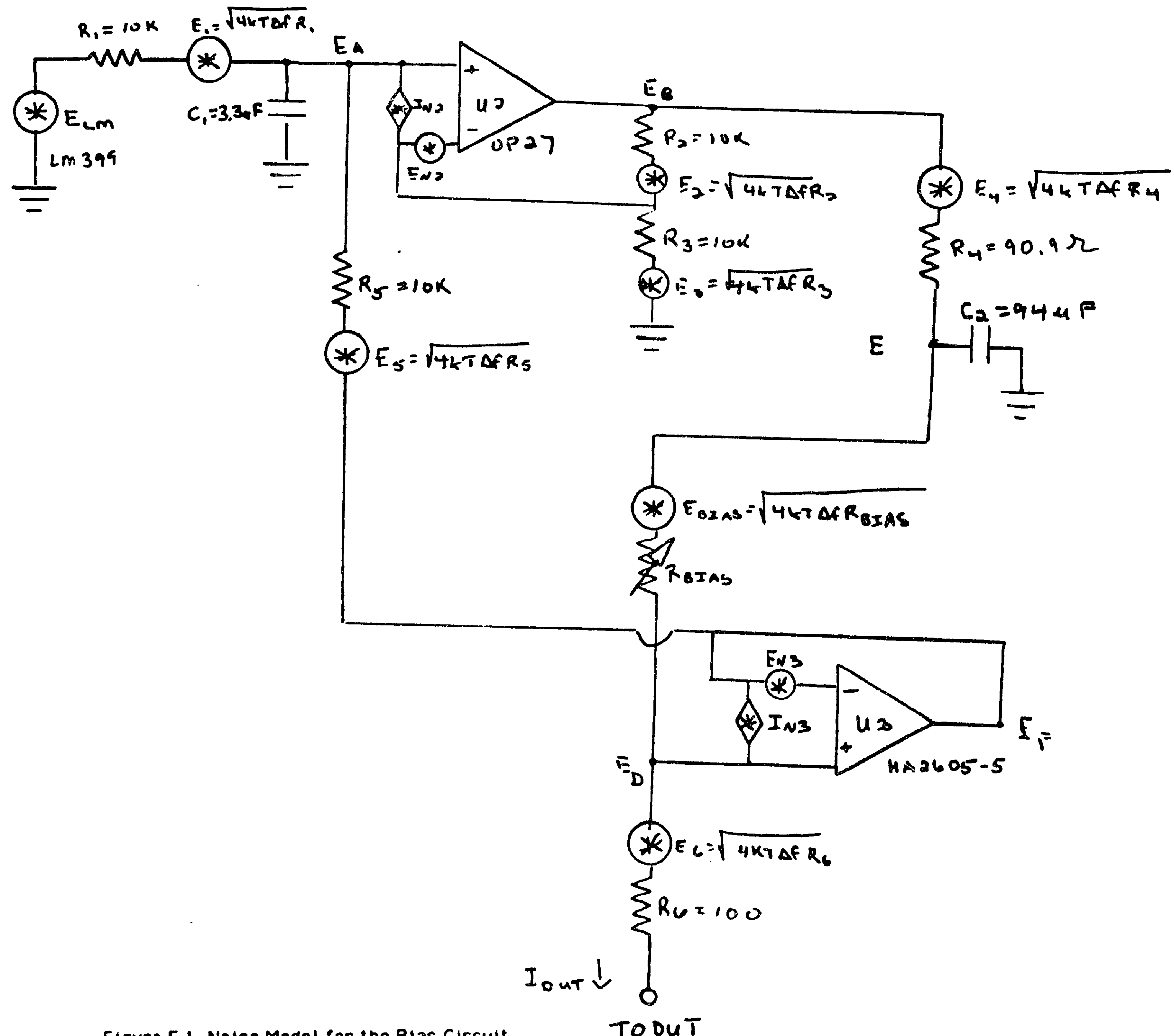


Figure F.1 Noise Model for the Bias Circuit

$$= \frac{1}{1 + R_4^2 C_2^2 w^2} (E_B^2 + E_4^2) \qquad \text{(F.3)}$$

The noise is converted from a noise voltage to a noise current by the bias resistor. At the DUT node, the noise current is equal to

$$I_{DUT}^2 = \frac{E_C^2 + E_{BIAS}^2 + E_6^2}{(R_{BIAS} + R_6)^2} \qquad \text{(F.4)}$$

The next step in the calculation is to put in the various values for the noise quantities and calculate I_{DUT}. The calculation will show which noise sources are dominant and whether the bias circuit contributes significant noise at the DUT node. In these calculations, I will assume a noise bandwidth, Δf, equal to 55 kHz minus 5 kHz, or 50 kHz. I will also assume that the circuit is operating at room temperature, so that 4kT equals 1.61 e-20. For determining the filter magnitude, a frequency of operation is needed, I will use 5 kHz. This will give me the largest magnitude, so I will be calculating the worst case. The specifications for the OP-27 gives the noise in units of $V/\sqrt{Hz}$. I will make this into volts by multiplying by the square root of the noise bandwidth. Then all the noise values being used and being calculated will be in terms of volts or amperes. The bias resistance has a wide range of values from 250 Ω to 2 MΩ. The 2 MΩ bias resistance will result in the largest thermal noise voltage. However, we are converting to a noise current by dividing by the resistance. Dividing by such a large resistance will result in a relatively small current, so I will calculate for the highest noise current that can be expected. This will occur for the 250 Ω resistance.

Table 5 shows the numerical values of the noise sources used in equations F.1 through F.4. Table 6 shows the final results for these four equations.

Table 5 Numerical Values of Noise Sources in F.1 through F.4

Source	Value (V^2)
E_{LM}^2	2.50e-9
$E_1^2 - E_2^2$	8.05e-12
E_{N2}^2	7.22e-13
$I_{N2}^2 R_2^2$	1.80e-12
E_3^2	8.05e-12
E_4^2	7.32e-14
E_{BIAS}^2	2.01e-13
E_6^2	8.05e-14

Table 6 Numerical Values of F.1 Through F.4

Noise Quantity	Value
E_A^2	2.33e-15 V^2
E_B^2	2.08e-11 V^2
E_C^2	2.90e-16 V^2
I_{DUT}^2	1.45e-18 A^2

From Table 5 we see that the noise of the voltage reference is more than two orders of magnitude above the noise of the 10 kΩ. However, the filter attenuates this noise significantly and it is the smallest noise quantity in equation F.2. In equation F.2 none of the terms really dominate over the others. The noise of the two 10 kΩ resistor, the input noise voltage and current of the operational amplifier are about the same order of magnitude. Table 5 shows that the noise at the output of the U2 amplifier is larger than the thermal noise of the 90.0 Ω resistor, but both noise quantities are attenuated by the filter. The attenuation is significant and E_C^2 is the smallest term in F.4. As equation F.2 is written, the thermal noise of the bias resistor is the largest term, but the thermal noise of R_6 is less than one order of magnitude smaller. Converting the voltage noise to a current by dividing by the resistance squared results in a small current noise.

If you assume the shot noise is about 8 e-17 A^2, (This assumes a reverse current of 5 mA and a 50 kHz noise bandwidth) the noise contributed by the bias circuit is almost two orders of magnitude less than the shot noise in this worst case scenario. The shot noise is not large enough to dominate over this bias circuit noise, so the bias noise may have some affect on the accuracy of the noise measurements. To operate at around 5 mA, the bias resistance should be larger than 250 Ω. So the noise current should be a little bit smaller. The shot noise should be large enough to dominate over the bias circuit noise. The final conclusion is that the noise of the bias circuit is not large enough to be considered a problem.

F.2 Noise Model for the Circuit Around the DUT

The noise model for the portion of the circuit around the DUT may be seen in Figure F.2. This model is the same as the one seen in Figure 4.1.2.3. The calculation of the noise of the system assumes that no signal is being applied at the CAL input. This is one of the assumptions that allows all the resistors to be put in parallel. The noise model itself is quite simple, consisting of only the thermal noise of the parallel combination of resistors, R_X, and the noise of the DUT. The noise E_Y is the signal that will be calculated and it is the signal that will be amplified by the three amplification stages. This quantity will indicate if the added thermal noise is lower or higher to the DUT noise.

$$E_Y^2 = \left(\frac{R_X}{R_X + R_{DUT}}\right)^2 E_{DUT}^2 + \left(\frac{R_{DUT}}{R_X + R_{DUT}}\right)^2 E_X^2 \qquad (F.5)$$

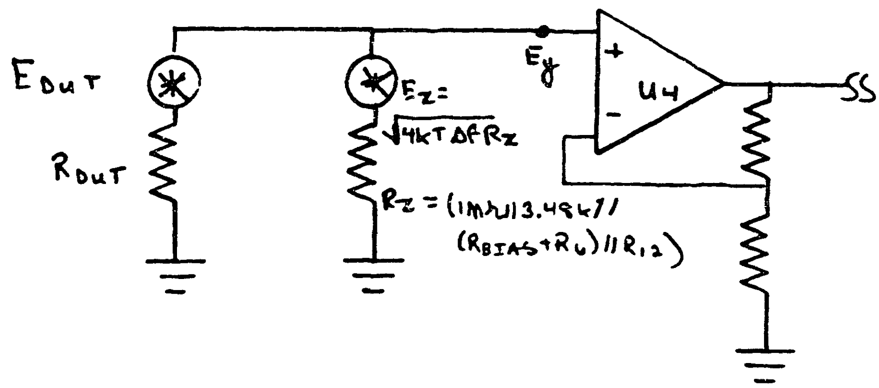


Figure F 2 Noise Model for Circuit Around DUT

To determine the noise of this section, the values of all the noise generators and resistors must be inserted into equation F.5. I will assume that the noise bandwidth is 50 kHz, the bias resistor, R_{BIAS}, is 250 Ω, the resistance of the DUT, R_{DUT}, is 10 Ω. Thus, the parallel combination of resistors, R_X, is equal to 308Ω. The noise of the DUT, E_{DUT}, is the shot noise multiplied by the R_{DUT}. Assuming the shot noise is 8 e-17 A^2, as I did above, the noise voltage, E_{DUT}^2, will be equal to 8 e-15 V^2. The results of the calculation is summarized in Table 7.

Table 7 Numerical Values of F.5

Noise Quantity	Value (V^2)
$EDUT^2$	8.00e-15
E_X^2	2.48e-13
E_Y^2	7.75e-15

From this equation one can see that the DUT noise is attenuated very slightly, while the thermal noise is severely attenuated. This is a desirable effect, because we want the added noise to be as small as possible. The shot noise is not sufficiently high to allow us to ignore the thermal noise. This thermal noise will also be amplified and will appear at the output of the circuit. So we must eliminate this noise quantity from future calculation for DUT noise.

F.3 Noise Model for the Amplification Stages

The noise model for the three amplification stages may be seen in Figure F.3. This is the same model that was seen in Figure 4.1.2.4. From this model, the noise at the output of the circuit, E_{OUT}, can be calculated. The input noise to this section, E_Y is the noise calculated in the previous section. This input noise is amplified by the stages and the noise of the stages is added to it. The following equations describe this amplification and addition process. In these equations, the magnitude of the three high pass filters, formed with a 2.2 μF and a 10 kΩ resistor, are not shown, because the magnitude at 5 kHz is equal to unity. In the U6 amplification stage, which is also a filter, assume that the capacitors are open.

$$E_B^2 = \left(\frac{R_2 + R_3}{R_3}\right)^2 (E_Y^2 + E_{N4}^2) + I_{N4}^2 R_2^2 + E_2^2 + \left(\frac{R_2}{R_3}\right)^2 E_3^2 \tag{F.6}$$

$$E_C^2 = E_B^2 \tag{F.7}$$

$$E_D^2 = \left(\frac{R_5 + R_6}{R_6}\right)^2 (E_C^2 + E_{N5}^2) + I_{N5}^2 R_5^2 + E_5^2 + \left(\frac{R_5}{R_6}\right)^2 E_6^2 \tag{F.8}$$

$$E_F^2 = E_D^2 + E_7^2 \tag{F.9}$$

$$E_G^2 = \left(\frac{R_9 + R_{10}}{R_{10}}\right)^2 (E_F^2 + E_{N6}^2) + I_{N6}^2 R_9^2 + E_9^2 + \left(\frac{R_9}{R_{10}}\right)^2 E_{10}^2 \tag{F.10}$$

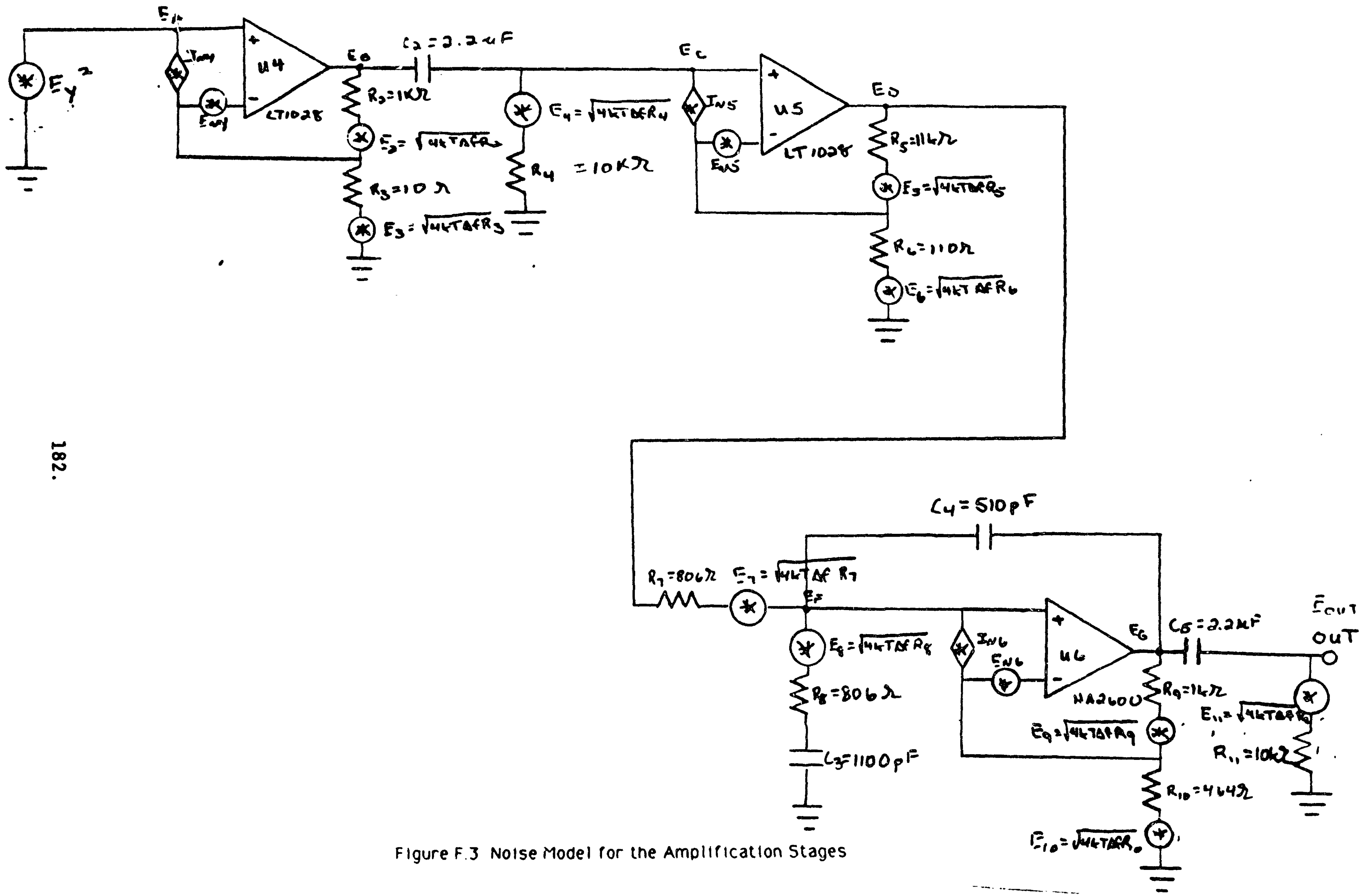


Figure F.3 Noise Model for the Amplification Stages

$$E_{OUT}^2 = E_G^2 \tag{F.11}$$

To identify noisy components and determine whether the noise produced by the amplifiers is large enough to swamp the noise being amplified, the actual noise values must be substituted into equations F.6 through F.11. As with the previous models, I assumed that the noise bandwidth, Δf, is 50 kHz and that the circuit is operating at room temperature. The noise specifications for the LT1028 and the HA2600 op amps, which may be found in Appendix E, are given in units of $V/\sqrt{Hz}$ and $A/\sqrt{Hz}$. These values will be converted to units of volts or amperes by multiplying by the square-root of the bandwidth. This way all noise values will be in the same units. Table 8 gives the numerical values for the noise sources used in equations F.6 through F.11. Table 9 shows the final results of the calculation.

Table 8 Numerical Values of Noise Sources in F.6 through F.11

Source	Value (V^2)
E_{N4}^2	6.05e-14
$I_{N4}^2 R_2^2$	1.28e-13
E_2^2	8.05e-13
E_3^2	8.05e-15
$I_{N5}^2 R_5^2$	1.55e-11
E_5^2	8.86e-12
E_6^2	8.86e-14
E^{72}	6.49e-13
E_{N6}^2	3.92e-11
$I_{N6}^2 R_9^2$	1.75e-15
E_9^2	8.05e-13
E_{10}^2	3.74e-13

Table 9 Numerical Values of F.6 Through F.11

Noise Quantity	Value (V^2)
$E_B^2 - E_C^2$	7.78e-10
$E_D^2 - E_F^2$	7.94e-6
$E_G^2 - E_{OUT}^2$	7.91e-5

Equation F.6 shows the effect of the first amplification stage on the input noise. The input noise is amplified as desired. The only problem is the input noise also contains some thermal noise from the parallel combination of resistors, this noise is also amplified and will reach the output of the system. We will have to eliminate this noise quantity from any calculations of DUT noise. The input noise voltage of the LT1028 is also amplified, by this stage. In addition, the noise of the 10 Ω resistor is multiplied by the ratio of R_2 to R_3 squared. Of these two extra noise contributing terms, the input noise of the LT1028 is the most significant. The rest of the equations, F.7 through F.11, show that none of the noise from the second and third amplification stages is added to the overall noise. This occurs because the noise at the output of the first amplification stage is substantially higher than any other noise source subsequently added. Thus, the only effect the second and third amplification stages have on the noise is to amplify it. The output noise, E_{OUT}, consists of the noise from three sources all amplified. These three sources are the input noise, E_X, the input noise voltage of U4 and the thermal noise of E_3. Of these three sources, E_X is the largest. The unwanted thermal noise contained in E_X will have to be eliminated in calculations for the DUT.

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