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LOCK-IN AMPLIFIER DIGITAL TECHNIQUES FOR D* SIGNAL-TO-NOISE CHARACTERIZATION OF INFRARED DETECTORS

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DEFINITION OF SYMBOLS

H = irradiance, w/cm^2, (sinusoidal fundamental)[1]

A = area of detector, cm^2

P = radiant power falling on detector, w

f = modulation frequency, Hz, (chopping rate)

Δf = equivalent noise bandwidth of measurement system, Hz

$\Delta f'$ = ENBW for the *average* of N signal samples, Hz

V_s = signal output of detector, V rms, (rti) (denoted as e_s in other ITHACO literature)

E_s = signal output of system, V dc, (rto)

V_n = detector noise within electrical bandwidth Δf, V rms, (rti)

e_n = spectral noise density of detector at f, V rms/$\sqrt{Hz}$, (rti)

E_n = output fluctuation superimposed on E_s, V rms (rto noise)

G = LIA gain, full scale output/full scale input

T = LIA time constant, sec

SNDR = Signal to spectral noise density ratio, $\sqrt{Hz}$ (alternatively expressed in dB = 20 $\log_{10}$ ratio)

R = responsivity of detector, V rms/w

NEP = noise equivalent power of detector, w/$\sqrt{Hz}$

D* = detectivity of detector material, cm $\sqrt{Hz}$/w

N = number of digitized samples taken to measure E_s and E_N

Q_n = quantization step size of ADC

r_s = sampling rate, Hz

t_o = sampling interval of integrating A-to-D converter, sec

t_m = total measurement time for N samples, sec

σ_x = reproducibility of signal measurement (fractional standard deviation, 68% confidence level)

σ_n = reproducibility of noise measurement (fractional standard deviation, 68% confidence level)

σ_{D*} = reproducibility of D* or NEP measurement

rto = referred-to-output of LIA (measurement system output voltage)

rti = referred-to-input of LIA (actual detector voltage)

MATHEMATICAL RELATIONSHIPS

Eq. 1 $P = HA$

Eq. 2 $V_n = e_n \sqrt{\Delta f} = R\ NEP \sqrt{\Delta f}$

2A $e_n = R\ NEP$

Eq. 3 $R = V_s / P$

Eq. 4 $SNDR = V_s / e_n = V_s \sqrt{\Delta f} / V_n = P/NEP$

4A $\frac{V_s}{e_n} = \frac{P}{NEP}$

Eq. 5 $NEP = P(V_n / V_s)(1/\sqrt{\Delta f})$

Eq. 6 $D^* = \sqrt{A}/NEP$

1. See reference #1, pg 325, for determination of H under chopper modulation. Typically will be only 40% of the unmodulated optical power.

Eq. 7 $r_s = 1/t_o$

Eq. 8 $t_m = N\, t_o$

Eq. 9 $\sigma_x = E_n / E_s = V_n / V_s = NEP \sqrt{\Delta f'} / P$

Eq. 10 $\sigma_n \sim \sqrt{1/(2N)}$, for $r_s \leq 2\, \Delta f$ (Nyquist limit)

Eq.10A $\sigma_n \sim \sqrt{1/(4N)}$, for two phase LIA if both channels are sampled

Eq. 11 $\sigma_{D*} = \sqrt{\sigma_x^2 + \sigma_n^2}$

Eq. 12 $E_s = G\, V_s$

Eq. 13 $E_n = e_n\, G \sqrt{\Delta f}$

Eq. 14 $\Delta f \sim 1/(2t_o)$, for $t_o > 10\, T$

Eq. 15 $\Delta f' \sim 1/(2t_m)$, for $t_m > 10\, T$

Eq. 16 $Q_n = 100 \mu V / t_o$ for Model 385 integrating ADC

INTRODUCTION

This article explains how to make either simultaneous or separate signal and noise measurements on an IR detector operating under chopper-modulated illumination. By this means a *noise equivalent power* (NEP) figure of merit can be obtained for the detector under test. This useful and accepted specification indicates approximately the lower limit at which signal can be distinguished from random noise generated in the detector by specifying the level of illumination at which the rms electrical signal output (sinusoidial fundamental thereof) equals the rms electrical spectral noise density at the modulation frequency. Since for most detectors, the noise is proportional to the square root of the sensing area, a corollary figure of merit which expresses the detectivity of the photosensitive material can be written if the detector area is known. [2] Called D* (pronounced "dee-star"), this intrinsic property of the material equals $\sqrt{A}/NEP$ and is independent of detector size.

In measuring NEP and D*, the electrical frequency (e.g., chopping rate) and source spectrum (e.g., blackbody temperature or monochromator wavelength) must be specified to conform to certain standardized conditions given in the literature. A typical such test, denoted D* (500° K, 900) for example, calls for a 500° K blackbody IR source and a 900 Hz chopping frequency. Sometimes a numeral "1" is included within the parentheses, for example, D* (500° K, 900, 1) to indicate 1 Hz bandwidth, but this is redundant since regardless of the actual bandwidth in effect during the determination of NEP and D*, the results should be normalized to 1 Hz. [3] In the interests of speed and accuracy, the measurements are conducted with the signal level considerably above the detector noise floor. The results are then linearly extrapolated to the threshold of detection acccording to the defining equation for NEP. [4]

The topic of measurement bandwidth Δf often leads to confusion and merits additional discussion. To begin with, Δf refers to the equivalent noise bandwidth of the system and *not* its 3 dB frequency response corner(s). In actual practice it is usually chosen to be wider than 1 Hz for two reasons. First, commonly used wave analyzers available to pioneering workers a quarter century ago for noise measurements typically had fixed bandwidths (e.g., 5 Hz) tunable over a wide center frequency range. [5] Second, in order to achieve a given standard deviation for fluctuations of the reading (i.e., measurement reproducibility), the measurement speed in any narrowband system will be directly proportional to bandwidth. Using Δf = 5 Hz for example, would be 5 times faster than using Δf = 1 Hz. Since the noise density spectrum of an IR detector exhibits a gradual slope, the ratio of bandwidth to modulation frequency $\Delta f/f$ can be made relatively wide (e.g., 10% or 20%) with excellent accuracy.[6] The "extra" noise on one side of the frequency of interest balances the "deficiency" on the other. The point is this: one is not constrained to a 1 Hz bandwidth, but rather can tailor Δf to any value one wishes when characterizing detectors, consistent with considerations of prevailing discrete frequency interference (e.g., line power harmonics), noise density profile and instrumentation limitations. For more details on noise bandwidth and the effects of noise, see reference #1, Sections 8.2 and 8.3.

Classical detector characterization methods involve taking two separate measurements; noise being obtained under quiescent conditions (e .g., dark detector) and signal being obtained under modulated illumination. This is done because conventional laboratory instrumentation using wave analyzers cannot make both measurements simultaneously. The procedure is inconvenient and subject to calibration and drift error. Drift of chopper speed relative to passband center frequency, in particular, has been a perennial source of trouble in measuring signal strength V_s. [7]

2. Reference #1, pg. 269
3. Reference #1, pg. 270
4. Reference #2, pg. 270
5. Reference #1, pg. 339; Reference #3, pg. 312, 313; Reference #4, pg. 11-59
6. Reference #2, pg. 270; Reference #3, pg. 303, 304
7. Reference #1, pg. 327

USE OF LOCK-IN AMPLIFIERS TO CHARACTERIZE DETECTORS

The Lock-In Amplifier (LIA) gives the user a choice of making signal and noise measurements either simultaneously or separately. A lock-in amplifier can measure spot noise (noise density at a given frequency) because it works on a narrowbanding principle. The noise in the band near the reference (modulation) frequency passes to the output where it appears as random fluctuations superimposed on the dc (signal) output. This effect is independent of signal and therefore is valid even under dark or unmodulated illumination conditions, provided the lock-in has an internally or externally supplied reference signal to set the measurement center frequency. Noise can be measured by including a rms meter in the output circuitry of the LIA, such as the analog Noise Option 01 available for ITHACO equipment. However much higher performance which is much better suited for detector characterization can be obtained using digital rms measurement techniques as described below.

The use of a dual-channel, heterodyning LIA with digital sampling capability such as the ITHACO DYNATRAC® Model 3990 improves on the historical methods in several respects:

1. The heterodyning concept allows the system passband to precisely track fluctuations in modulation frequency due to chopper motor speed instability. Some truly awful choppers are in use, often built into blackbodies. A DYNATRAC® instrument will obtain correct data from such a chopper.

2. Digital noise measurement works over the full dynamic range and frequency range of the LIA. A single instrument, the ITHACO Model 3990 system consisting of a Model 399 Lock-In Amplifier and a Model 385EO Integrator/Coupler for GPIB data transfer, will measure signal and noise at chopping rates from below 1 Hz to above 100 kHz and will handle SNDR from 1 to 10,000 (80 dB range).

3. System bandwidth Δf is precisely known and easily changed over the range 25 Hz to 0.001 Hz.

4. Computer interfacing allows convenient automated data collection as well as control over the critical parameters of LIA sensitivity and system bandwidth.

5. Reproducibility of signal and noise measurement is calculable and controllable. There is no guessing involved as to the accuracy of results. The user may trade time for accuracy at will.

6. Signal and noise follow exactly the same electrical path. There exists no calibration uncertaintly due a multiplicity of instruments sharing in the measurements. Signal and noise measurement errors as well as detector drift are cancelled ratiometrically for D* and NEP.

7. Results are obtained very rapidly due to simultaneous measurement of signal and noise. Furthermore, a dual-channel LIA can acquire a noise measurement twice as fast as a single channel instrument. This is because the noise in each channel (ASINØ and ACOSØ) is uncorrelated and therefore both channels can be sampled simultaneously to double the sampling rate.

8. The LIA responds only to the sinewave fundamental of the detector ouptut signal. Waveshaping and prefiltering would be redundant. The LIA is also highly efficient at eliminating high levels of common mode and differential mode interference.

9. A dual-channel LIA can serve as a spectrum analyzer for analysis of interfering frequencies which are not necessarily coherent with the reference input. Sweeping the reference frequency will allow the accurate measurement of troublesome line frequency harmonics near the modulation frequency, for example.[8]

10. Very high input sensitivity will allow direct connection to the detector in many cases. In Model 391A, 393 and 399 DYNATRAC® instruments, this is facilitated by having a fully floating, quasi-differential front end operating from an isolated power supply, which incorporates a driven transformer shield. When a preamp is required for impedance matching or driving long cables, ITHACO has available a series of compatible models. These can operate from the isolated power supply to prevent ground loop and isolation woes.

11. Since the microprocessor-based 385 ADC resides in a completely separate chassis and employs optical isolation of its analog and digital circuitry, digital switching noise cannot find its way into the sensitive LIA electronics

Thus a relatively compact and reasonably priced instrument, the Model 3990 LIA, combines the functions of high gain amplifier, wave analyzer, rms meter and computer interface in a single package. It affords the power, convenience and flexibility needed for superior characterization of detectors, not only for NEP and D* but also for responsivity, spectral response, frequency response, noise spectra, and so forth.

8. Reference #5, Section 3.5.2.2

DIGITAL LIA MEASUREMENT TECHNIQUES

Digital measuring methods involve repeatedly sampling the LIA dc output. The average of the N samples, E_s, represents the referred-to-output (rto) signal of the LIA. The standard deviation of the samples represents the rms rto noise, E_n, for the bandwidth of the system. The system equivalent noise bandwidth Δf depends on both the LIA Time Constant T and the sampling time t_o of the digitizer that operates on the LIA output. ITHACO DYNATRAC instruments employ an integrating type of D-to-A conversion, and thus t_o can have a significant effect on the overall Δf. The sampling time can be increased as needed in increments of 1/60 of a second. This can be very important in the process of filtering non-random interference [9] or to provide finer resolution of noise when SNDR is on the order of 60 dB or larger (up to a practical limit of about 1 in 2^{16} quantization error).

The gain of the system G equals (LIA full scale dc volts out) divided by (LIA full scale rms input sensitivity): Δf can be readily calculated by a computer.[10] Thus the raw rto data can be scaled to LIA referred-to-input (rti) values for signal and noise as given by Eq. 12 and 13. Δf can be estimated using eq. 14 (will hold to very good accuracy for $t_o > 40T$).

The wider we can make Δf, the faster we can accumulate N samples without statistical over-sampling due to exceeding the Nyquist rate. Sampling too rapidly relative to the LIA Time Constant T would lead to a less accurate noise measurement than predicted by Eq. 10 or Eq. 10A, due to autocorrelation. (Empirical tests show that over-sampling by a factor of $\sqrt{2}$ times the Nyquist rate - i.e. $rs = 2\sqrt{2}\,\Delta f$ - can be tolerated before auto-correlation effects begin to effect reproducibility).

The user must determine the values of N, t_o and T which will yield the best compromise for the speed/accuracy tradeoff. Typically one would want to set the ADC sampling interval to dominate the system bandwidth Δf by having $t_o \geq 10T$. This can be done if good shielding and grounding practice has been employed to minimize discrete frequency interference, allowing the LIA to operate with a small time constant. With the ADC thus dominant, the sampling rate inherently equals the Nyquist rate and one obtains the theoretical maximum speed of noise measurement. The speed of noise measurement in most cases will determine the speed with which a detector figure of merit can be obtained.

To achieve the fastest D* or NEP determination when simultaneously measuring signal and noise, one must select an illumination level large enough so that the signal reproducibility is on the order of ten times better that of the noise. This will occur at a signal level V_s approximately 40 dB above the noise floor e_n (e.g., SNDR = P/NEP = 100). This is also in accordance with standard practice by workers in the IR field when measuring signal and noise separately. Justification for this is that a detector typically exhibits good linearity up to at least 60 dB above the NEP level. [11] Furthermore this level of SNDR (40 dB) lends itself readily to simultaneous signal and noise measurement by a 3990 system with freedom from secondary noise error sources which get progressively more troublesome at the higher SNDR levels, such ADC quantization noise, chopper speed instability and light source level fluctuations. The nominal 40 dB SNDR figure leaves a range of ± 10 dB over which individual devices can vary without sacrificing accuracy or speed. For SNDR at the 20 or 60 dB level, simultaneous measurements are still relatively convenient and fast, although speed and accuracy tend to suffer moderately. Pushing SNDR to extremes (e.g., 0 dB or 80 dB) can be done if one is careful, however the technique is more suitable for the research lab than for reliable production testing of devices.

9. Reference #6, Figure 2 and related text
10. Reference #5, Appendix A
11. Reference #1, pg. 268, 359

A REPRESENTATIVE DETECTOR TEST

Figure 1 depicts a test setup for characterization of a photoconductive detector. Assume the detector has the following typical characteristics:

$$R = 3 \times 10^5 \text{ v/w}$$

$$A = 0.1 \text{ cm} \times 0.1 \text{ cm} = 0.01 \text{ cm}^2$$

$$D^* = 10^{10} \text{ cm} \sqrt{\text{Hz}}/\text{watt (blackbody)}$$

Then from Eq. 6 & 2

$$NEP = \sqrt{A}/D^* = 10 \text{ pw}/\sqrt{\text{Hz}}$$

$$e_n = R\,NEP = 3 \ \mu\text{V}/\sqrt{\text{Hz}}$$

For greatest measurement speed and accuracy we want SNDR = V_s/e_n to equal 100 (40 dB). Therefore we size the blackbody aperture and distance from the detector plane to deliver $V_s = 100 \times e_n = 300\ \mu\text{V}$ signal to the LIA.

$$P = V_s/R = 1 \text{ nw of radiant power}$$

To allow adequate headroom for signal level variations we set the LIA for a 1 mV full scale sensitivity ($G = 10V/1mV = 10^4$). From Eq. 12 and 13, the output signal and noise of the 3990 system will be given by:

$$E_s = V_s G = 3 \text{ V dc}$$

$$E = e_n\ G\sqrt{\Delta f} = 30 \text{ mV rms} \sqrt{\Delta f}$$

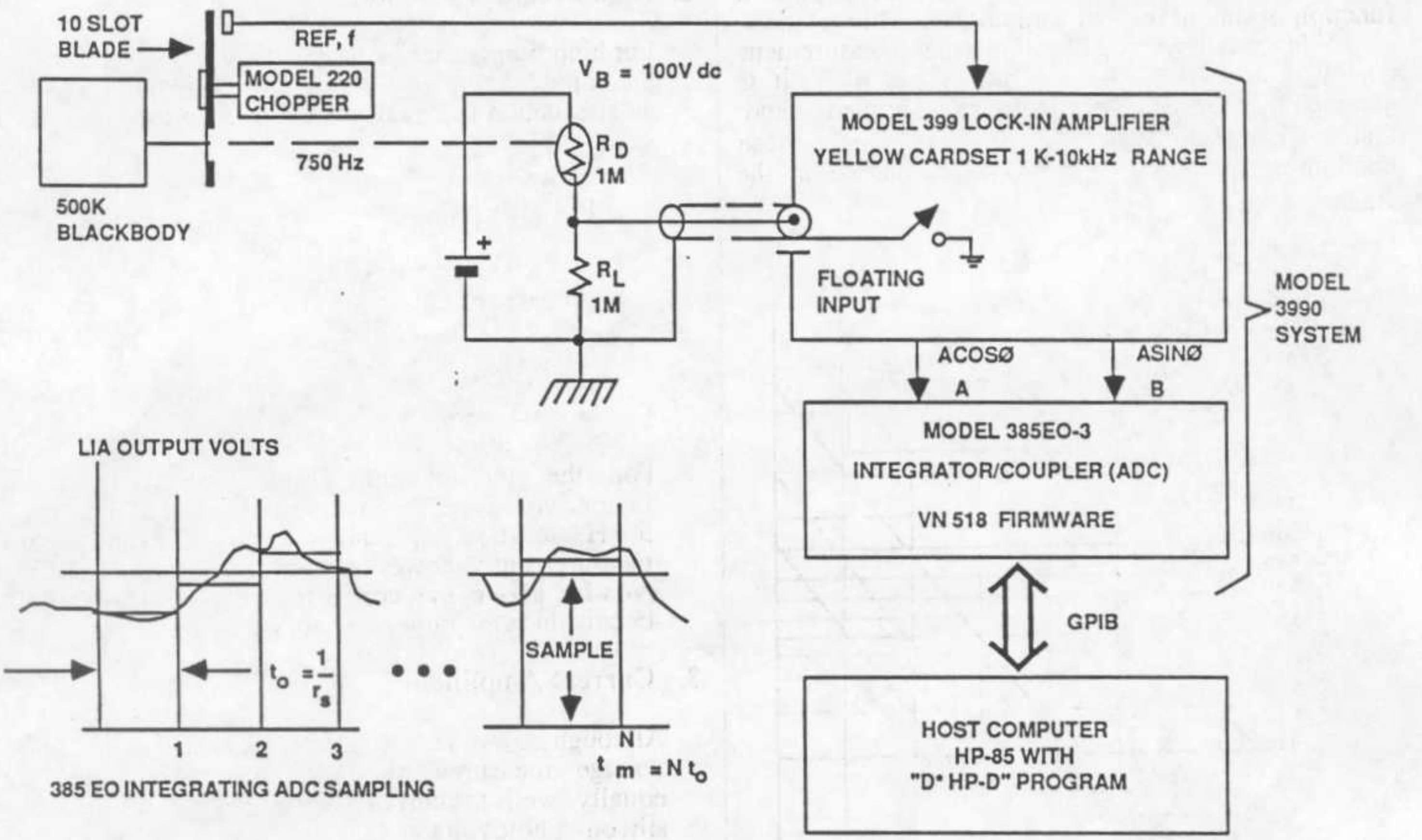


Figure 1 D* MEASUREMENT SETUP USING MODEL 3990 LOCK-IN AMPLIFIER

The relatively high level of noise will allow us to sample both channels for noise and to run the LIA system bandwidth "wide open" using the fastest possible Time Constant, (T = 1.25 msec) and shortest possible ADC sampling interval (t_o = 16.67 msec). Under these conditions, Δf equals 26 Hz as viewed at the GPIB data output of the Model 3990 system and therefore:

$$E_n = 30 \text{ mV} \sqrt{26} = 153 \text{ mV rms}$$

$$E_n \text{ (p-p)} = 6\, E_n = 918 \text{ mV, peak to peak}$$

The expected quantization Q_n for t_o = 16.67 msec from Eq. 15 is .1 mV/t_o = 6 mV p-p, and will have an inconsequential effect on the noise measurement. Lets assume we choose to take 600 samples to measure D*. Then from Eq. 8, 9, 10A and 15 we have:

$$t_m = N\, t_o = 10 \text{ seconds}$$

$$\sigma_n = \sqrt{1/(4N)} \times 100\% = \pm 2.04\%$$

$$\sigma_x = (\sqrt{1/2Nt_o} \text{ /SNDR}) \times 100\% = \pm 0.224$$

We assert that the composite uncertainty of D* is mathematically analogous to cascading two noisy amplifier stages. The reproducibility of the ratio of V_n / V_s used to calculate D* is therefore the rms sum of the noise and signal reproducibilities σ_n and σ_x (Eq. 11). This means that under the conditions given above (SNDR=P/NEP=100) we can make D* measurement with an uncertainty of σ_{D*} = 2.05% at the 0.68 confidence level in just 10 seconds (±6.16% at the 3σ, 0.997 confidence level).

If we were to decrease the illumination to 0.1 nw for a SNDR = P/NEP = 10 (e.g., 20 dB) we would find that it would take 22 seconds to get a comparable reproducibility (σ_n = 1.38%, σ_x = 1.50%, σ_{D*} = 2.03%). On the other hand,were we to increase the illumination to 10 nw (60 dB SNDR) and kept t_m = 10 seconds, we would find virtually no improvement in D* reproducibility (σ_n = 2.04%, σ_x = 0.022%, σ_{D*} = 2.04%).

The HP-85 host computer program called "D* HP-D" measures e_n and V_s. It reports the exact bandwidth Δf and the exact reproducibility values of σ_x, σ_n and σ_{D*}. Given P and A, it then calculates NEP and D*.

Figure 2 depicts approximate D* reproducibility as a function of the number of samples, N. This relationship holds nearly exactly for the usual measurement conditions of SNDR ≥ 30 and $t_o \geq 10$ T. It is assumed that both LIA channels are sampled and that the LIA rolloff is 12 dB/octave. Under these conditions $\sigma_{D*} \sim \sigma_n \sim \sqrt{1/(4N)}$, as charted in the figure.

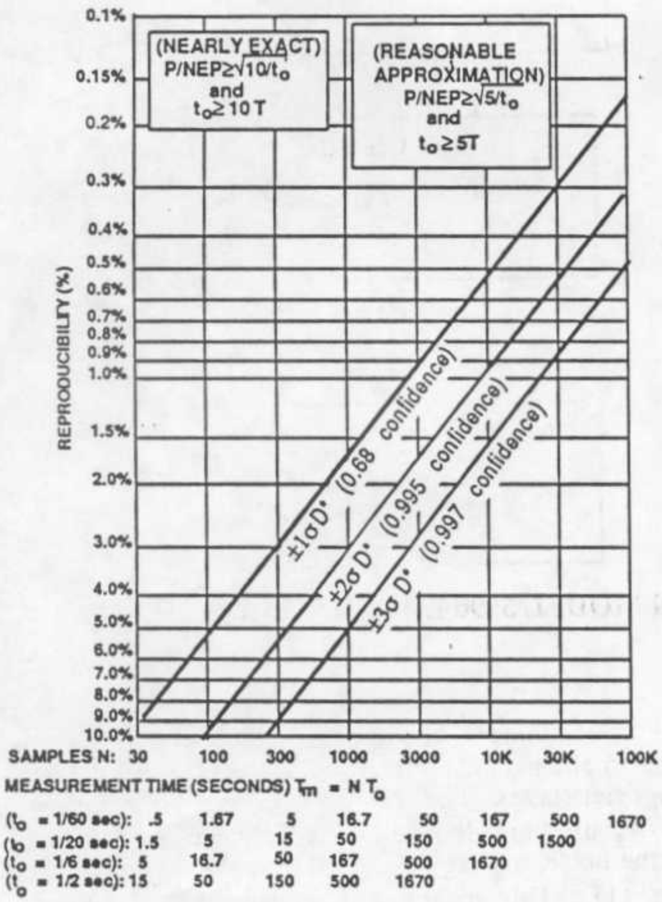


FIGURE 2 D* REPRODUCIBILITY VERSUS NUMBER OF SAMPLES

APPLICATIONS COMMENTS

1. Amplifier Noise

For high impedance sources the detector noise will overwhelm the LIA referred-to-input self noise. The noise adds vectorially so that even a 10:1 detector to LIA noise ratio would result in only a 1/2% error. Models 391A, 393 and 399 have very low current noise as well as very low voltage noise, which is important in the direct connection to higher impedance detectors. For low impedance sources a matching preamp (Model 166) may be needed.

2. High Frequency Rolloff

For high impedance sources, stray capacitance in the input circuit on the order of 75 pF will cause an attenuation in signal and noise given by:

$$\text{Rolloff Factor} = \sqrt{\frac{1}{1+(2\pi fRC)^2}}$$

R = Source impedance = $R_D R_L / (R_D + R_L)$

C = LIA input + detector + cable capacitance
(40pF) (10pF) (25pF)

For the given example (Figure 1), this rolloff factor would be 0.985 at 750 Hz and 0.647 at 5 kHz for both e_n and V_s. The NEP and D* measurement, however, remains mains valid even far above the corner frequency of the input circuit since the ratio of e_n to V_s is unaffected.

3. Current Amplifiers

Although we've restricted our discussion to voltage measurements, the discussion applies equally well to current output devices such as silicon photovoltaic detectors. These devices often require the insertion of current preamplifier in the signal path. Its transimpedance R_f defines its current to voltage scaling. To get a referred-to-input signal or noise current measurement, the LIA rti voltage readings must be scaled as follows:

$$I_s = V_s / R_f$$

$$i_n = e_n / R_f$$

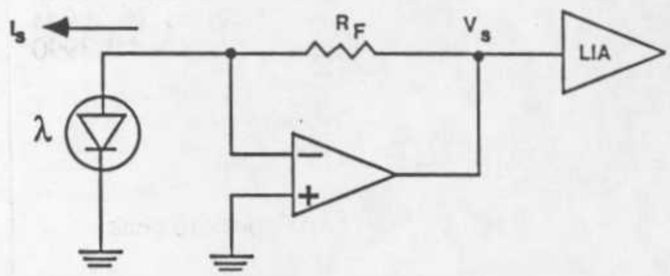


FIGURE 3 CURRENT PREAMPLIFIERS WITH PHOTOVOLTAIC CELL

The measured value of i_n will contain significant contributions due to the thermal noise of R_f and due to preamplifier input current and voltage noise. Calculation will be needed to determine the actual noise generated by the detector. See reference #7 for details.

4. Chopper Quality

Motor short term speed fluctuation can cause significant error when measuring signal and noise simultaneously. The motor jitter will show up as random phase noise which could mask the actual detector noise. This effect would be non-existent if no signal is present (e.g. dark noise and signal are measured separately). Blade jitter due to aperture edge irregularity, on the other hand, will have negligible effect upon D* measurements.

The ITHACO Model 220 Chopper is an excellent choice, as the motor speed excursions are small enough and slow enough to be tracked without error by a DYNATRAC ® Lock-In Amplifier. The Model 230 Chopper is less suitable because its smaller diameter blade performs a less effective flywheel function.

Reference #5, Section 5.2, describes the results of realistically simulating signal and noise measurement in a chopper optical system using a Model 3990 LIA and a Model 220 Chopper.It presents the experimental determination of motor jitter effects, as well as ultimate limits of operation. Note that as SNDR increases beyond 60 dB, the chopper phase noise will eventually force us to phase the LIA and obtain data from a single channel, doubling D* measurement time. In high SNDR situations, the use of two identical blades on the Chopper to increase the inertia will be helpful, as it reduces both the amplitude and the frequency of the motor speed fluctuations.

5. LIA Reference Tracking Speed

For simultaneous signal and noise measurement, the phase locked loop (PLL) in the LIA reference circuit must be able to track a fluctuating reference frequency. The faster the better, since this greatly increases the SNDR range over which the LIA can be operated dual channel. By using an ITHACO LIA which employs plug-in cardsets, such as the Model 391A, 393 or 399, the speed of tracking can be optimized for the modulation frequency employed, affording a dramatic improvement in dynamic tracking accuracy. One should always use the highest possible reference frequency range of the lock-in amplifier. For example, under the condition described in Section 5.2 of reference #5, better results were obtained by replacing the ORANGE cardset with the higher frequency YELLOW cardset and operating in the nominal 1 kHz - 10 kHz range. When this was done the Model 230 performance was nearly as good as that of the Model 220 Chopper. One could manually run the chopper speed up and down 10% or during sample acquisition and discern no loss in noise measurement accuracy.

6. Blade And Speed Choice For Chopper

The number of slots in the blade has no direct bearing on observed phase jitter due to motor instability. The number of slots, however, should be chosen so that the motor runs between 33% and 100% of full speed for best speed jitter performance. At 90 Hz for example, one ought to use a 2-slot blade spinning at 45% of full speed (2700 rpm).

7. Considerations When Operating Below 500 Hz

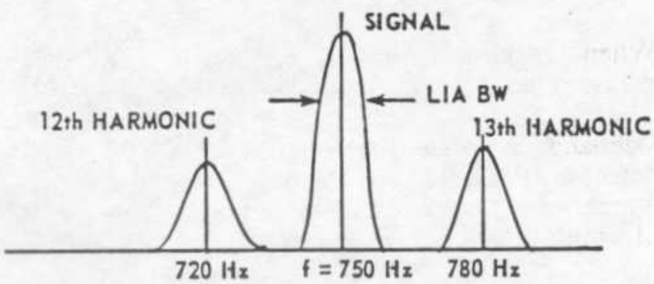
When making *simultaneous* signal and noise measurements below 500 Hz, particularly if the SNDR is 50 dB or greater, the upper sideband component of the LIA phase sensitive detector (PSD) may become troublesome. It will appear as a sinusoid superimposed on the dc LIA output with a frequency double that of the modulation frequency, f (view on oscilloscope). The effect can be large enough to mask the noise to be sampled if one attempts to operate the system with maximum bandwidth (LIA T and ADC t_o set to minimum values).

To solve this problem, one must decrease the system bandwidth below its nominal 30 Hz "wide open" value by a ratio roughly proportional to the ratio f/500 Hz. The penalty is proportionally slower measurement time when T and t_o are increased above the minimum. Thus it may take five times as long at 90 Hz as at 500 Hz to obtain the same accuracy if the noise level is low. Moral: decrease SNDR to the 20 to 40 dB range when doing low frequency work. See Appendix B.

One also may use the Model 385 integrating ADC to very good advantage in the suppression of the spurious 2f output. By setting the LIA T and ADC t_o to yield approximately the same bandwidth (will be equal when $t_o = 4T$) an 18 dB/octave attenuation above the corner frequency will be obtained. This affords us a much more effective rejection of the 2f component while maintaining a given Δf than could be obtained using the LIA with an ordinary fast-sampling ADC. The payoff is an order of magnitude faster measurement when working at low chopping rates. The technique is the same as would be used to suppress an output beat frequency waveform resulting from an interfering frequency lying near f (see reference #6 scope traces and text). Note that by having t_o span a cycle or so of the spurious ac LIA output, one can achieve accurate results even when the output ripple exceeds the noise (viewed at the 385 input).

8. Discrete Frequency Interference

Interfering components lying within the system bandwidth can cause error similar to that produced by the 2f effect discussed above. On a scope the LIA output will exhibit a sinusoidal beat frequency superimposed on the dc signal and noise (e.g. 30 Hz if f=750 Hz and the interference at 780 Hz). Again the solution lies in bandwidth restriction and again the penalty is slow measurements. The usual culprit in low level work is line frequency harmonics, which can never be more than 30 Hz away from f. Figure 4 shows the results of using a dual channel LIA to analyze for possible trouble of this sort.



399 LIA INT MODE
(sweep from 700 Hz to 800 Hz)
T = 12.5 msec (10 Hz B
A (vector sum) OUTPUT.
DETECTOR MODULATED AT FIXED 750 Hz.

FIGURE 4 USE OF A VECTOR SUM LIA TO DETECT INTERFERENCE

Large interferences as shown will result in extremely slow measurements, with the time strongly dependent on desired accuracy, level of interference and nearness of the interfering frequency to f. *One therefore should choose f to fall between line harmonics*. Workers often choose f to lie 5 Hz "off center" (e.g. 745 Hz instead of 750 Hz). Do not choose f = n x 60 Hz!

Experimentation is advised to find the best values of T and t_o. Start by trying successively larger values, keeping $t_o \sim 4T$, and look for the point at which the measured noise shows no further reduction. Of course it would be much preferable to get rid of the interference at the source (things get much better when the line frequency interference is 20 dB or more below the signal).

CONCLUSION

This article has focused on the Model 3990 System as the tool that best handles D*. measurements. Other ITHACO Lock-In Amplifiers also can be used in conjunction with the 385EO (GPIB) or 386EO (RS-232) Integrator/Coupler ADC devices with good results.

Model	Comments
393 DYNATRAC 5 nV/$\sqrt{Hz}$ max 0.03 pA/$\sqrt{Hz}$ max	Same as 3990 except no LIA sensitivity programmability. 385 will upgrade existing unit. Gain programmability often unnecessary. Cardset optimization of PLL speed.
397EO DYNATRAC 15 nV/$\sqrt{Hz}$ typ	With 385 forms 3970 system. Has sensitivity programming. Should have Option 12 to speed up PLL. Option 06 is optional for gain programmability. Option 10 is mandatory for dual phase dc output. No phase adjustment, so do not use for SNDR > 60 dB. Non-isolated front end supply. Very reliable and simple to operate in production environment. Built in current preamp for photovoltaic or PMT detectors.
391A DYNATRAC 5 NV/$\sqrt{Hz}$ max 0.03 pA/$\sqrt{Hz}$	One 385 can handle two independent 391A's. Fully floating front end supply. Single phase, manually adjusted. Half the speed due to only one channel.Cardset optimization of PLL.
3961 5 nV/$\sqrt{Hz}$ max	With 385 forms fully programmable D* measurement system for automated production testing. Requires low impedance input (use preamp). Operate in BPL5 mode. Should use good quality chopper, Model 220.
3962 5 nV/$\sqrt{Hz}$ max	One 385EO will handle two independent 3962's. Single phase version of 3961 - same comments apply, except measurement speed is halved. Automatic phase setting simplifies operation.
3921 25 nV$\sqrt{Hz}$ typ	Lowest cost. One 385 will handle two 3921's. Often will require preamp for overcoming input noise voltage or current. Half the speed due to single channel. Autophase tracking eliminates manual phase adjustment. Limited ability to handle interference due to modest dynamic reserve and 6 dB/octave output filtering on time constant.

APPENDIX A

D* HOST COMPUTER PROGRAMS

ITHACO has available software which customers may use "as is" or modify to suit their purposes. The programs are based on the HP85 program "NOISE8" as described in reference #5,Appendix C. They are written for use with ITHACO DYNATRAC Models 391A, 393, 397EO (3970) or 399 (3990).

1. "D*HP-D". D* program for HP-85 with GPIB interface and ITHACO DYNATRAC Lock-Ins. Supplied on tape. See ITHACO IAN 44, "HP-85 Software for D*Measurement Using DYNATRAC Lock-In Amplifiers".

2. "D*PC-D". D* program for IBM PC operating in BASICA with National Instruments Model GPIB-PC2 interface board and ITHACO DYNATRAC Lock-Ins. Requires National GPIB-PC Rev. C Binary Handler software. Supplied on 5.25 inch floppy disc ITHACO P/N A13157. See ITHACO IAN 45, "IBM-PC Software for D* Measurement Using DYNATRAC Lock-In Amplifiers".

Shown below is the printout obtained using "D*HP-D" with the D* demonstration setup shown in Figure 5.

The interpretation is as follows:

The unknown IR power P was estimated assuming an effective rms sinusoidal responsivity of 12Ω/nw for the type 5055 detector. This yields about 1.5 x 10^5 v/w for the setup shown. For a 75 μV detector signal we therefore have: $P = V_s/R = 75\ \mu V/1.5 \times 10^5 = 0.5$ nw.

V_s: LIA rti signal; vector sum of V_s (a) and V_s (b) data. Channel B has been nulled to zero in this instance for observation of chopper jitter effects. Ch.B reproducibility is on the same order as the baseline uncertainty due to detector noise (1% of full scale).

e_n: LIA rti noise = $\sqrt{(1/2)(e_n(a)^2 + e(b)^2)}$ Note that e_n (a) ~ e_n (b), therefore we have no evidence of chopper phase noise error here.

SNDR: 20 log (139.7) = 42.9 dB for the combined Ch.A & Ch.B noise and signal data (±1σ reproducibility = ±σD* = ±2%.)

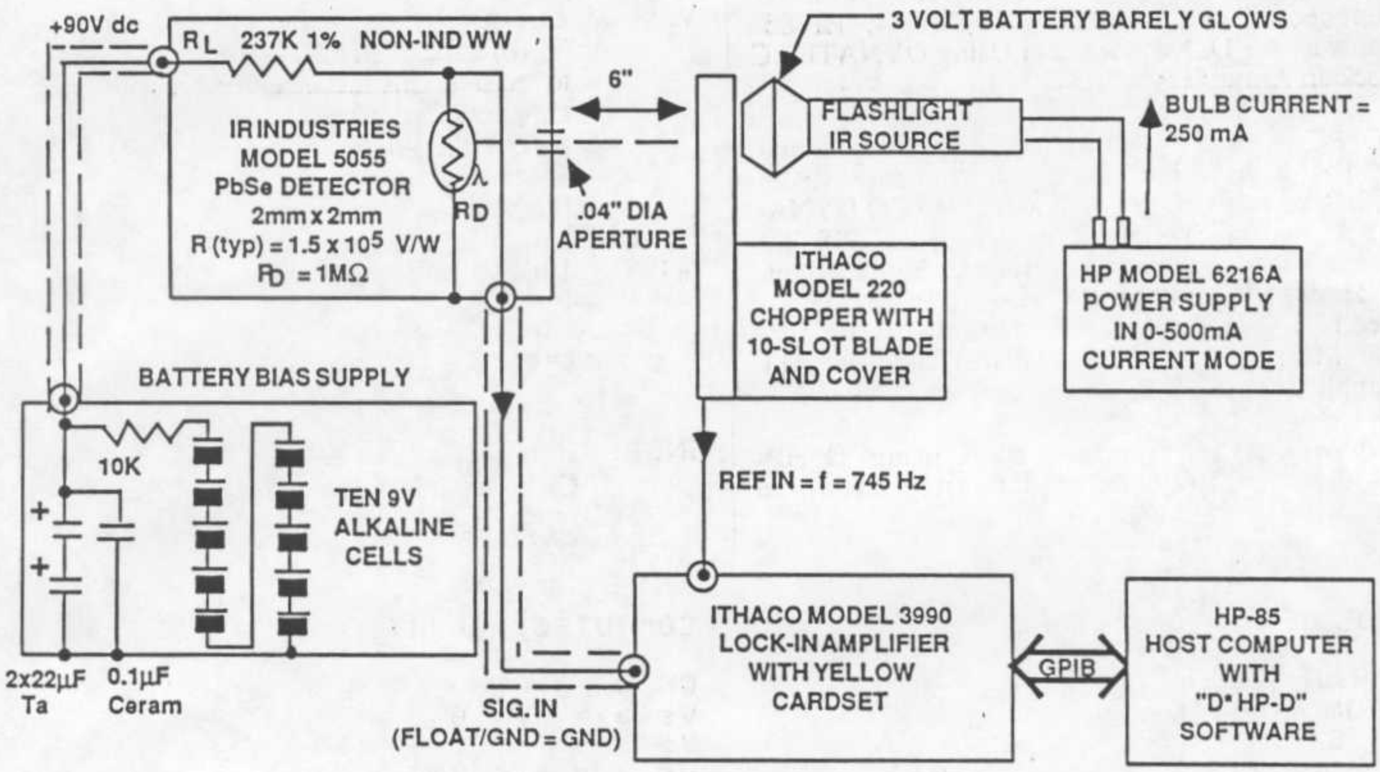
```
MEASUREMENT CONDITIONS

LIA TIME CONS= .00125
LIA NOM ENBW = 100
LIA F.S. MV  = .1
385 I,To(SEC)=     1       .017
385 EQUIV NBW= 30
SYSTEM ΔF, Hz= 25.7812750599
SAMPLES, N   = 600
T(meas) NI/60= 10
SIGNAL BW ΔF'= 4.99882809571E-2
385 PARAMETER CODES:
A1,B1,D03,I0001,MA,OC,R000,S00,T
B,Z0


RAW DATA OUTPUT FROM 385

CH.A OUTPUT,VDC Es(a)= 7.40454
CH.A FLUCT,VRMS En(a)= .2727

CH.B OUTPUT,VDC Es(b)= .11202
CH.B FLUCT,VRMS En(b)= .26547
```

```
COMPUTED RESULTS/REPRODUCIBILITY

CH.A, B, & VECT.SUM SIGNAL(+-σx̄)
Vs(a)=  74.045 μVRMS   (±  .16%)
Vs(b)=   1.120 μVRMS   (±10.44%)
Vs   =  74.054 μVRMS   (±  .16%)
CH.A,B,V.S. NOISE/ROOT Hz (+-σn)
en(a)= 537.073 nVRMS   (± 2.85%)
en(b)= 522.833 nVRMS   (± 2.85%)
en   = 530.001 nVRMS   (± 2.02%)
1Hz S/N RATIO=Vs/en=P/NEP(+-σD*)
SNDR(a)=          137.87 (± 2.85%)
SNDR(b)=            2.14 (±10.82%)
SNDR    =         139.72 (± 2.03%)


DETECTOR FIGURE OF MERIT.  USE:
a,b DATA; SING CHAN LIA READINGS
comp DATA; DUAL CHN LIA READINGS
RADIANT POWER P = .5 NW
DETECTOR AREA A = .04 CM.SQ.
                          ( +-σD* )
NEP(a)     = 3.627E-012 (± 2.85%)
NEP(b)     = 2.334E-010 (±10.82%)
NEP(comp)  = 3.578E-012 (± 2.03%)

 D*(a)     = 5.515E+010 (± 2.85%)
 D*(b)     = 8.570E+008 (±10.82%)
 D*(comp)  = 5.589E+010 (± 2.03%)
```

NEP (comp): We read the composite Ch.A &Ch.B result since we are taking data from both channels (vector sum information). In general NEP(a) and NEP(b) will be meaningless unless the LIA is phased for a maximum output on one channel (as we have done in this example for Ch.A).

D* (comp): Again we observe the composite result. D*(a) and D*(b) refer only to single phase LIA operation.

When this experiment was repeated with the chopper slowed to 90 Hz (9% of full speed), results to 2.22% σ_{D*} accuracy were obtained in 10 seconds. Changes to Figure 5 conditions were: LIA Time Constant = 4 msec, LIA Ref. Range = YELLOW 100Hz - 1kHz. The observed SNDR decreased to 30 dB due to increased detector 1/f noise at this frequency, (e_n = 2.4 $\mu V/\sqrt{Hz}$). No 2f difficulty was encountered (per Application Comment #7). This example demonstrates that the use of superior quality DYNATRAC instrumentation prevents errors as measurement conditions become more difficult.



CONDITIONS:	DETECTOR TEMP	= 25°C
	LIA TIME CONSTANT	= 1.25 msec
	LIA SENSITIVITY	= 1 mV x .1 = 100µV
	LIA REF RANGE	= 1K - 10K
	LIA PHASE	= null A SINø for phase jitter diagnostics (optional)
	385 "I" PARAMETER	= 1 (t_o = 1/60 SEC, r_s = 60 Hz)
	# OF SAMPLES	= 600 t_m = N t_o = 10 seconds)
	DATA COLLECTION	= Both channels of LIA (Vector Sum Mode)
	σ D* REPRODUCIBILITY	= ±6% at 99.7% confidence level

FIGURE 5 D* DEMONSTRATION SETUP

APPENDIX B

Measuring Detector Noise in the 1/F Region Using the Model 3970 Lock-In Amplifier

Typically 1/F noise measurements are made at octave intervals such as 10 Hz, 20 Hz, 40 Hz, 80 Hz, ... In order to achieve maximum measurement speed, the question becomes: how do we choose lock-in time constant (T) and 385 ADC sampling interval (t_o) to achieve the widest practical bandwidth. Two considerations are important. First, the equivalent noise bandwidth of the lock-in taken alone (LNBW) must not be so wide that it includes the origin (e.g., 50 Hz bandwidth at 10 Hz measurement center frequency). Second, the composite system bandwidth, $\Delta f = f(T,t_o)$, must not cover so much of the 1/F noise spectral density curve as to distort the true spot noise at the center frequency (f).

The Model 3970 Lock-In is the preferred instrument for this type of measurement. It covers the range from 10 Hz - 10 kHz, without changing reference ranges or cardsets. Compared to the Model 3990 it has a rather large ac input bandpass filter signal conditioning passband at low frequencies (prevents potential errors in computing Δf). Furthermore, it easily can be modified for switching under host computer control to a longer time constant for the lowest frequency points. Empirical testing with a darkened detector has shown that only two time constants are required; .025 sec (5 Hz LNBW) and .0025 sec (50 Hz LNBW), to cover the range 10 Hz to 10 kHz. The 385 sampling interval t_o is then chosen to achieve the requisite bandwidth, Δf, without significant time measurement penalty. Table 1 shows the optimum T and t_o parameters. For 160 Hz and above, the 3970 can be operated with a "wide open" bandwidth. For each frequency, the number of samples, N, was chosen to yield a noise measurement reproducibility of $\sigma_n = 0.0333$ (±10% uncertainty at the 0.997 3σ confidence level). This allows for meaningful time comparisons.

f Hz	T sec	I	t_o sec	LNBW Hz	CNBW Hz	Δf Hz	N	t_m sec
10	0.025	12	0.2	5	2.5	2.03	218	43.6
20	0.025	6	0.1	5	5	3.20	240	24.0
40	0.025	1	.017	5	30	4.85	955	15.9
80	0.0025	2	.033	50	15	13.2	218	7.3
160	0.0025	1	.017	50	30	22.8	218	3.6

f = passband center frequency
T = lock-in time constant
I = 385 integration time code
t_o = 385 sampling time = I/60
LNBW = lock-in equivalent noise bandwidth
CNBW = 385 equivalent noise bandwidth
Δf = composite measurement bandwidth
N = # of samples for ±10% 3σ reproducibility
t_m = N t_o = measurement time

TABLE 1 OPTIMUM 3970 PARAMETERS FOR LOW FREQUENCY NOISE MEASUREMENT

When chopped IR light was allowed to fall on the detector for the frequencies listed above, essentially the same noise values were obtained. The radiation was fixed at a level which yielded a 17 dB SNDR at 10 Hz increasing to a 33 dB SNDR at 160 Hz (due to 1/f decrease in noise at higher frequencies). This demonstrates that the same values of T and t_o as shown above could be used if one were making simultaneous signal and noise measurements for D* characterization in this region. The frequency values 20, 40, 80, 160, 320, ... are good choices for D* because they lie away from the line frequency harmonic series, 60, 120, 240, ..., yet do not coincide with multiples of 30 Hz.

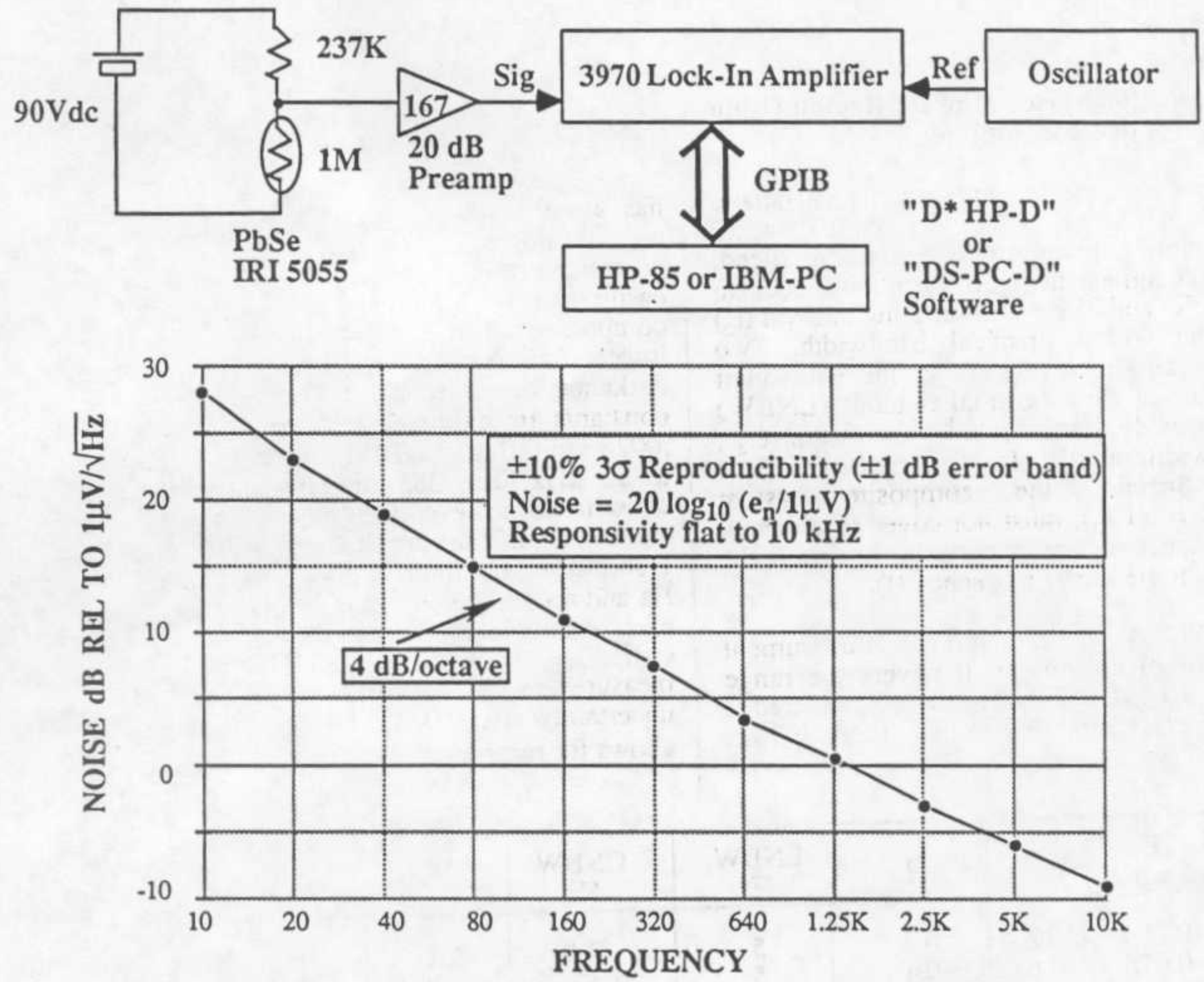


Figure 6 Pb/Se Detector 1/F Dark Noise Measurements Results

Figure 6 plots the 1/F curve for a typical PbSe photoconductive detector. The Model 167 Voltage Preamplifier was connected directly to the detector head. This arrangement prevents interference pickup by the input wiring, and suppresses any microphonic tendency of the lock-in. It also provides 1 microvolt sensitivity with much lower voltage and current noise than possible with the 397EO alone. Furthermore, it reduces capacitance, allowing for flat responsivity to beyond 10 kHz.

To make faster measurements at the lowest frequencies, one could increase Δf by lowering T and t_o, then compensate for the error introduced by having too wide a bandwidth relative to the 1/F curve. The compensation could be determined either analytically or emprically.

REFERENCES

1. R.D. Hudson, *Infrared System Engineering*, John Wiley & Sons, New York (1969).

   Chapters 7,8,9 contain a clear and readable account of detectors, noise and methods for measuring figures of merit. Be cautioned however that the definitions for NEP, NEI and D <u>are not</u> normalized to 1 Hz (see Table 9.1, pg. 322), but that D* <u>is</u> normalized to 1 Hz. This unfortunately confusing failure to carry through the √Hz factor for NEP, NEI and D occurs commonly in IR literature, leading to a variety of conventions for denoting (or not denoting) bandwidth, for example NEPλ (3.2 μm, 800, 10) for a determination with f = 800 Hz, Δf = 10 Hz, leaving you to wonder whether the reported value was or was not calculated with a √Hz factor.

2. Kruse, McGlaughlin & McQuistan, *Elements of Infrared Technology*, John Wiley & Sons, New York (1962).

   Chapter 8, Section 8.2, consistently normalizes the measurement bandwidth for NEP, NEI, D and D*.

3. Hotter, et al, *Fundamentals of Infrared Technology*, Macmillan, New York (1962).

   See Chapter 12, Test Procedures. NEP, NEI, D definitions omit bandwidth normalization.

4. Wolfe & Zissis, *The Infrared Handbook*, Office of Naval Research, Washington, D.C. (1978).

   See section 11.4, Detector Characterization. Bandwidth consistently normalized for all parameters.

5. J. L. Scott, ITHACO Application Note 36, *Digital Techniques for Random Noise Measurement Using Lock-In Amplifiers*, ITHACO Inc., Ithaca, New York (1985).

6. Scott, Pease & Fisher, *Digital Lock-In Techniques for IR Detector and Fiberoptic Testing*, <u>Laser Focus/Electro-Optics</u>, Sept. 1985 (Reprint available from ITHACO).

7. ITHACO Application Note 50, *Noise Analysis and Gain Considerations in Selecting the Right Current Preamplifier*, ITHACO, INC., Ithaca, New York (1986)