DIGITAL LOCK-IN TECHNIQUES FOR IR DETECTOR AND FIBEROPTIC TESTING

Digital computation of the spot noise accompanying the signal input of a lock-in amplifier produces results far superior to traditional analog methods. It provides much faster and more accurate measurements, works over an extremely wide S/N ratio range, and covers the full frequency spectrum of the instrument. Examples of photodetector testing and fiberoptic characterization are given.

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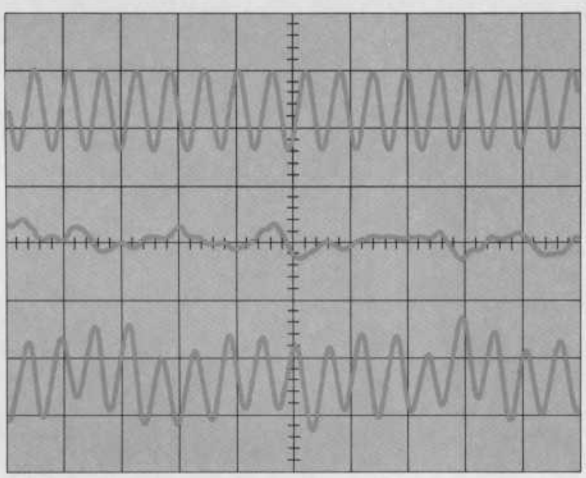

Lock-in amplifiers will measure a signal with high accuracy even when that signal is buried in discrete frequency interference and random noise. They have been used for years in laboratories to measure extremely small signals. More recently, the LIA has found use in the production testing of optical fibers and solid-state detectors. In these applications, the LIA measures both the signal and the random noise to establish signal detection limits, determine the maximum possible measurement speed, and calculate the S/N ratio. In addition, solid-state detector manufacturers use the LIA to measure detector spot noise at one or more frequencies.

The noise measurement capability of LIAs can be greatly improved using A/D converters in conjunction with microprocessors programmed with noise calculation firmware, making noise

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measurements with the LIA very practical to use in test fixtures. This article explains how these digital techniques are implemented using an integrating A/D converter/computer interface unit to make better noise measurements. It also discusses some of the theory of noise measurement.

The LIA measures ac signals obscured by noise. It acts as a very selective ac voltmeter, in effect acting as a narrowband filter capable of extremely high Q to reject both random noise and discrete frequency interference (DFI). It produces a dc output proportional to the rms voltage of the fundamental of an input signal (e_s) that is coherent with a second input, the reference frequency (f_r). It can be thought of as a bandpass filter that automatically locks its center frequency to f_r, followed by an rms-to-dc converter.

Noise measurement using LIAs

The LIA can be used to measure the spot noise surrounding f_r, that is, the Gaussian noise density e_n in volts rms/$\sqrt{\text{Hz}}$. This can be done either independently of a signal measurement or in conjunction with one. f_r can be scanned over a range of frequencies to obtain a noise density profile if desired.

The LIA can do this because it must inevitably allow the portion of the input random noise lying within its effective passband to appear in its output. This causes fluctuations in its dc output level, the amplitude of which can be measured to yield the referred-to-output (rto) noise, (S). This can be readily scaled to a referred-to-input (*rti*) value using the known equivalent noise bandwidth of the measurement system (B) and the ac-input to dc-output gain (G) of the LIA:

$$e_n = \frac{S}{G\sqrt{B}} \text{ V rms}/\sqrt{\text{Hz}} \qquad (1)$$

Usually the LIA employs a two-pole (12 dB/octave) lowpass output filter with RC time constant τ to supply the overall bandlimiting, in which case $B = 1/(8\tau)$, as explained in the figures accompanying the sidebar "Basic Principles in Noise Measurements."

Analog and digital noise measurement

Typically, one measures noise for one of two reasons. First, a knowledge of the S/N ratio (S/N ratio = e_s/e_n) allows predictions on how long it will take to measure a signal to a given confidence level.

Second, the amount of noise present defines the limits of resolution of the measurement system, beyond which the signal is unrecoverable (at least in the time allotted). These considerations represent two manifestations of the general problem of speed vs. accuracy tradeoffs that must be faced in any noisy system.

Historically, analog techniques have been used to determine the amount of lock-in output fluctuation and thus measure noise (see Fig. 1). Typically the noise measurement circuit rectifies the ac component of one channel of the lock-in output and lowpass filters it to give an average value. When this is multiplied by $\sqrt{\pi/2}$, the output-referred noise, S, is obtained.[7]

A major drawback of the analog method is slow measurement time due to the time constants of the coupling and filtering circuits, and to the restriction to narrow bandwidths (long time constants) relative to the reference frequency. Further limitations include: restricted range of operating frequencies, small dynamic range for S/N ratio, low accuracy, cumbersome control setup, and difficulty in reading a wavering analog meter.

Digital techniques can overcome all of these problems if correctly implemented. Some commercially available LIAs have simply used digital techniques to emulate the analog circuitry, making measurements more convenient but leaving the problems of speed, accuracy, and flexibili-

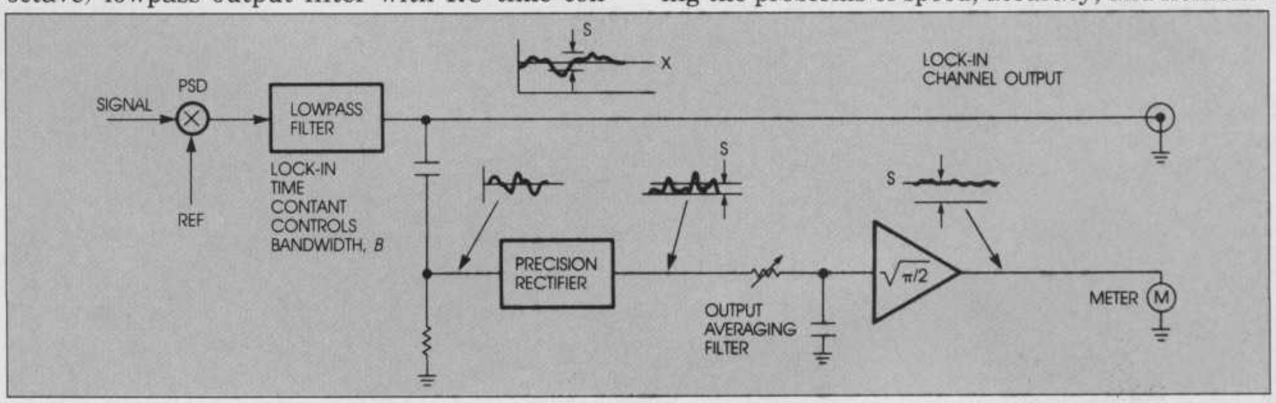


FIGURE 1. Analog noise measurement technique.

BASIC PRINCIPLES IN NOISE MEASUREMENTS

Basic principles involved in noise measurements, using a highly simplified LIA block diagrams is illustrated in Fig.1. Both frequency domain spectra and time domain waveforms are shown. The heart of the LIA is the phase-sensitive detector (PSD), which synchronously rectifies the input spectrum, downshifting it so that the frequency of interest ends up as a dc output and all other components appear as ac fluctuations. This allows a humble RC lowpass filter to act as the narrowbanding device. τ can be set very long to impose bandwidths as low as $B = 0.001$ *Hz.* (See Ref. 6).

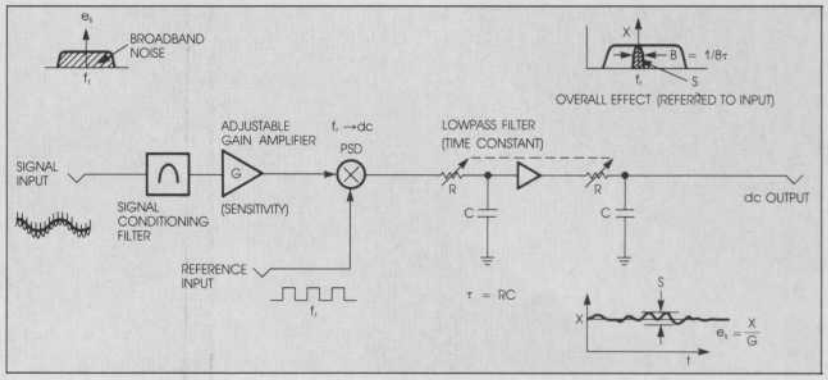


FIGURE 1. Elementary LIA signals and noise.

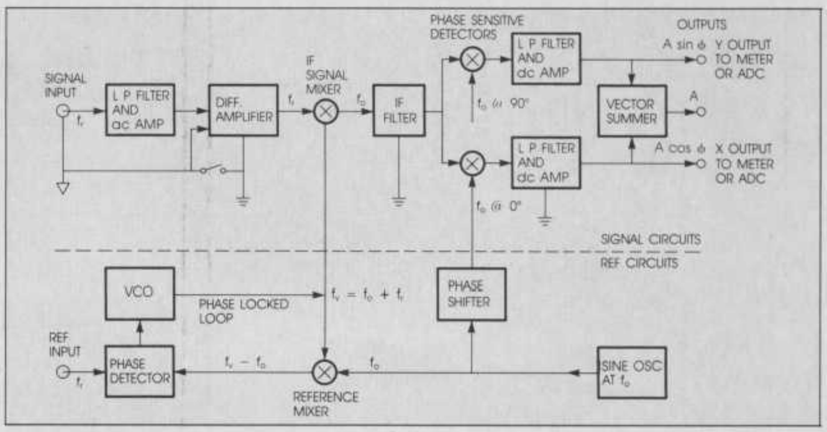


FIGURE 2. A modern dual-channel heterodyning LIA.

The elementary LIA design suffers from a number of performance limitations, the most important of which are susceptibility to error due to harmonics of the reference present in the input, poor overload characteristics, and sensitivity to the phase of the signal relative to the reference. Over the years a number of elaborations have been added to overcome these difficulties, including front end high Q filters, linearly multiplying phase-sensitive detectors, signal heterodyning to an intermediate frequency and dual-phase sensitive detectors. ITHACO LIAs employ the heterodyning, dual-phase concept (see Fig. 2) to provide the convenience of phase insensitive measurements and the ability to track a changing reference frequency without the compromises of inadequate dynamic reserve, poor accuracy, residual harmonic responses and increased self-noise encountered with the other modifications.

Since the two phase-sensitive detectors are operated in quadrature, they resolve the signal into its *X* and *Y* components. These components can be combined either by analog means (vector summer as shown) or by calculation from the digitized channel outputs (see Fig. 3).

The dual-phase LIA also can double noise measurement speed. This is because the noise in the two channels is uncorrelated and unrelated to the channel signal level, allowing results from the two channels to be combined.

Heterodyning LIAs employ relatively low Q filtering in the ac signal conditioning circuitry. This results in a relatively wideband front end, compared to the bandwidth restriction of the post-detector dc filtering, and thus a negligible effect on noise measurement accuracy. LIAs that use high Q bandpass filters instead of heterodyning to deal with harmonic rejection will have noise measurement errors due both to computational errors in trying to compensate for the variable effect of the front end passband on noise bandwidth and to misalignment of its center frequency with the reference. Furthermore, narrowband filters generate noise, raising the noise floor of the system and masking weak noise measurements.

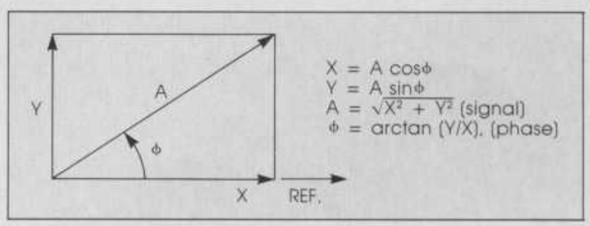


FIGURE 3. Relationship between dual-channel lock-in outputs, signal, and phase.

ty unsolved. On the other hand, the correct digital technique can overcome all of the shortcomings.

The proper digital method begins with the observation that the standard deviation of the LIA output fluctuation is mathematically identical to its rms noise. Thus we may sample the LIA output to obtain an estimate of the rto noise (S) in volts rms. (We shall use the convention of underlining variable symbols to distinguish actual from measured quantities.)

$$\underline{S} = \sqrt{\frac{1}{N}\sum_{j=1}^{N}[\underline{X} - x_j]^2} \quad \text{(rto noise)} \qquad (2)$$

$$\underline{X} = \frac{1}{N}\sum_{j=1}^{N} x_j \quad \text{(rto mean signal)} \qquad (3)$$

The reproducibility of our estimates for $\underline{S}$ and $\underline{X}$ depends on the number of samples taken (N). By taking a sufficiently large number of samples the accuracy of our estimate can be made as high as we wish. For noise, the error band that surrounds a measurement can be readily calculated as follows (see ref. 8), assuming the samples are spaced far enough apart in time relative to the bandwidth to prevent autocorrelation (statistical oversampling):

$$\sigma_n = \sqrt{\frac{1}{2N}}, \; N \geqq 30 \qquad (4)$$

We define σ as the fractional standard deviation that we would observe in a series of determinations of S (or e_n), each of length N. If we chose $N = 450$ for example, we can be 68% confident that a given measurement lies within $\pm 3.3\%$ ($\pm\sigma_n$) of the actual noise, and 99.7% confident that it lies within $\pm 10\%$ ($\pm 3\sigma_n$). Figure 2 is derived from Eq. 4 to assist in choosing N to yield acceptable accuracy.

Similarly the signal reproducibility σ_x for the average of N samples (for the estimated signal, X or e_s), assuming no oversampling, is given by:

$$\sigma_n = \frac{\sqrt{B/N}}{e_s/e_n} = \frac{\sqrt{B/N}}{\text{S/N ratio}} \qquad (5)$$

Thus, once we measure e_s and e_n, we may predict how much we can trust a given signal measurement, whether it be a single sample or the average of N samples, under the actual measurement conditions in use.

Noise measurement speed

It has been determined empirically that oversampling will not occur if the sampling rate r_s,

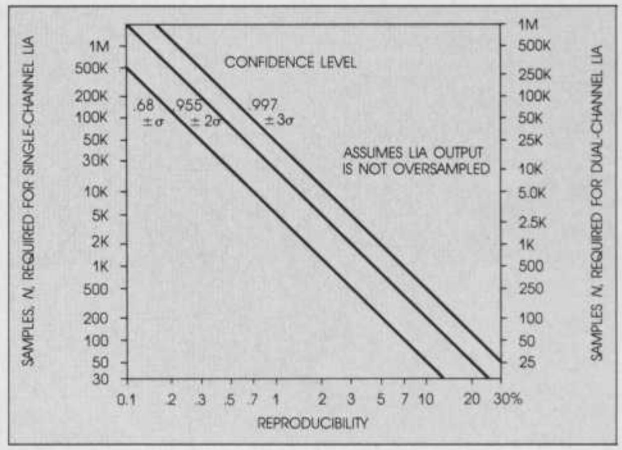


FIGURE 2. Number of samples required to achieve a desired degree of noise measurement reproducibility at several confidence levels.

satisfies the inequality:

$$r_s \leqq 2\sqrt{2}\,B \qquad (6)$$

Sampling faster than the $2\sqrt{2}\,B$ rate will lead to greater reproducibility errors than predicted by Eq. 4 or 5 or Fig. 2, in proportion to the degree to which the limit is exceeded. Usually there is no harm in oversampling, provided that the error predictions are compensated. In some special circumstances oversampling is desirable in order to provide somewhat faster noise measurements.

Eq. 6 applies best to the situation wherein a fast (wideband) A/D converter samples the LIA output. Actually, the ITHACO integrating A/D converter (see sidebar "Sampling the LIA Output") averages each sample over an interval t_o, with samples being taken back-to-back. This process has an associated equivalent noise bandwidth, $1/(2\,t_o)$. As it turns out, under most real world measurement circumstances the best results are obtained by allowing this bandwidth to be narrower than the lock-in bandwidth, $1/(8\,\tau)$.

If this is done (by making t_o at least five times larger than τ), oversampling cannot occur.[1] With the integrator thus dominant, the measurement time, t_m, simply equals Nt_o and our reproducibility predictions will always be valid. Figure 3 allows one to determine measurement time, given the number of samples and the A/D converter sampling time or noise bandwidth.

Making digital measurements

Several applications observations can be made concerning the use of the dual-channel integrator/coupler as interface between analog LIA and host computer.

In cases where neither the LIA nor the coupler dominates, the system equivalent noise band-

SAMPLING THE LIA OUTPUT USING A V/F ADC

The ITHACO Model 385EO Integrator/Coupler uses a voltage-to-frequency integrator to sample the lock-in output. This microprocessor-based computer peripheral device performs A/D conversion, makes baseline corrections and calculates vector sum, phase angle, sum, difference, and ratio information from the raw data. It communicates with the host via the GPIB (or an RS-232 port, Model 386EO). Its firmware also will accumulate noise ($\underline{S}$) and averaged signal ($\underline{X}$) data for both channels for a specified number of samples (up to 999,999) when triggered by the host. For upgrading the capability of existing equipment, the 385EO/386EO Integrator/Coupler can be teamed with practically any make or model of LIA with excellent results.

The use of an integrating sampling technique involves no performance compromises, and confers some unique advantages for noise measurement over the conventional sample-and-hold/fast A/D conversion method employed by other manufacturers. Since the length of the sampling period, t_o, is programmable, it may be increased to improve the resolution of the digitization process when necessary. The resolution equals $t_0/10^5$. For example, when trying to measure a very small amount of noise superimposed on a relatively large signal (for example, S/N ratio = 80

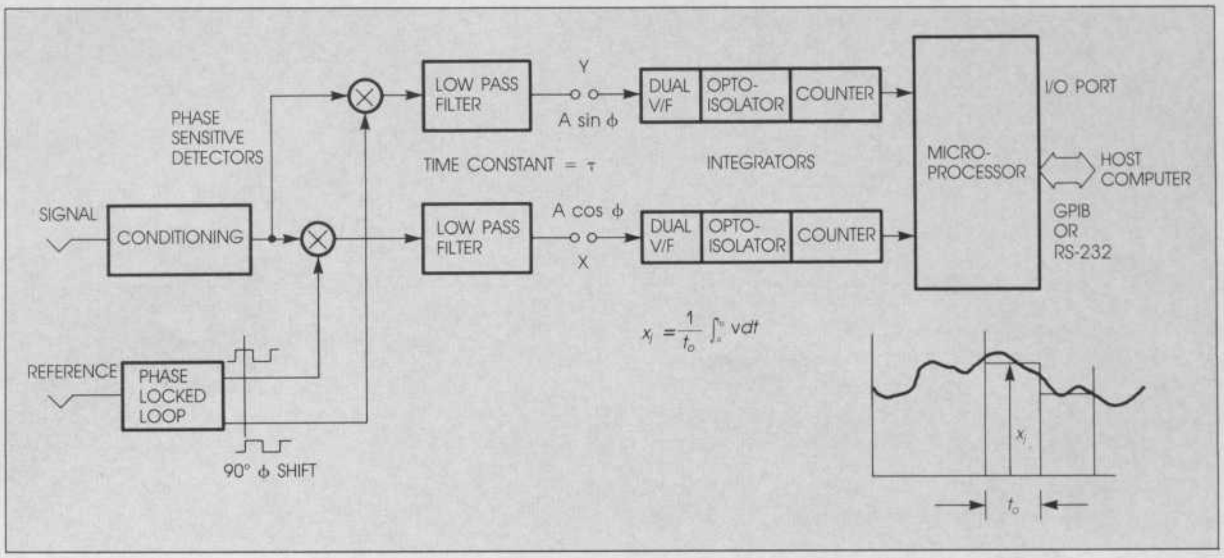
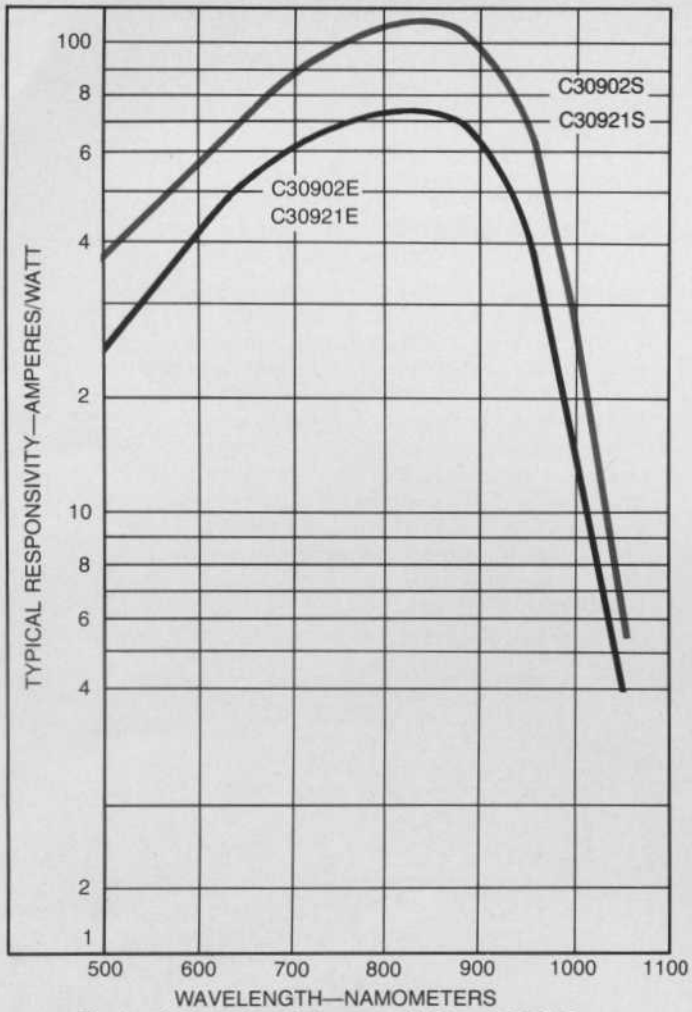
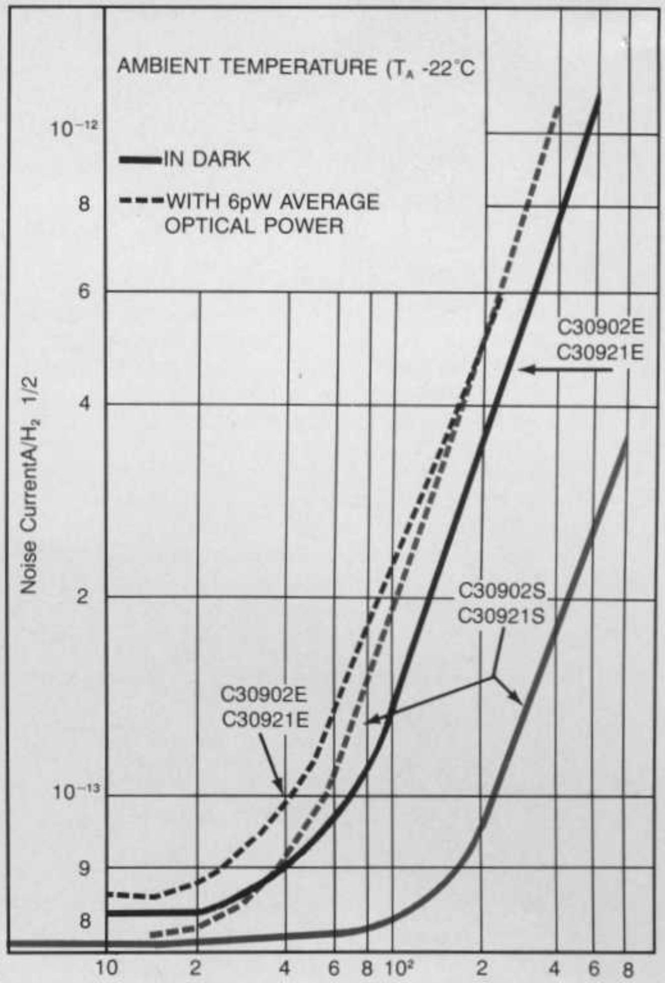


FIGURE 1. Lock-in amplifier sampled by integrating A/D converter.



Typical Spectral Responsivity at 22° C



—Typical Noise Current vs Gain

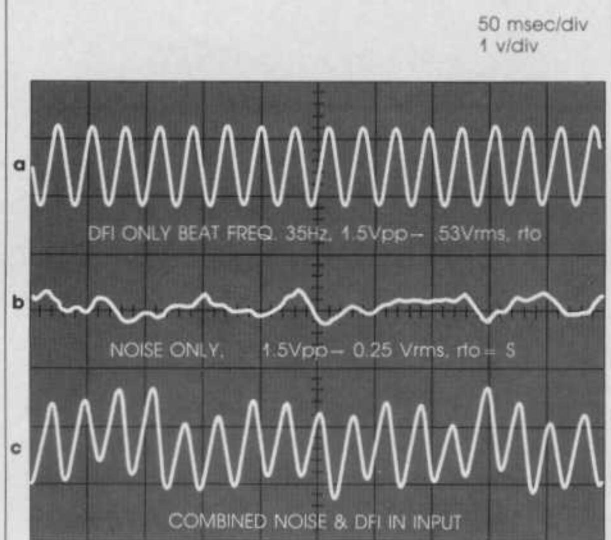


FIGURE 2. LIA output waveforms with DFI only, noise only, and both present. The rto interference due to DFI measured for **a** was 25 $nV/\sqrt{Hz}$, which when vectorially added to the actual e_n measured for **b** (89.3 $nV/\sqrt{Hz}$) leads to an estimated error of +3.8% (92.6 $nV/\sqrt{Hz}$). The value actually measured for **c** was 92.3 Hz. 14400 samples were taken over 8 minutes for each measurement to achieve a reproducibility of $\sigma = \pm 0.5\%$. The averaging of the interference is for a typical non-ideal condition, $t_o = 1.25\ (1/\Delta f)$ to span 1¼ cycles of the beat frequency. The LIA being used (ITHACO Model 399) would require a 0.125 second (1Hz NBW) time constant to obtain this level of DFI suppression if a wideband ADC had been used instead of an integrating type. This would require a measurement time 7.65 times longer to achieve the same noise measurement reproducibility, due to the bandwidth restrictions on sampling rate.

Conditions: LIA time constant τ = 12.5 ms (10 Hz) NBW
385 Integ. time, t_o = 33.3 ms (15 Hz) NBW
DFI = 6.2 μV rms @ 600 Hz
LIA reference = 635 Hz
Noise, e_S = 89.3 $nV/\sqrt{Hz}$
LIA gain, G = 10 Vac/ 10 μV rms = 10^6
LIA + Integrator Eq. NBW, B = 7.65 Hz

dB) we can push the resolution above 15 bits to overcome the masking effect of quantization noise and allow accurate measurements.

The ability to select the length of time for sampling yields a second advantage in that it provides an additional and adjustable amount of signal filtering beyond what the LIA provides. When discrete frequency interference (DFI) is present in the input, such as harmonics of 60 Hz in low level work, this can lead to a dramatic increase in noise measurement speed. DFI close enough to the reference frequency to produce a difference frequency within the passband of the LIA can mask the noise. This necessitates increasing the time constant to reduce the lock-in bandwidth and exclude the interference, with consequent slow measurement times. But the use of a sampling period set long enough to span several cycles the beat frequency will average out the DFI fluctuations, allowing up to an order of magnitude reduction in τ for the same accuracy, with a proportional increase in speed. The beat frequency, as seen on an oscilloscope, (see sidebar Fig. 2) can be larger than the noise amplitude without ill effect.

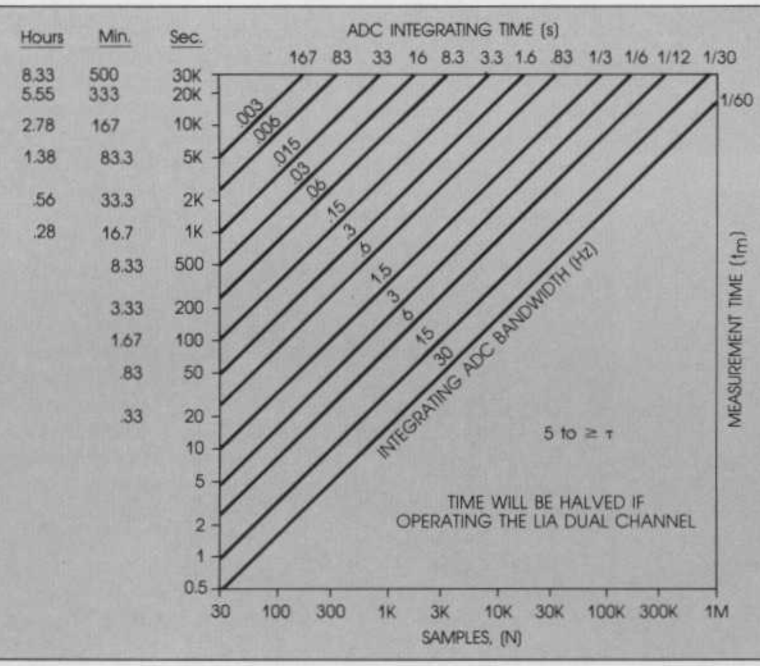


FIGURE 3. Measurement time assuming integrator dominates system noise bandwidth.

width B becomes a somewhat complicated function of t_o and τ. The host computer can readily calculate its value. (See Ref. 1 and 2).

When using the coupler to simultaneously measure signal and noise, the effective equivalent noise bandwidth for the average signal, $\underline{X}$, usually turns out to be very nearly equal to $1/(2\,Nt_o)$. Thus (from Eq. 5) the signal reproducibility is:

$$\sigma_x = \frac{\sqrt{1/(2Nt_o)}}{e_s/e_n} \qquad (7)$$

When using a dual channel LIA, the host may compute the vector sum signal from the quadrature components. The noise may be obtained from *either* quadrature channel, independent of phasing. (Note that obtaining noise by sampling the vector sum is mathematically invalid). To double the measurement speed, the host can combine the channel X and Y noise data using:

$$e_n = \sqrt{[e_n^2\,(x) + e_n^2\,(y)]/2} \qquad (8)$$

During a longer noise measurement, signal level drift, such as that caused by changing illumination levels, can cause severe errors if the noise is small. To combat this, the measurement should be subdivided into M short segments of N samples a piece. The host then should perform an rms combination of the multiple measurement runs to obtain noise to the desired level of accuracy.

To get the fastest spot noise determination, the system bandwidth should be set as wide as possible, consistent with DFI, reference frequency, and the spectral noise density profile of the input. If the LIA can be operated at its fastest time constant (for example, 1.25 ms), and the coupler at its shortest sampling time (1/60 s), ±10% reproducibility @ ±3 σ (99.7% confidence level) can be achieved in less than four seconds, for example, as can be determined from Fig. 2 and 3.

Noise measurement of IR detectors

Manufacturers of photoconductive infrared detectors must screen their product for acceptable noise performance when operating under illumination. Typically the miniscule amount of noise present compared to the signal output will make it difficult to get acceptable accuracy while achieving an adequate test throughput. The use of digital noise measurement will solve this potentially major bottleneck.

A representative test configuration is shown in Fig. 4. The blackbody temperature is set to radiate IR energy in the desired region of the spectrum. The adjustable aperture controls the strength of the beam to suit the detector under test. The illumination should not be so high as to make noise measurements impractical. A rotating chopper blade modulates the beam at the frequency of interest and also provides a synchronous reference signal to the LIA. This frequency is chosen to be as distant as possible from potential or known interfering frequencies. For example, 750 Hz lies midway between the 12th and 13th harmonics of the power line frequency (Δf = 30 Hz).

Let us assume the sensitivity of the detectors

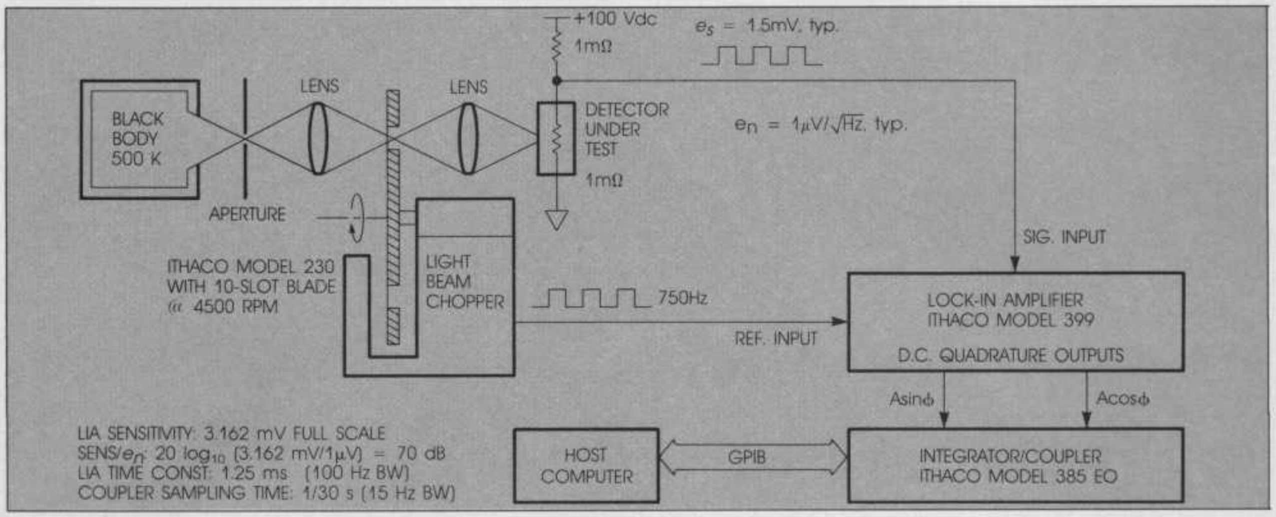


FIGURE 4. Simplified detector noise measurement setup.

will yield an output of between 1 and 3 mV and that a noise density of $1\mu V/\sqrt{Hz}$ must be measurable to $\pm 10\%$ at 99.7 (3 σ) confidence level. We further assume that the use of a high-quality chopper introduces less than 100 $nV/\sqrt{Hz}$ noise due to phase jitter arising from motor speed instability (that is, it will cause less than $+0.5\%$ error when vectorially added to $1\mu V/\sqrt{Hz}$). Let us keep the quantization noise small enough to cause no more than a $+3.5\%$ error when vectorially added to the nominal noise level. As is shown in reference 1, choosing $t_o = {}^1/_{30}$ second will yield a small enough A/D converter step size relative to the output fluctuations of the measurement system to accomplish this. If we scale our noise measurement down 2%, we can safely predict a ±2% contribution to our error budget due to quantization and phase jitter effects, leaving ±8% for sampling effects. From Fig. 2, we discover that we must take 400 samples operating dual channel to get ±8% at 3 σ. Since t_o = 1/30 second, Fig. 3 tells us that the measurement will require 12 seconds to complete.

If we were to reduce the aperture size, we could reduce the S/N ratio, allowing a much faster measurement (less than four seconds). This works because the problem is the reverse of the usual problems faced by LIAs: here we are trying to find the noise buried in the signal! Increasing the illumination would make matters worse, slowing acquisition time and forcing phased operation on a single lock-in channel to get around chopper phase noise interference.

Measuring fiberoptic attenuation

Here we find the opposite extreme, trying to measure a faint signal all but obliterated by detector noise. The problem is to measure the detector noise in order to predict how long it will take to obtain the desired measurement accuracy, given the signal output level of the detector. The light losses in the monochromator, in coupling the beam into the tiny core of single-mode fiber, in the kilometers-long spool of fiber, and in the output coupling to the detector represent a very large signal loss. On top of this, the wavelength being transmitted is not necessarily in the "sweet spot" of the fiber transmission window.

Measurement of noise also is useful for knowing the effectiveness of baseline correction, to determine if the LIA signal input or gain change transients have settled (output transients will be detected as excessive noise), and to discover degraded detector noise performance at end of life.[10]

Using the current-input preamplifier shown, the combined input-referred Johnson noise due to Z_s and R_f is: $\sqrt{4k\tau(1/R_f + 1/Z_s)} = 0.182\ pA/\sqrt{Hz}$, which will cause great measurement uncertainty due to output fluctuations when the signal drops into the sub-picoampere region. By letting the coupler integrate over a long enough period, we can obtain the degree of signal smoothing necessary to bring this error within bounds. In practice, ITHACO's Model 385 subdivides this measurement period into N intervals in the course of a simultaneous signal and noise measurement per Eq. 2 and Eq. 3. In this case, the signal

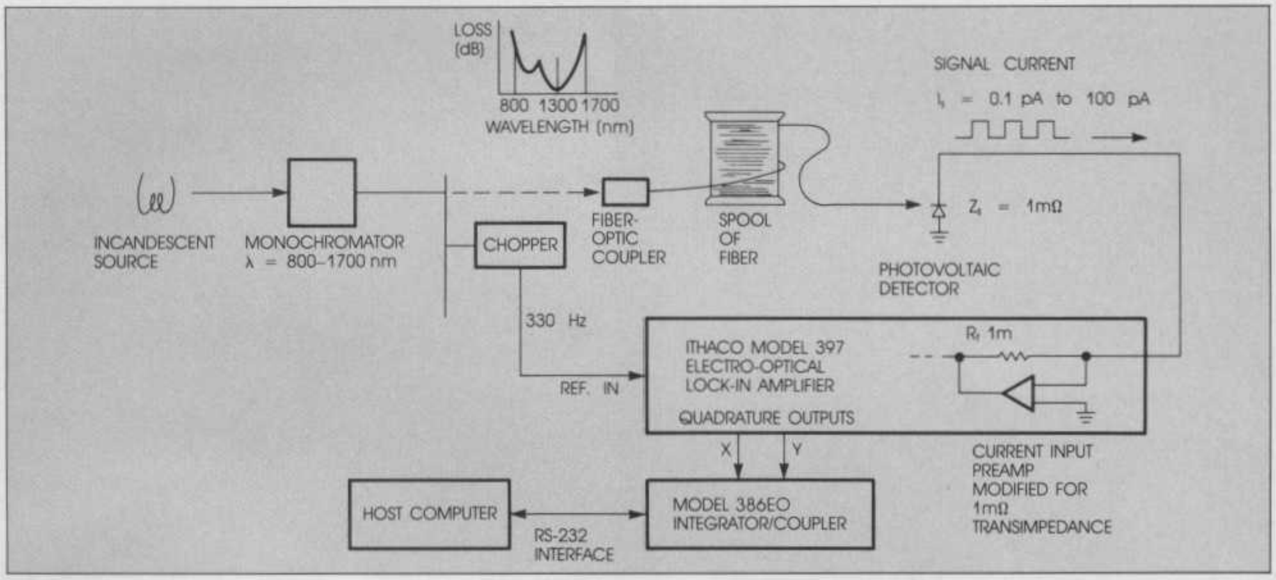


FIGURE 5. Simplified setup for measuring loss in a single-mode optical fiber. Attenuation is measured by taking the ratio of the signal passing through the entire spool to that measured when the leading end is broken off and connected directly to the detector.

reproducibility is given by Eq. 6, which can be rewritten as follows to obtain the signal measurement time:

$$t_m = Nt_o = \frac{1}{2(e_s/e_n)^2(\sigma_x)^2}, \text{ seconds} \qquad (9)$$

Given the desired signal reproducibility σ_x and the measured values of e_s and e_n we can know exactly how long it will take to get our answer. For example, suppose the signal rms amplitude falls to a level equal to the input noise density, $0.182\, pA/\sqrt{Hz}$, and we want $\sigma_x = 0.05$ for $\pm 15\%$ accuracy at $\pm 3\sigma$. Then $t = 1/2\ (0.05)^2 = 200$ seconds. The only way to improve on this is to increase the signal, reduce the noise, or accept less accuracy in the signal measurement. A tenfold increase in signal would drop the measurement time to 2 seconds, for example.

For this type of testing, the noise remains constant from test to test and spool to spool but the signal varies widely. This suggests that an adaptive routine could be programmed into the host for go/no-go test limit, wherein additional runs can be averaged in with the initial run, if necessary, effectively lengthening t_m until σ_x is reduced to the point that you can be sure that e_s is in or out of bounds. This would save time in verification of very good product, which would require only one run of N samples. It reserves extra time automatically only for cases of product near the edge of acceptability, which might require multiple runs each of length N to get the requisite confidence for a pass/fail decision. The payoff is in faster throughput.

Flexible technique

The use of standard deviation computations in conjunction with a heterodyning, dual phase lock-in, and an integrating A/D converter yields noise measurement performance that far outstrips previously available analog or digital methods. The unique features allow formerly impossible measurements to be made with confidence and ease. More common measurements can be made much faster and more accurately. The flexibility of the technique lends itself to a wide range of measurement problems over the full range of frequencies, signals, and noise levels found in LIA applications. The user tailors the system to his needs using easily written host computer programs.

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