Phase Sensitive Detectors (also known as lock-in detector and phase coherent detectors)

Consider the following signals:

Suppose $v_b$ is a signal we wish to detect and $v_a$ is the noise which comes with $v_b$, i.e.,

$$v_{total} = v_a + v_b$$

The information we want is the amplitude of $v_b$ which is slowly time varying.

How can we detect the signal in the presence of the noise? The noise has large Fourier components so near the frequency of the signal that they cannot be filtered out by conventional filters. The signal is lost if $v_{total}$ is applied to our peak detector of Lecture 4.
This loss occurs because the average of $v_b = 0$ and the large noise signal determines when the diode will conduct. Sometimes when the diode is conducting $v_b$ is positive and charge is added to C due to $v_b$. Sometimes when the diode is conducting $v_b$ is negative and charge is taken from C due to $v_b$. The average voltage developed across C due to $v_b$ is zero. Clearly what is needed is an "intelligent" switch instead of the diode. This switch would close for $v_b$ positive and open for $v_b$ negative. Even better would be a "switch" which also inverted $v_b$ when it was negative and applied it to C; i.e., a full wave rectifier with intelligence enough to know when $v_b$ was going to be positive or negative. With such a switch it would be $v_e$, the noise, which averaged to zero. This desirable control of the switch can be obtained if we know exactly the frequency of $v_b$. An external generator exactly at the frequency of $v_b$ can be used to drive the switch. Notice that the frequencies must be exactly the same, right down to frequency drifts, or the signal and switch get out of phase. In practice a common oscillator provides both the signal frequency and the switch drive. An example makes the process obvious.

A very simple system to study the optical absorption of a sample might look like this in block form

![Block Diagram of Optical Absorption System]

The motor drive monochromator samples the light passed by the sample. The slow variations of light intensity with time are recorded on the strip chart recorder. There are many reasons why the signal might be buried in the noise. The sample may pass only a little light. If the experiment is in the infrared region of the spectrum, the source will be weak and the detectors not too efficient. Besides these problems, there is the fundamental one that we are looking at slowly varying signals (i.e., low frequency signals) for which the $1/f$ noise problem is the worst. We can greatly improve our system as follows:

![Improved Block Diagram of Optical Absorption System]

We have arranged for our signal to be at frequency $\omega_m$ which is chosen to be high to avoid $1/f$ noise of the detector and amplifier. Further, our "intelligent" switch knows just when to switch to follow the signal. The detection is "phase sensitive" in that the switch is in phase with the signal. The filter (time
constant) determines the system band width which is often in the 1 to 0.1 Hz range, but may be 0.01 Hz or even less. The Princeton Applied Physics Notes which follow provide more insight into the applications of Phase Sensitive Detectors.

**Signal Averaging**

If we were actually to build our optical spectrometer with phase sensitive detector as just described, we would be disappointed with its performance with the narrow (and thus slower) filter settings. We would find slow drifts on the recorder which were non-reproducible. In this particular system the major cause of drift would be slow variations in the intensity of the light source. The phenomenon of drifts associated with narrow filter settings is common to most systems. This drift problem can be beaten by the nice, but expensive, technique of "signal averaging." Instead of putting the output of the phase sensitive detector directly onto the recorder, we interpose

The time constant (filter bandwidth) is set for a small value so that a scan through the spectrum can be done rapidly. Many scans are done and during each one the signal is digitized and the digital number added to the contents of a counter. Each counter corresponds to a particular portion of the scan. In each counter the signal adds each time; but the noise, being random, does not grow as fast. The noise addition is a random walk problem and the RMS voltage grows as the square root of the number of scans. The signal grows proportional to the number of scans so the signal-to-noise ratio increases as the square root of the number of scans.

**Mixing and Super Heterodyning**

We have had two examples of shoving signal from one part of the frequency domain to another. In both of those cases we moved a "dc" signal to some higher frequency to avoid $1/f$ noise. A more general form of frequency manipulation involves combining two frequencies to give a third frequency. The general term for this activity is "mixing." If the two frequencies are equal and the resultant third frequency is "dc," the name "phase coherent detection" is also used.
An example of mixing where none of the frequencies involved is near dc is provided by the superheterodyne receiver.

Block diagram of a superheterodyne receiver.

Two requirements of a radio receiver are that it be tunable and that it receive a narrow band so as to avoid interference from adjacent stations. A brute force approach would be to use narrow band tunable filters. Such an approach would be very difficult and expensive. A more reasonable approach is to mix the received signal with a signal from a variable frequency oscillator, the local oscillator. The mixer is a nonlinear element. The quadratic term is the mixer's response yield product term. Using the same trig identities as used earlier for modulation, we obtain sum and difference frequencies. The amplifier following the mixer is narrow band at a fixed frequency, the so-called intermediate frequency or I.F. To be received, the input or R.F. frequency must satisfy either

\[ \omega_{R.F.} + \omega_{local\ oscillator} = \omega_{I.F.} \]

or

\[ \omega_{R.F.} - \omega_{local\ oscillator} = \omega_{I.F.} \]

The local oscillator frequency is chosen so that the two \( \omega_{R.F.} \) choices are far apart and can be separated by the selectivity of the first stage of the receiver.

Phase Shifter

We often have need for a circuit to shift the phase of a signal. For example, in our phase-sensitive-detector augmented optical spectrometer, the phase of the switch and the phase of the optical signal are by no means automatically the same. If nothing else, the phase of the optical signal depends on how the chopper is mounted on the motor shaft. The following circuit is a useful one for phase shifting. Note that the phase shift is a function of frequency. This dependence, unfortunately, always occurs. A frequency independent phase shifter would be most useful. The closest we come to this
ideal is to use a phase locked loop to shift all the harmonics in a square wave together. Of course, in this case we really do not shift the input, but generate a new phase related square wave.

\[ V_{in} \]

\[ V_{o} = -V_{in} \]

The circuit analysis goes easily with a trick. We pretend there are two sources of \( V_{in} \), one feeding the top \( R_1 - R_1 \) line, and one feeding the bottom \( R_2 - C \) line. then we use superposition to calculate separately the two contributions to \( V_o \) from \( V_{in} \).

Suppressing the bottom source leaves the noninverting input connected to ground through an unimportant \( R_2 \). We have an inverting amplifier of gain 1.

If we suppress the upper source we have a noninverting amplifier of gain \( R_1 + R_1/R_1 = 2 \) fed by a voltage divider made up of \( R_2 \) and C.

\[ V = 2 \left[ V_{in} \frac{R_2}{R_2 + \frac{1}{SC}} \right] \]

adding we get

\[ V_o = \left[ -1 + 2 \frac{R_2}{R_2 + \frac{1}{SC}} \right] V_{in} = \left[ -1 + \frac{2S}{S + \frac{1}{R_2C}} \right] V_{in} = \left[ \frac{-S - \frac{1}{R_2C} + 2s}{S + \frac{1}{R_2C}} \right] V_{in} = \left[ \frac{S - \frac{1}{R_2C}}{S + \frac{1}{R_2C}} \right] V_{in} \]

The response function has a zero at \( S = 1/R_2C \) and a pole at \( S = -(1/R_2C) \).
Expressed in magnitude and phase form:

\[ V_o = \frac{Me^{j\phi_2}}{Me^{j\phi_1}} V_{in} = e^{j(\phi_2 - \phi_1)}V_{in}. \]

In other words, the output is of the same magnitude as the input and differs from it by a phase \( \phi_2 - \phi_1 \). Clearly \( \phi_2 - \phi_1 = \pi \), for low \( \omega \), and \( \phi_2 - \phi_1 = 0 \) for high \( \omega \). What is "high" and what is "low" depends, of course, on the size of \( 1/R_2C \).

For \( \omega = 1/R_2C \)

\[ \phi_2 - \phi_1 = \frac{3\pi}{4} - \frac{\pi}{4} = \frac{\pi}{2}. \]
Lock-in Amplifiers—Where And How To Use Them

Because of their inherent versatility, lock-in amplifiers can be used in a wide variety of applications ranging from basic physics to electronic testing to electrochemistry. For example, lock-in amplifiers can be used as narrow band ac voltmeters, phase meters, quadrature detectors, and in some cases, tuned amplifiers and wave analyzers. In addition, they can also be used to detect a "tagged" signal in a complex experiment.

Lock-in amplifiers are ideally suited for use as narrow band ac voltmeters because they are frequency dependent. As a result, a lock-in will respond to only that portion of the signal that is in synchronism with the reference signal. All components of the signal not occurring at the reference frequency are ignored. Although the method used to accomplish this discrimination between the frequency of interest and other frequencies differs from conventional narrow-band filters, lock-in amplifiers do provide 6, 12 or 18 dB/octave roll-offs, depending on the characteristics of the instrument used. Of greater importance than the roll-off rates provided is the extremely narrow bandwidths obtainable with a lock-in amplifier — as narrow as .0004 Hz.

Virtually all lock-in amplifiers can be used as phase meters. This is possible because the output of a lock-in amplifier is equal to the magnitude of the signal occurring at the reference frequency multiplied by the cosine of the angle between the input and reference signals at the detector. As a result, maximum lock-in output, or panel meter deflection, occurs when the cosine is equal to one — a condition that exists when the angle between the input and reference signals at the detector is zero. Since virtually all lock-in amplifiers contain a built-in reference-channel phase shifting network, it is necessary only to adjust the calibrated front-panel "phase adjust" control to provide maximum meter deflection. When maximum deflection is obtained, one need only read the amount of phase shift "dialled in" to determine the phase angle between the signal of interest and the reference at the experiment's signal source. Some lock-in amplifiers, such as the P.A.R. Corp. Model 129A and the Models 126 and 124A, with phase measurement option, contain circuitry that automatically computes the phase angle between the signal of interest and the reference, and provides a direct readout of the computed angle on a front-panel meter.

In many application areas it is desirable to measure the amplitude of specific signal components that occur out of phase, but at the reference frequency. In most cases these out-of-phase components are in quadrature with the in-phase component and may represent reactive components in a complex ac signal. The same mechanism that permits a lock-in amplifier to function as a phase meter also permits it to measure the amplitude of signal components occurring at any given phase angle, relative to the reference signal. All that need be done is to "dial in" the phase angle of interest using the calibrated "phase adjust" controls. The lock-in amplifier will then automatically provide a measurement of the signal component existing at the selected phase angle.

Certain lock-in amplifiers contain tuned signal channels. Included in this class of lock-in are the Models 124A, 120, 122 and HR-B. By monitoring the input to the phase sensitive detector, which is, in fact, the output of the signal channel, via the BNC connectors provided on the lock-in, it is possible to utilize the signal channel as a tuned amplifier. The signal of interest is connected to the lock-in amplifier input and the amplified, filtered output is taken from the monitor BNC connector. No reference signal is required since the phase sensitive detector is not utilized in this application. The degree of filtering obtained depends on the lock-in amplifier being used. The Model 124A provides circuit Q continuously variable from 1 to 100 while the Model HR-B provides circuit Q ranging from 5 to 25.

The Models HR-B and 124A can also be used as wave analyzers. In this application, the high-Q tuned signal channel is employed to filter out all but one specific frequency component of the input signal. By taking measurements at many different frequencies across the lock-in amplifier's operating frequency range, it is possible to obtain a complete frequency spectrum for the signal of interest. An automatic scanning wave analyzer featuring constant bandwidth can be constructed using a Model 124A or Model 126 with a Model 230 Multiplier being operated as a squarer and integrator. The Model 186 contains a built-in noise measurement mode featuring a calibrated 1 Hz bandwidth that is quite useful in determining the best operating frequency in a complex experiment.

Lock-in amplifiers can be used to detect tagged signals even in the presence of otherwise identical signals. In a typical application of this type, the signal of interest is chopped at the lock-in amplifier reference frequency. Because of the frequency and phase sensitivity of the lock-in amplifier's phase sensitive detector, the lock-in will respond only to that part of the complex signal that has been chopped at the reference frequency.
Lock-in Amplifiers—How They Work

The lock-in amplifier is basically a phase sensitive ac voltmeter that compares the input signal against a reference signal, producing a dc level proportional to that part of the input which is synchronous with and in phase with the reference. This comparison takes place in a circuit known as the phase sensitive detector, which can be thought of as nothing more than an electronic polarity reversing switch as shown in Figure A.

The reference signal can be generated within the lock-in itself, if it is equipped with an internal reference oscillator, or can be generated externally by some other source. Regardless of the reference signal source, it is applied to circuitry that converts it into a near-perfect square wave occurring at precisely the reference frequency. This is applied to the phase sensitive detector through a special network that permits the reference square wave to be shifted from 0 to 370 degrees in phase. This phase-shiftable reference square wave is used to "throw" the electronic polarity reversing switch between its inverting and non-inverting positions at the reference frequency.

The input signal contains the information of interest at the reference frequency as well as random noise components and perhaps spurious signals at various discrete frequencies. This composite signal is applied to the phase sensitive detector through the signal channel which contains suitable amplifiers, the sensitivity controls and, if the lock-in is so equipped, input filters.

The relationship that exists between the coherent component of the input signal and the reference square wave at the phase sensitive detector is shown in Figure B. When properly adjusted, the electronic polarity reversing switch transitions are synchronized to the polarity reversals occurring in the coherent component of the input. As a result, the switch will be in its non-inverting position when the coherent component is in its positive half-cycle, and in its inverting position when the coherent component is in its negative half-cycle. Under these conditions, the output will consist of a series of positive half-cycles with spurious signals and noise superimposed.

It should be noted that the output is a series of positive half-cycles related to the coherent component that is "in-phase" with the reference square wave at the detector. The built-in phase-shifting network makes it possible to measure the amplitude of any phase component of the coherent input. For example, to measure the quadrature component it is only necessary to insert a 90 degree phase shift in the reference channel to compensate for the 90 degree shift between the coherent input and its quadrature component.

The detector output is passed through a low-pass filter to produce a dc voltage that is proportional to the average value of the coherent input. In the filtering process spurious signals and random noise are greatly suppressed. The degree to which they are suppressed is the signal-to-noise improvement ratio of the lock-in amplifier, which is equal to the square root of the ratio between the input white noise bandwidth and the output noise bandwidth. The input noise bandwidth is fixed by the operating parameters of the experiment. Normally, these parameters are determined by the requirements of the experiment and are not subject to change. The output noise bandwidth is determined by the output filter time constant which is variable and can be set on some lock-ins to be as narrow as 416 microhertz. A detailed discussion of noise is given on page 6.
Noise—What It Means And What It Does

Noise can be defined as unwanted disturbances that are superimposed on the signal of interest. The operation of a lock-in amplifier is such that virtually all the random noise and other interfering signals accompanying the signal of interest are eliminated, resulting in an output that is a dc level virtually free of ac components.

In the case of random noise the performance of a lock-in amplifier can be specified in terms of its “signal-to-noise improvement ratio”. This is the ratio of the signal-to-noise ratio at the input to the signal-to-noise ratio at the output; in other words, how much the signal-to-noise ratio is improved by the lock-in. For example, if the input signal-to-noise ratio were 1:10 and the output signal-to-noise ratio were 10:1, the signal-to-noise improvement ratio would be 100:1. The improvement ratio is equal to the square root of signal source bandwidth divided by the equivalent noise bandwidth of the lock-in amplifier. The equivalent noise bandwidth is equal to (1/4TC) for 6 dB/octave output filters and (1/8TC) for 12 dB/octave output filters, where TC is the selected time constant. As can be seen, doubling the time constant effectively halves the equivalent noise bandwidth. However, the signal-to-noise improvement ratio is proportional to the square root of the equivalent noise bandwidth. Thus, in order to increase the improvement ratio by a factor of 10, for example, it is necessary to increase the output filter time constant by a factor of 100.

The total noise reaching the phase sensitive detector is a combination of both the noise accompanying the input signal and the noise generated internally by the lock-in signal channel electronics. Under most experimental circumstances, the noise accompanying the input is fixed and cannot be further reduced. The input noise bandwidth of the lock-in is also usually fixed, either by virtue of an untuned or wideband signal channel or by virtue of the bandwidth selected in a tuned lock-in. Thus, the only two variables are the internally generated noise and the output filter time constant.

Under ideal conditions, where the internally generated noise is equal to zero, the output noise level would be equal to the input noise level divided by the signal-to-noise improvement ratio. If, however, the internally generated noise were to effectively increase the noise reaching the detector by 10 dB, the output noise level would also increase by 10 dB. In order to compensate for this increase, it would be necessary to increase the output filter time constant by a factor of 10.

When viewed in terms of the output of a lock-in amplifier, the experimenter is faced with a choice between a worsened signal-to-noise ratio or a greatly increased output filter time constant. A worsened signal-to-noise ratio reduces resolution, thus increasing the minimum amplitude at which the coherent component of the input signal can be detected. On the other hand, greatly increasing the output filter time constant beyond that required without internal noise greatly increases the time required to obtain the same results, and greatly reduces the ability of the lock-in to follow rapid changes in the level of the coherent input.

A far better alternative to worsened signal-to-noise ratio or increased time constant is the reduction of internally generated noise. This noise is generated by all parts of the circuitry, but the noise generated within the preamplifier predominates since it is further amplified by following stages. One way of specifying preamplifier noise performance is to speak of its noise figure, which indicates the amount of noise it adds to the source thermal noise. Source thermal noise, or Johnson noise as it is usually referred to, is used as the standard of comparison because it is completely predictable, always present and is the least amount of noise that can possibly accompany the signal. For pure resistance sources its value in volts rms can be calculated from the equation:

\[ e_n = \sqrt{4kTBR_s} \]

where:

- \( e_n \) = rms noise voltage within bandwidth measured
- \( k \) = Boltzmann’s constant = 1.38 \times 10^{-23} \text{ joules/kelvin}
- \( T \) = absolute temperature of source in kelvins
- \( B \) = bandwidth over which measurement is taken
- \( R_s \) = Source resistance

Noise figure is not constant; rather, it varies as a function of frequency, source resistance and temperature. When the loci of all points having the same noise figure are plotted as a function of source resistance and frequency at a constant temperature, the result is a noise figure contour. A full set of noise figure contours specifies the noise performance of the preamplifier.

Noise figure contours can be used in a number of ways. If the frequency and source resistance are fixed, the optimal preamp can be selected by comparing noise figure contours. Conversely, an experiment can be optimized by selecting an operating frequency and/or source resistance that provide the best noise performance with the available preamplifier. In addition, once the operating parameters have been established, the minimum discernible signal can be calculated in terms of volts per root hertz by solving the equation: minimum discernible signal = \( e_n \times 10^{NF/20} \) where \( e_n \) is the source thermal noise in volts per root hertz and NF is the noise figure in dB as shown on the noise figure contour for the preamplifier being used. Minimum discernible signal is the signal level which appears equal to the noise level at the output.

Princeton Applied Research has available an illustrated technical note entitled “How to Use Noise Figure Contours” which, as the title implies, describes how to apply published noise figures to your experiment. Also available is a colorful wall chart which greatly simplifies preamplifier selection, and which provides useful data about commonly used physical constants and relationships. Both are yours for the asking.
When Princeton Applied Research introduced the first commercially available lock-in it was intended for use primarily in photometric applications. Since then the lock-in has been used in a wide variety of measurements in virtually every discipline. The applications summarized in these pages are just a few of the areas in which the lock-in amplifier can be used. For assistance with your specific application, write or telephone your local P.A.R. Corp. representative or the Princeton headquarters.

Weak Photoluminescence Decay

The measurement of repetitive low-intensity light signals is becoming increasingly more important. One method available of obtaining these measurements is to photograph the CRT display produced by each light signal. However, this method is extremely time inefficient and does not provide adequate sensitivity because of detector noise and low frequency drift. However, by modulating the light signal and using a lock-in amplifier to detect the modulated signal, detector noise and low frequency drift can be significantly reduced. In one such application, which involves the study of photoluminescence decay in alkali halides, a mechanical light chopper is operated at half the exciting lamp’s repetition rate. The resulting decay is sampled through time with the output applied to the lock-in. The lock-in reference frequency is the chopping frequency. The electromagnetic disturbances resulting from the lamp flash are eliminated since they occur at twice the reference frequency—a frequency to which the lock-in is insensitive. All other noise is greatly attenuated because it is not synchronous with the chopping frequency.

Self-Balancing Bridge

In this servo loop application, lock-in amplifiers can be employed in self-balancing ac bridges for use with strain gauges and other transducers. In one example, an exciting voltage is placed across the bridge and across a series network consisting of a fixed resistor and a photoconductor. The outputs of both are applied to a summing amplifier, one to the inverting input and the other to the non-inverting input. The output of the summing amplifier is applied to a lock-in amplifier utilizing the exciting voltage as its reference input. The dc output of the lock-in is used to drive an incandescent lamp located at the photoconductor. When the voltage output of the bridge and the photoconductor network are equal, the lamp will be lit to some value of output, determined by the setting of output zero adjust control on the lock-in. If the transducer in the bridge changes its resistance value because of a change in the parameter it is monitoring, the output voltage of the bridge will also change, causing a change in the input to the lock-in. The result of this change will be a change in the output of the lock-in, and thus in the brilliance of the lamp. As the lamp output changes, the resistance value of the photoconductor will also change and eventually a new equilibrium will be obtained. By calibrating the current flowing through the photoconductor network, a direct readout of the measured parameter can be obtained.

Background noise near null could cause errors by obscuring the null point. The major advantage to using a lock-in amplifier in this application is its immunity to background noise of all types, thus providing greater accuracy and loop stability than otherwise obtainable.
Birefringence In Materials Research

Considerable research is being directed towards the characterization of flux-grown crystals, including magnetic garnets such as YIG, YbB₁₂, GdDub, and CVB₁₂. One approach is to measure the effects on a beam of polarized light passing through a sample crystal held under stress. By slowly moving the sample across the beam, it is possible to obtain data regarding the sample's uniformity. A typical characterization experiment utilizes a monochromatic light beam and a quartz retarder to modulate the beam at some convenient frequency in the range of a few hundred hertz. The modulated beam is passed through the sample and through a motor driven compensator to a cooled lead sulphide detector. The detector output is applied to the input of a P.A.R. Corp. lock-in. The reference for the lock-in is obtained from the same generator used to drive the quartz retarder. The output of the lock-in is used to drive the motor which controls the compensator. This servo loop is adjusted so that it is in balance when all of the birefringence introduced into the light beam by the sample is eliminated by the compensator. As the crystal sample is moved across the light beam, any change in its internal structure causes a change in the birefringence introduced into the beam. A change in birefringence causes the output of the lock-in to change, which in turn causes the motor to readjust the compensator until balance is again achieved. A hard copy of the results is obtained by connecting the output of the lock-in to the Y input of an x-y recorder, the X axis of which is driven by a position transducer indicating the movement of the sample through the light beam. A P.A.R. Corp. lock-in was chosen for this application because of its ability to ignore signals produced by incident light and quiescent detector current while measuring only that portion of the light modulated by the quartz retarder.
IR Quantum Counting

In order to observe and measure certain energy level changes in rare earths, it is necessary to pump the sample with multi-wavelength IR radiation. An immediate problem in these double-excitation experiments is measuring the specific quantum change of interest in the presence of many other quantum changes caused by the broad band IR pumping. One practical approach is to “tag” the quantum change of interest so that it is easily identifiable. A typical tagged-signal experiment utilizes two independent monochromatically filtered IR light sources to provide pumping excitation. One light source is used to excite the electrons of each rare earth atom in the sample from its ground state to an intermediate excited state. The second light source, which is mechanically choppered at a convenient frequency, is used to raise the excited electrons of interest to the desired higher energy level. The resultant fluorescence is monitored by a photodetector. A P.A.R. Corp. lock-in, which is reference locked to the mechanical light chopper, is used to measure the output of the photodetector. Since the desired fluorescence is tagged by the chopped IR pump, it is coherent with the lock-in amplifier’s reference; the other fluorescence signals are not. Because of the lock-in amplifier’s ability to reject all signals that are not coherent with the reference, the resulting output will be proportional to only the tagged quantum changes. Superior dynamic reserve enables the P.A.R. Corp. lock-in to extract very weak tagged signals from high fluorescence background levels. In addition, the low noise preamplifier and narrow-band filtering employed in the P.A.R. Corp. lock-in enable it to extract the tagged signal even when it is so weak that it is obscured by the dark current of the detector.

Laser Stabilization

Lock-in amplifiers are ideally suited to applications where a modulated signal must be stabilized with a dc control voltage. In a typical laser stabilization application, a piezoelectric transducer is fitted to one end of the laser tube. The reflecting mirror at that end of the tube is attached to the transducer so that changes in a dc voltage applied to the transducer will cause changes in the cavity length, and thereby cause changes in the operating frequency. The output of the laser is mixed with a standard frequency to produce a beat frequency within the lock-in’s operating range. The lock-in is driven by an externally generated reference signal equal in frequency to the beat produced when the laser is operating at the desired frequency. When the beat frequency is equal to the reference frequency, the lock-in amplifier’s dc output will be at its maximum.

Any change in laser frequency will cause a change in the beat phase, and thereby a decrease in the dc output of the lock-in. This change in dc output causes the crystal transducer to change the position of the mirror, and thereby stabilize operating frequency of the laser. By carefully adjusting the transducer, it is possible to obtain completely automatic frequency stabilization. Other schemes have been designed that expand upon this basic approach to provide very tight control of the laser frequency.

[Diagram of Laser Stabilization Measurement System]

[Diagram of IR Quantum Counting System]
Applications
(Continued)

Temperature Controller

It is important in many areas of research to maintain the experiment at a specific temperature above or below ambient. Even though there are some excellent temperature controllers available commercially — the P.A.R. Corp. Model 152 for example — there are some applications where a customized temperature controller is desirable. One widely used method for controlling temperature is to balance the output of a temperature sensor against a stable reference voltage, chop the difference signal and detect it with a lock-in amplifier, the output of which is used to control the heating element.

In the system illustrated in Figure A, the output of the temperature detector, in this case a thermocouple, and the stable voltage are applied to a modulator. The two signals are algebraically summed and modulated by an electronic chopper driven by the reference output of the lock-in amplifier. The modulated signal is applied to the lock-in amplifier input where it is phase sensitive detected to produce a dc level which is applied to the heating element. The zero offset control of the lock-in amplifier is used to set the desired heater element quiescent temperature level. When the temperature detector output is equal to the set voltage applied to the modulator, the modulator output will be zero; thus, only the voltage set manually with the zero offset control will be applied to the heater element. If the temperature drops, the output of the detector will change in such a manner as to increase the input to the lock-in, which in turn increases the voltage being applied to the heater element. If the temperature rises above the desired level, the input to the lock-in will drop which causes the voltage being applied to the heater element to drop.

Another commonly used temperature controlling system is illustrated in Figure B. The temperature sensor, which is a negative temperature coefficient thermistor, is part of a bridge. $R_1$ and $R_2$ are equal in resistance. $R_3$ is adjusted to equal the resistance of the thermistor at the desired temperature. The output of the bridge is applied to the differential input of a lock-in amplifier, the output of which is used to power the heating element through a suitable amplifier. Bridge excitation is obtained from the reference oscillator of the lock-in, or from an external oscillator which is also used to provide the lock-in with its reference signal. The zero offset control of the lock-in amplifier is used to set the desired heater element quiescent temperature. When the system is in balance at the desired temperature, the bridge output will be zero; thus the lock-in amplifier's output will reflect only the setting of the zero offset control. If the temperature should increase, the bridge output will go negative, which in turn reduces the voltage applied to the heater element. If the temperature decreases, the bridge output will go positive, which increases the voltage applied to the heater element.

As in most servo loops, the system rolloff rate must be maintained at no more than 6 dB/octave if stable operation is to be obtained. Therefore, the output filtering of the lock-in amplifier must be set for 6 dB/octave. The sensitivity of the system is set with the lock-in's sensitivity control; the response time by the lock-in's time constant switch. Since the lock-in responds only to the coherent output of the modulator, the system is virtually immune to noise.