496 Lock-In Amplifier and Integrating ADC

BENTHAM

Overview

The Bentham 496 module combines a lock-in amplifier and an integrating analogue to digital converter (ADC). This unit is typically housed in the Bentham 417(T): power supply, USB interface and display.

Lock-In Amplifier

The lock-in amplifier takes an AC input signal (the detector's response to a chopped optical source), to provide a proportional DC output (0-10V) to the ADC. This latter output can be viewed on the 417(T) bin channel A. This process is achieved by the comparison of the detector and optical chopper reference signals, locking-in.

The 496 amplifier is a dual input device, with square wave demodulation to give the best possible signal to noise ratio with the signal encountered in chopped light systems. The high impedance voltage input is ideal for connection from current pre-amplifier (Bentham 477), detectors operating in the photoconductive mode (such as the lead salt detectors), or to the output of the Bentham DH-Py pyroelectric detector.

To operate, it is necessary to determine the relative phase difference between the recovered detector signal and the chopper reference, and to note the phase to obtain maximum positive signal to input to the ADC. The device is auto-ranging sensitivity, the time constant provides analogue averaging to the signal and should be minimised where possible, with preference given to digital averaging of the ADC.

Whilst this amplifier is typically computer controlled via USB/I^2C, it is possible to operate locally, by depressing the black rem button. One may change the present sensitivity, time constant, phase quadrant, phase variable and input.

Reference input, the output signal from the chopper control module is also required for use. The attributes of the lock-in amplifier vary per detector used. This should be set up before use.

Integrated ADC

The ADC uses a continuously running voltage to frequency converter to produce a pulse train whose frequency is proportional to the instantaneous input voltage. The pulses are then accumulated in a counter. The ADC provides 2000 counts per volt with a maximum of 20000 counts.

At 100ms intervals, the contents of the counter is transferred to an output buffer and the counter reset to zero. The total number of pulses accumulated by the counter in any counting period represents the true average of the signal during that counting period; if the accumulated pulses from a number of counting periods are added and normalised then a true average over a longer period is obtained.

The input to the ADC can be either from the amplifier, or another, auxiliary source, selected by depressing the aux button.

The input to the ADC is offset giving the unit a small negative range. This ensures that negative going noise peaks, occurring in near zero signals, are correctly averaged while retaining most of the available resolution for positive signals.

Front Panel

The front panel of the 496 is displayed below. Features of this are as follows:

Figure 1: Front panel of 496 lock-in amplifier/ADC

- Remote button: this allows the user to return control of the unit to the front panel (local mode). When the accompanying yellow LED is lit the unit is being control via the computer
- Input Button: in local mode pressing this allows the user to switch between the two inputs
- 5 line LCD screen: this shows the input being currently used, measured voltage and chopping input frequency, phase of the input signal, the signal-to-noise ratio, the gain currently being applied and the integration time
- Gain +/-: in local mode this allows the user to vary the amount of gain between 2^{-4} and 2^{14}
- Integration time +/-: this allows the user to vary the gain used in local mode. Increasing the integration time can help with noisy signals but also increases the time take to perform the measurement
- Input 1 and 2 BNC connection: used to input external signals into the 496
- Ref BNC: Attach the chopper output here

BenWin+ Setup Window

The window used to set the 496 up in BenWin+ is displayed below:

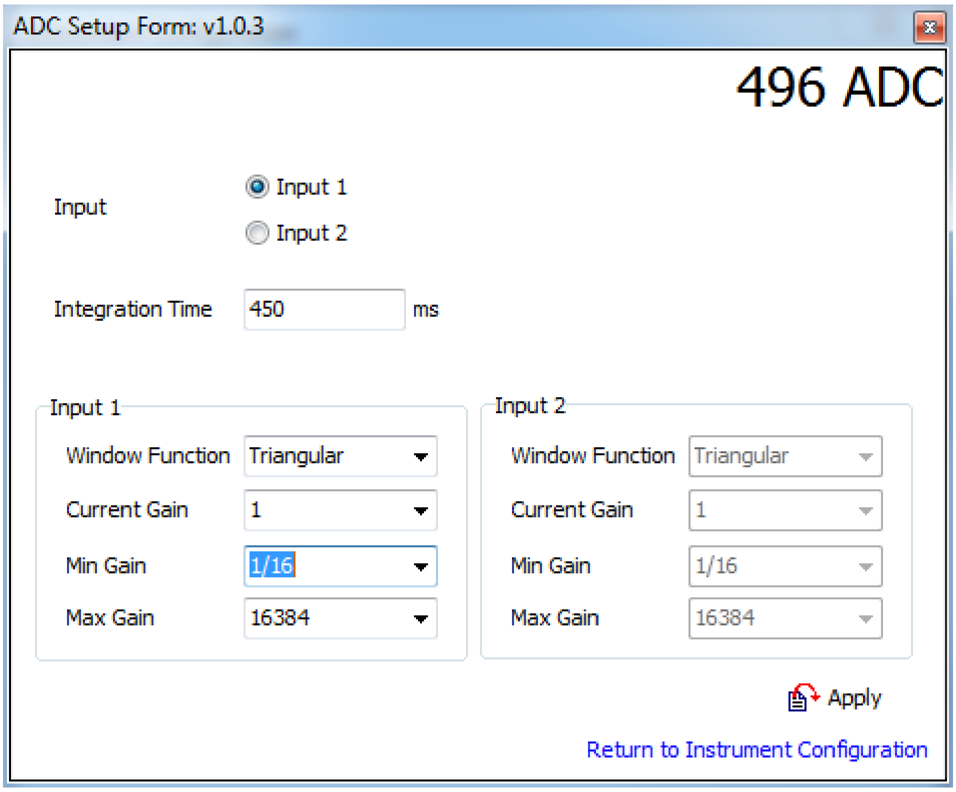


Figure 2: BenWin+ control panel for the 496

Features of this are much the same as the front panel with regards to input selection, integration time and current gain. However, this window also offers additional functionality such as the setting the minimum and maximum gain to be used and the Window Function which applies a window weighting function to the signal to minimise spectral leakage. Different window functions have different frequency response characteristics and therefore the user should choose the optimal window for their needs by considering the frequency resolution and dynamic range requirements.

Specification

Inputs	2 remotely/manually selected
Frequency range:	10Hz to 2kHz
Gain Ranges	Binary gain, 2^{-6} to 2^{14}
Input Impedance	100MΩ/25pF, Pseudo differential
Dynamic Reserve	Not applicable- digital demodulation
Gain Accuracy	+0.15%
Gain Stability	200ppm/°C
Output Stability	5ppm/C to 500ppm/C depending on sensitivity
Time Constant	10ms to 10s
Display	Signal, frequency, phase and SNR displayed to two decimal points
ADC Resolution	16 bit
ADC Speed	52 kHz
Input Range	0 to 10V
Linearity	< 0.025% departure from linearity from zero to full scale
Interface	USB (via 417/417T base unit)

Lock-In Amplifier Theory of Operation

Although the 496 is completely autonomous when locking in on the inout signal, it may be useful for the user to understand the concept of how it operates. As such the following is a brief description of how a typical lock-in amplifier works.

All lock-in amplifiers, whether analogue or digital, rely on the concept of phase sensitive detection for their operation.

Stated simply, phase sensitive detection refers to the demodulation or rectification of an ac signal by a circuit which is controlled by a reference waveform derived from the device which caused the signal to be modulated. The phase sensitive detector effectively responds to signals which are coherent (same frequency and phase) with the reference waveform and rejects all others.

In a light measurement system the device which causes the signal to be modulated is usually a chopper, the reference waveform is an output coherent with the chopping action provided by the chopper and the ac signal is the signal from the photodetector.

As the lock-in is a solution to a measurement problem we can usefully describe its action and composition by looking at the sort of problems that occur when light measurements are pushed to the limit.

Consider a simple light measurement system being used to measure transmission. Light from a stable light source is passed through a sample and reaches a detector. The resulting electrical signal from the detector is amplified and displayed on a meter. The meter reading gives an indication of the amount of light transmitted by the sample.

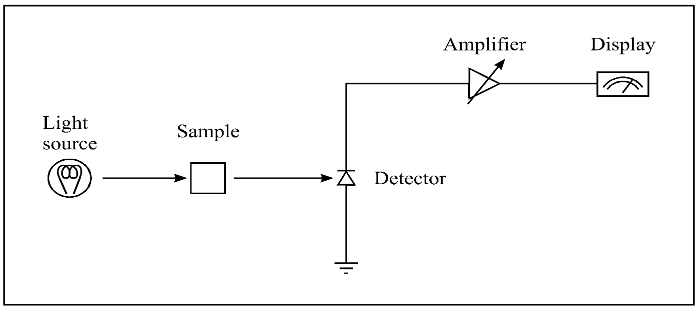


Figure 3: Schematic of Lock-in Device

With medium to high transmission samples this system would be expected to give precise and reproducible results. Samples of increasing optical density could be accommodated by increasing the gain of the amplifier. What is always noticeable in such systems, however, is that as the signal level falls and the amplifier gain is increased so the precision with which the results can be recorded also falls. This is due to noise.

Noise in this sense is anything which contributes to the meter reading but which is not due to the parameter being measured. It is generated in all parts of the electrical circuitry but in light measurement systems it is dominated by noise from the detector or noise associated with the optical signal.

The following diagram shows the distribution of noise and signal power from the optical detector in terms of power per unit bandwidth as a function of frequency. This can be used for a situation where the density of the sample is so high that the signal is smaller than the noise with the result is that the instrument becomes unusable.

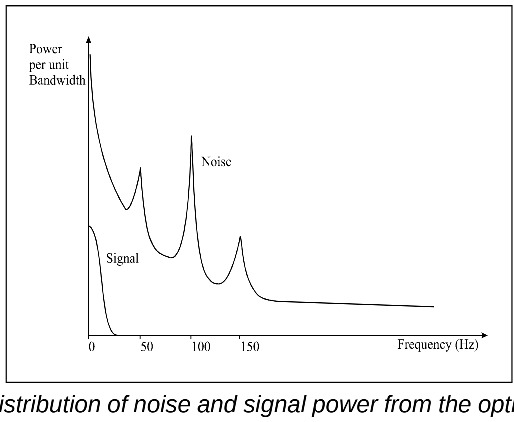


Figure 4: distribution of noise and signal power from the optical detector in terms of power per unit bandwidth as a function of frequency

The most noticeable feature of this curve is the steady increase in noise power which occurs as 0 Hz is approached.

In our transmission measuring instrument this low frequency noise has several sources including flicker noise associated with semiconductor devices, variations in dark current (especially in photomultipliers) and variations in ambient light leaking into the instrument and reaching the detector.

At higher frequencies the spectrum flattens out to give a reasonably constant shot noise background which is associated with the quantum nature of light. The small peaks of 50Hz and 150Hz are due to electrical interference from the power system. The larger peak at 100Hz is due to light from room lighting leaking into the instrument and reaching the detector. It is important to note that the y axis in this diagram is in units of power per unit bandwidth so the total noise and signal powers are represented by the area under the corresponding curves. Clearly therefor we can immediately improve the signal to noise ratio in this system by using an electronic filter to reject the higher frequency components which do not contain any signal information. Unfortunately the information relating to the transmission of the sample is also near zero hertz, so using a low pass filter to reject all noise components above say 30Hz will give only a small improvement in signal to noise ratio.

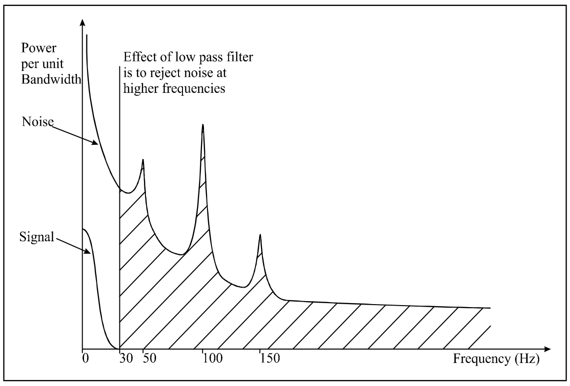


Figure 5: Effect of Low Pass Filter

What we really need to do if we want to measure high optical density with this system is to move the signal information away from the high noise zero hertz region.

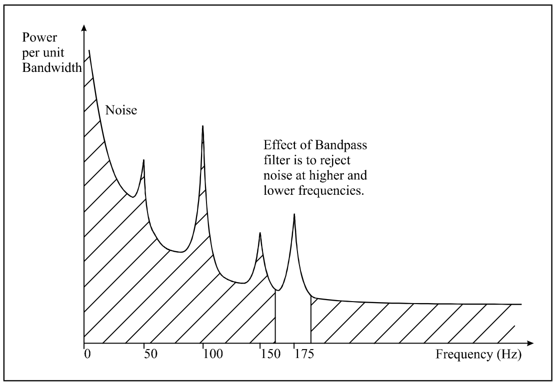


Figure 6: Effect of Bandpass Filter at 175 Hz

We can do this by placing an optical chopper, which will periodically interrupt the light between the light source and the detector. The diagram below shows the detector output spectrum in this new situation with the chopper running at 175Hz.

From the diagram this looks like a good move. We have moved the signal away from a region where the background noise is high to a region where it is low. We can now pass the signal through an electronic bandpass filter which will reject both the noise at higher and lower frequencies (including zero hertz) and hence significantly improve the signal to noise ratio.

The problem now is that the signal is ac, i.e. its average value is zero so to record a value from it we must first rectify it. We might end up with an arrangement as shown in Fig 7 which includes an amplifier, a tuned filter whose centre frequency is at 175Hz, a rectifying circuit and a display.

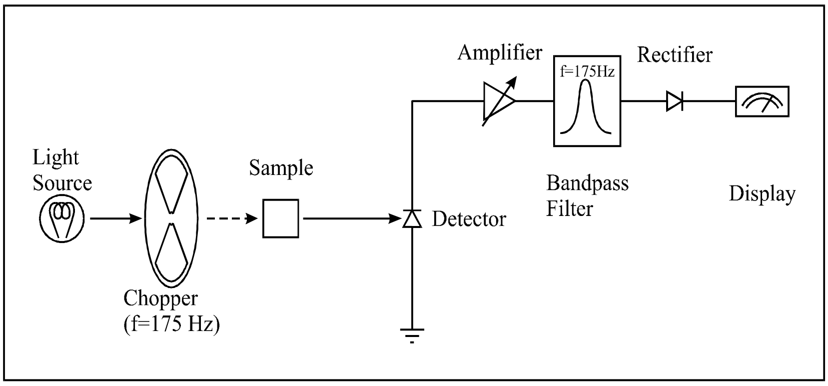


Figure 7: A standard measurement system using lock-in amplifier

Tuned amplifiers, as these devices are called, are used in some simple systems but they suffer from 3 major disadvantages.

The first concerns Q.

Q is defined as the centre frequency of a filter divided by its bandwidth.

In this application the narrower the bandwidth of the filter, the greater is the noise rejection.

The maximum Q typically achievable for a tuned amplifier is in the region of 100, but in a demanding measurement situation we might need a Q of 1000 to achieve acceptable signal to noise ratio.

Secondly, if such a filter could be produced any small shift in chopping frequency would result in large changes in output due to misalignment between signal frequency and filter centre frequency.

The third problem lies in the rectifying device and the way it responds to noise which passes through the filter. Using a normal rectifying device such as a semiconductor diode the noise itself will be rectified and will thus give rise to a dc level at the meter which will be indistinguishable from that derived from the signal.

All these problems are overcome by the phase sensitive detector.

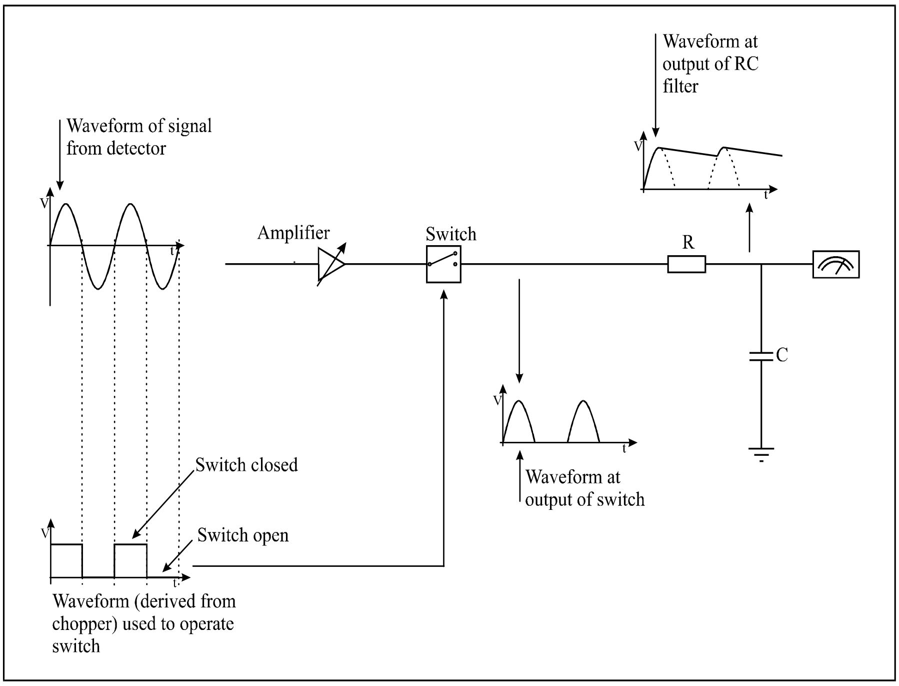


Figure 8: Synchronous Filter

As first glance this device looks very simple.

As before there is circuitry to amplify low level signals, but now there is no tuned filter or rectifying diode. Instead the amplifier is followed by a switch which is operated by a waveform derived from the chopper. When the level from the chopper is high the switch is closed and the output of the amplifier is connected directly to a low pass filter consisting of a resistor (R) and a capacitor (C). When the output of the chopper is low the switch is open and no connection is made.

Rectification of the signal occurs when the waveform controlling the switch is exactly in phase with the ac signal at the input to the switch, hence the sometimes used description, synchronous rectifier. More importantly when the switch is closed the noise associated with the signal passes through un-rectified to the low pass RC filter beyond where it is smoothed or averaged to its mean value of zero.

The device behaves as a bandpass filter and performs the same function as a tuned amplifier followed by a rectifier but with the following advantages.

- The effective bandwidth and hence noise rejecting capability of the device is determined only by the values of the components used in the low pass filter. In fact the bandwidth is given by 1/4T where T is the time constant 1 of the RC filter. A time constant of 2.5 seconds will thus give a bandwidth of 0.1Hz which at centre frequency of 175Hz corresponds to a Q of 1750.

- The centre frequency of the filter is locked (hence lock-in) onto the chopper's frequency. The signal can never drift outside the pass band of the filter.
- Any noise which is present at the output is equally distributed about the mean value resulting from demodulation of the signal. It does not give rise to a dc level as with the diode rectifier and increasing the time constant will reduce its' magnitude.

An alternative approach is to visualising the phase sensitive detector is to consider the switch as a multiplier.

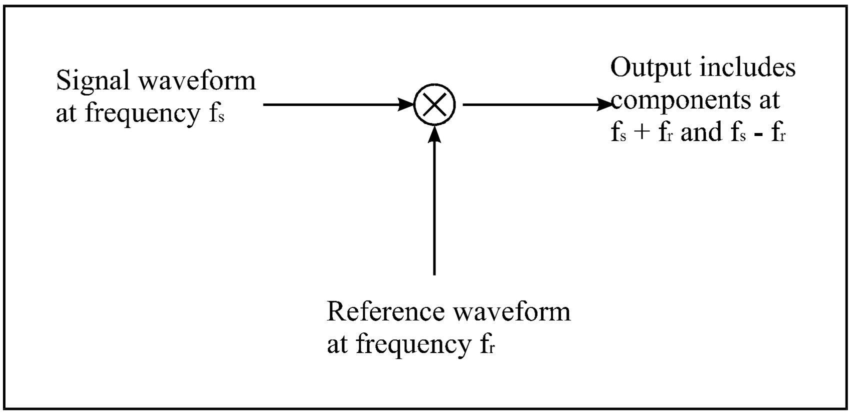


Figure 9: Alternative visualisation of phase detector

Assuming that both the signal and reference waveforms are sinusoidal then the output of the multiplier will contain components at frequencies of fs+fr and fs-fr where fs and fr are the frequencies of the signal and reference waveforms respectively.

If fs=fr as is the case where the reference waveform is derived from the device which is modulating the signal then there will be an output at 0Hz i.e. dc. Any other component in the signal e.g. a noise component at a frequency of fn will give rise to an ac output at frequencies of fn-fr and fn+fr which will be smoothed or averaged to the mean value of zero by the low pass filter.

As the time constant (RC) of the filter is increased so the attenuation of the higher frequency components from the multiplier will increase thus effectively reducing the bandwidth of the overall device.

Some Practical Points

So far we have considered the phase sensitive demodulator as a single pole switch which is capable of synchronous rectification of only half of the signal. For a signal recovery device such a waste of signal information would be unacceptable. Commercial lock-in amplifiers therefore include a full wave synchronous demodulator which usually works by using inverting and non-inverting amplifiers to produce anti-phase versions of the signal (i.e. 180° out of phase).

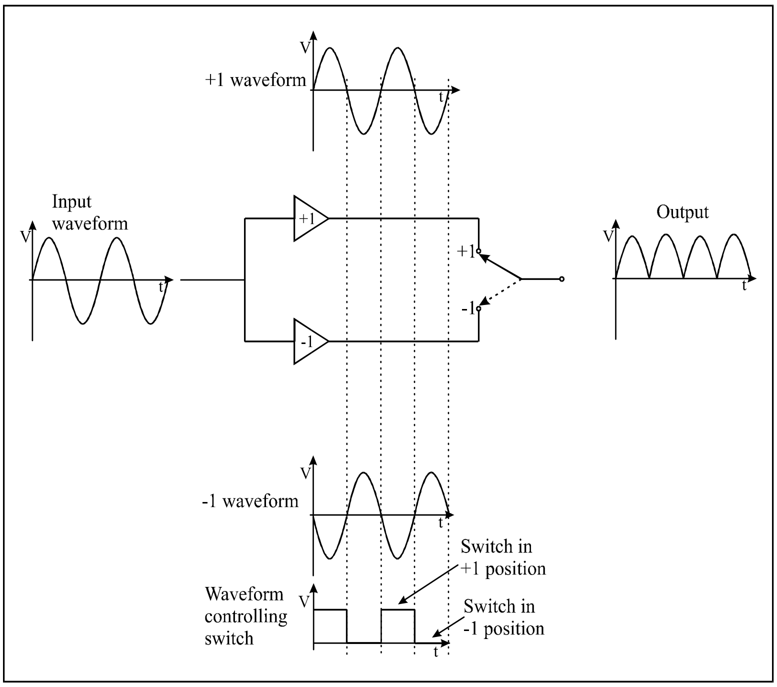


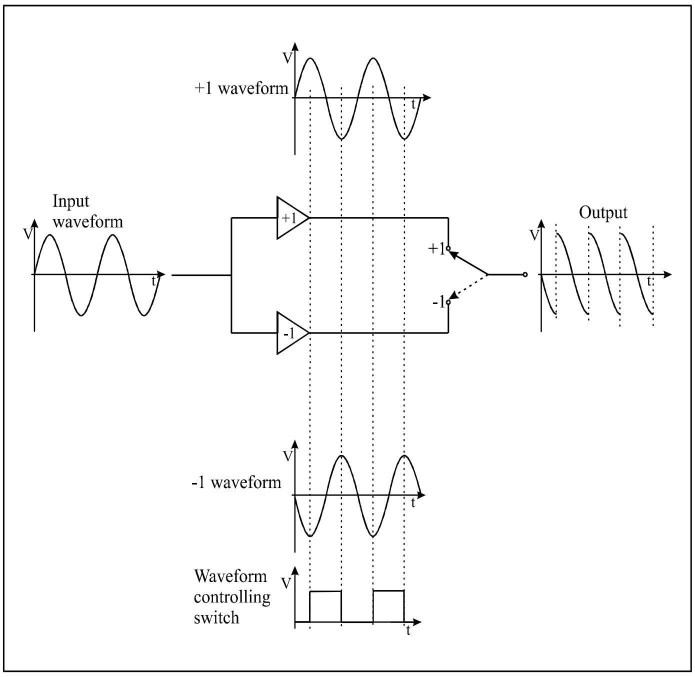
Figure 10: Rectifier operation

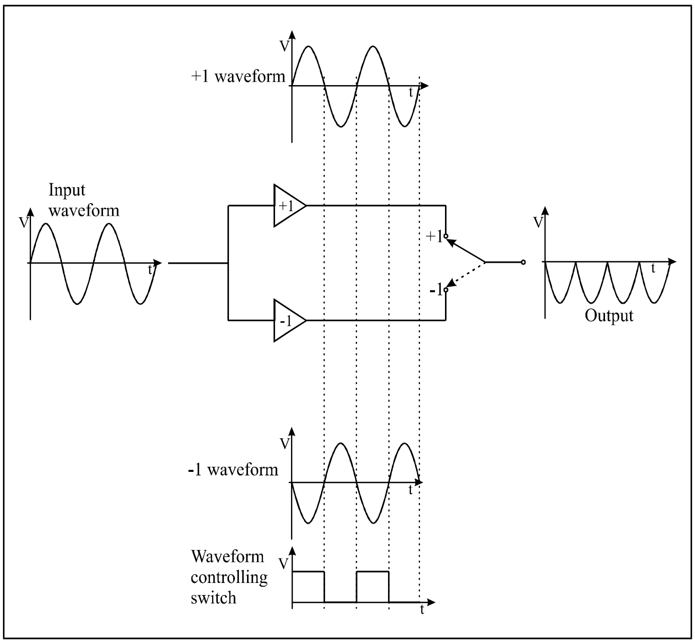
In all the diagrams shown so far the reference waveform available at the phase sensitive demodulator has conveniently been in phase with the signal - the condition for maximum output and hence best signal to noise ratio.

In any real experimental set-up we cannot always be sure that the relationship between the signal from the detector and the reference waveform from the chopper will be in phase. The phase relationship will be dependent on the position of the chopped light beam relative to the reference pick-up on the chopper and also on any phase modification of the signal introduced by the detector.

To allow perfect phase matching at the demodulator, lock-in amplifiers include comprehensive and flexible phase shifting circuitry which allows a phase shift of over 360 degrees to be introduced. The phase controls on the lock-in usually take the form of a continuously variable of 0-95 degrees and 3 fixed increments of 90 degrees giving 365 degrees in total.

The following diagrams show how the output of the psd varies with the reference to signal phase relationship. In figs 11a, 11b and 11c the reference waveform has been shifted by 90°, 180° and 270° respectively from that in Fig 10.


+1 waveform
Input
waveform
Output
+1
-1
-1 waveform
Waveform
controlling
switch


+1 waveform
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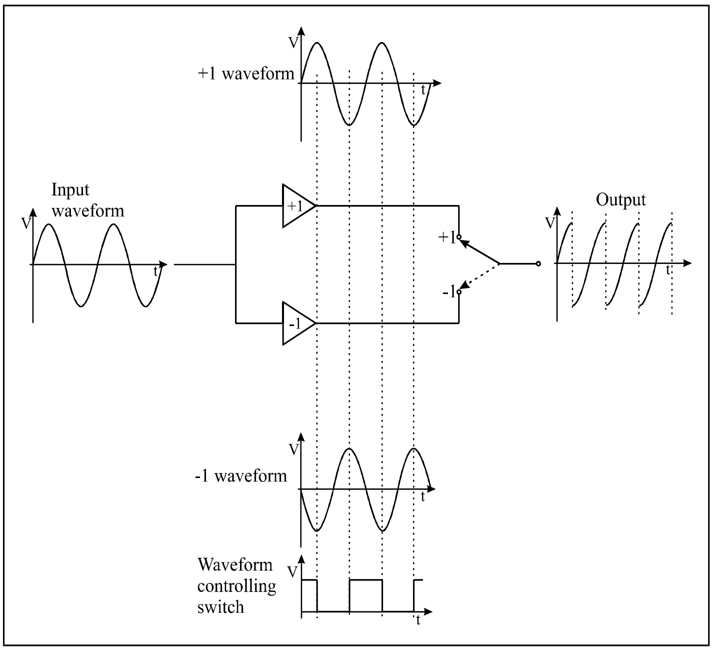

Figure 11a: phase shift of 90, 11b: phase shift of 180 and 11c: phase shift of 270

Harmonic Response

In the multiplier model of our psd we made the assumption that the two waveforms were sinusoidal, but in practice, when using a simple switch we are effectively multiplying the signal not by a sine wave but by a square wave (the switch is either closed or open). Now a square wave is equivalent to a sine wave of the fundamental frequency plus all the odd harmonics of the fundamental 2. A psd using a simple switch therefore behaves as a filter centred at the fundamental but with windows at each odd harmonic.

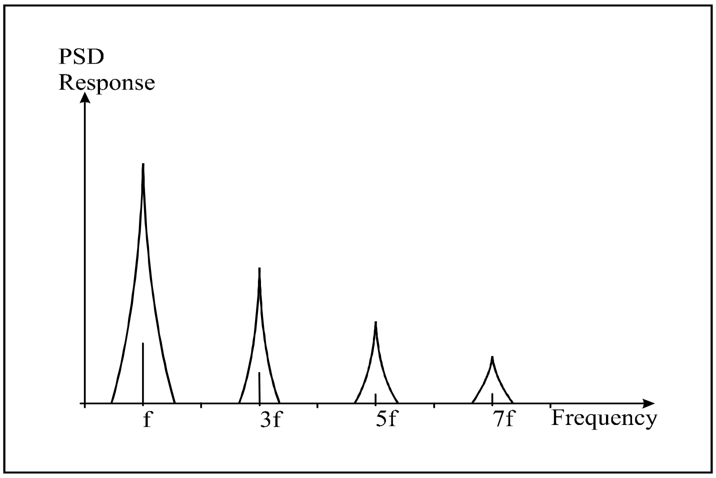

Figure 12: PSD response

At first glance this looks to be a problem in that the noise present in these harmonic windows will be passed through to the low pass filter and degrade the signal to noise ratio at the output of the lock-in.

However, this problem is not as bad as it looks, especially if the modulating device is a chopper.

First of all the harmonic windows do not have the same “sensitivity” as the fundamental but reduce as 1/harmonic reflecting the Fourier composition of the square wave switching the demodulator 2.

Secondly, if the modulating device is a chopper, then the signal itself will not be sinusoidal but more typically triangular or trapezoidal both of which have significant odd harmonic content so the harmonic windows will also transfer additional signal information to the output of the lock-in.

The most dangerous situation is one in which a harmonic window coincides with a point in the spectrum where a large discrete interference is present but this should never occur in a properly designed light measurement system with the correct choice of chopping frequency.

It is unfortunately the case that lock-ins are sometimes used to compensate for incorrect optical design.

The most commonly encountered misconception regarding the use of lock-in amplifiers in light measurement concerns their ability to disregard a constant dc light level most commonly caused by ambient light leaking into the system and reaching the detector. This after all is the reason for including the chopper so that the signal information is shifted away from the dc region.

The next intuitive step taken by many is to assume that there is no longer any need to make the system light tight - as long as ambient light reaching the detector does not cause saturation everything will be OK.

This is not correct.

Referring back to the earlier paragraphs we pointed out that the shot noise background is caused by the light itself and is associated with its quantum nature.

Ambient light leaking into chopped light systems will always degrade signal to noise performance even though it does not give rise to a dc output from the lock-in.

You can often improve signal to noise ratio just by making things light tight - if you cut down light leaks you may be able to reduce the time constant and make your measurements faster. The time constant of the RC circuit in seconds is given by the product of the resistance in ohms and the capacitance in farrads.

WEEE statement

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