Customer Information

How To Reach Us

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Standard Terms

Billing:	Net 30
Prices:	Prices in this catalog are U.S. list prices. International prices may vary.
Leasing:	Lease terms are available. Contact SRS for details.
FOB:	Sunnyvale, CA 94089
Shipping:	Shipment is by UPS Ground unless otherwise specified.

Warranty and Repairs

Warranty:	All instruments unless otherwise stated are warranted to be free from any defects in material and workmanship for one year from the date of purchase. Some instruments have different warranties. See the individual instrument descriptions for details.
Calibration:	All instruments are calibrated to NIST traceable standards. A certificate of calibration is available at no charge if specified at the time an order is placed.
Repair Charge:	10% of the instrument price. Minimum repair charge is $175. All returned items must be accompanied by an RMA number to ensure proper servicing. Contact SRS to obtain a RMA number.
Options:	There is a $250 charge to retrofit any option.

International Representatives

A complete list of Stanford Research Systems' international representatives can be found on page 248.

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SRS Catalog
PART 2

FFT Signal Analyzer

Model SR785 — 100 kHz Two Channel Dynamic Signal Analyzer

- ***DC to 102.4 kHz Bandwidth***
- ***90 dB Dynamic Range***
- ***16-bit A/D Conversion***
- ***Low Distortion Synthesized Source***
- ***ANSI Standard Octave Measurements***
- ***145 dB Dynamic Range Swept Sine Mode***
- ***Order Tracking***
- ***20 Pole/Zero Curve Fitting***
- ***Time/Histogram Mode***
- ***8 Msamples Transient Capture Memory***
- ***Internal and External Triggering***
- ***File Output to SDF, MAT, UFF, and ASCII***
- ***Graphics Outptut to GIF, EPS, PCX, HPGL***
- ***3.5“ MS-DOS Compatible Disk Drive***
- ***GPIB and RS-232 Interfaces Included***
- ***Windows Data Viewer Program Included***

SR785 Overview

The SR785 Two Channel Dynamic Signal Analyzer is a precision, full-featured signal analyzer that offers state-of-the-art performance and a wide selection of features at a price that's less than half that of competitive analyzers. Building on its predecessor, the SR780, the SR785 incorporates new firmware and hardware that make it the ideal instrument for analyzing both mechanical and electrical systems. For measurements involving modal analysis, machinery diagnostics, vibration testing, servo systems, control systems, or acoustics, the SR785 has the features and specifications to get the job done.

Standard measurement groups include FFT, order tracking, octave, swept-sine, correlation, time capture and time/histogram. The SR785 brings the power of several instruments to your application: a spectrum analyzer, network analyzer, vibration analyzer, octave analyzer and oscilloscope.

A unique measurement architecture allows the SR785 to function as a typical dual channel analyzer with measurements like cross spectrum, frequency response, coherence, etc. Alternatively, the instrument can be configured so that each input channel functions as a single channel analyzer with its own span, center frequency, resolution, and averaging. This allows you to view a wideband spectrum and simultaneously zoom in on spectral details. The same advanced architecture provides storage of all measurement building blocks and averaging modes. Vector averaged, RMS averaged, unaveraged and peak hold versions of all measurements are simultaneously acquired and can be displayed without re-taking data.

Averaging

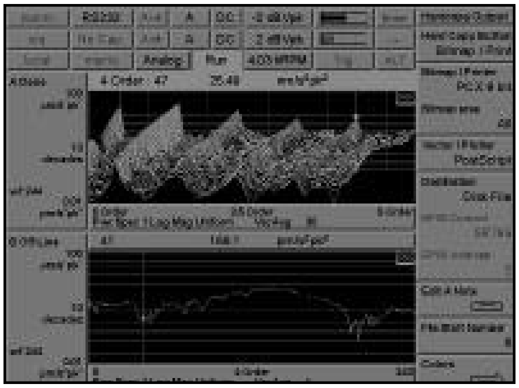
The SR785 comes equipped with a wide selection of averaging techniques to improve your signal to noise ratio. Choose RMS averaging to reduce signal fluctuations, vector averaging to actually eliminate noise from synchronous signals, or peak hold averaging. In the order-tracking measurement group, time averaging is available. Both linear and exponential averaging are provided for each mode.

Because the SR785 is so fast, there's no need for a separate "fast averaging" mode. For instance, in a full span FFT measurement with a 4 ms time record, 1000 averages take exactly 4 seconds, during which the SR785 still operates at its maximum display rate.

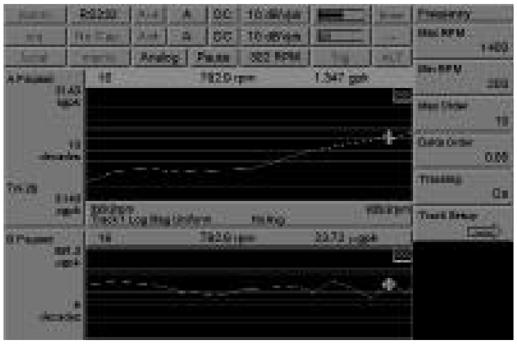
For impact testing, the average preview feature allows each time record or spectrum to be accepted or rejected before adding it to the measurement.

Order Tracking

Order tracking is used to evaluate the behavior of rotating machinery. Measurement data is displayed as a function of multiples of the shaft frequency (orders), rather than absolute frequency. Combined with a waterfall plot, the SR785 provides a complete history or "order map" of your data as a function of time or RPM. Using the slice feature, the amplitude profile of specific orders in the map can be analyzed.



Order Maps



Tracked Order Display

In tracked order mode the intensity of individual orders vs. RPM is measured. Unlike other analyzers, there's no need to track a limited number of orders to ensure full speed measurements. The SR785's speed allows simultaneous tracking of up to 400 orders.

Run-up and run-down measurements are available in both polar and magnitude/phase formats. RPM profiling is provided to monitor variations of RPM as a function of time. A complete selection of time and RPM triggering modes is included allowing you to make virtually any rotating machinery measurement.

Octave Analysis

Real-time 1/1, 1/3, 1/12 octave analysis at frequencies up to 40 kHz (single channel), or 20 kHz (2 channel) is a standard feature of the SR785. Octave analysis is fully compliant with ANSI S1.11-1986 (Order 3, type 1-D) and IEC 225-1966. Switchable analog A-weighting filters as well as A, B and C weighting math functions are included. Averaging choices include exponential time averaging, linear time averaging, peak hold and equal confidence averaging. Broadband sound level is measured and displayed as the last band in the octave graph. Total power, impulse, peak hold and Leq are all available. Exponentially averaged sound power (Leq) is calculated according to ANSI S1.4-1983, Type 0.

Octave displays can be plotted as waterfalls with a fast 4 ms storage interval. Once data is stored in the waterfall buffer, the SR785 can display centile exceedance statistics for each 1/1, 1/3 or 1/12 octave band as well as for Leq.

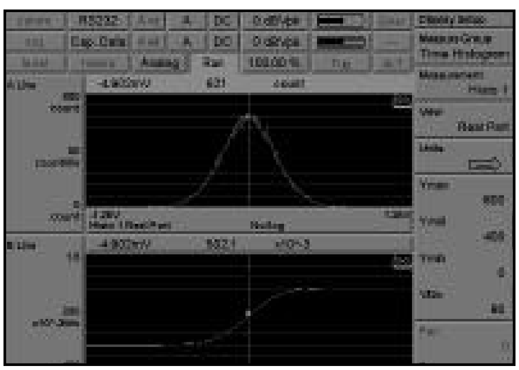
Swept-Sine Measurements

Swept-sine mode is ideal for signal analysis that involves high dynamic range or wide frequency spans. Gain is optimized at each point in the measurement producing up to 145 dB of dynamic range. A frequency resolution of up to 2000 points is also provided. Auto-ranging can be used with source auto-leveling to maintain a constant input or output level at the device under test (to test response at a specific amplitude, for instance).

Auto-resolution ensures the fastest possible sweeps and adjusts the frequency steps in the scan based on the DUT's response. Phase and amplitude changes that exceed user-defined thresholds are measured with high frequency resolution, while small changes are measured using wider frequency steps between points. A choice of linear sweeps with high resolution, or logarithmic sweeps with up to eight decades of frequency range is provided.

Time/Histogram

Use the time/histogram measurement group to analyze time domain data. A histogram of the time data vs. signal amplitude is provided for accurate time domain signal characterization. Statistical analysis capabilities include both probability density function (PDF) and cumulative density function (CDF). The sample rate, number of samples and number of bins can all be adjusted.



Histogram and CDF

Time Capture

The SR785 comes with 2 Msamples of memory (8 Msamples optional). Analog waveforms can be captured at sampling rates of 262 kHz or any binary submultiple, allowing you to optimize sampling rate and storage for any application. For example, 8 Msamples of memory will capture 32 seconds of time domain data at the maximum 262 kHz sample rate, or about 9 hours of data at a 256 Hz sample rate. Once captured, any portion of the signal can be played back in any of the SR785's measurement groups except swept-sine. The convenient Auto Pan feature lets you display measurement results synchronously with the corresponding portion of the capture buffer to identify important features.

Unit Conversion

Automatic unit conversion makes translating transducer data easy. Enter your transducer conversion directly in V/EU, EU/V or dB (1V/EU). The SR785 will display the result in units of meters, inches, m/sec^2, in/sec^2, m/s, in/s, mil, g, kg, lbs., N, dynes, pascals or bars. Built-in ICP power means you don't need an external power supply for your accelerometer. Acoustic measurement results can be displayed in dBSPL, while electrical units include V, V^2, dBV and dBm.

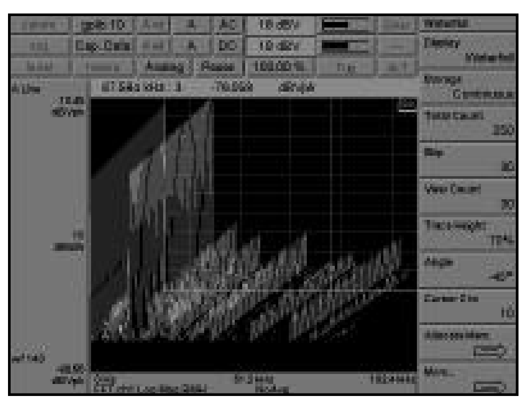
Source

The SR785 comes standard with five precision source types. Generate low distortion (-80 dBc) single or two-tone sine waves, white noise, pink noise, chirps, and arbitrary waveforms. The chirp and noise sources can be bursted to provide activity over a selected portion of the time record for FFT measurements, or to provide impulse noise for acoustic measurements. The digitally synthesized source produces output levels from 0.1 mV to 5 V and offsets from 0 to ±5 V, and delivers up to 100 mA of current.

Arbitrary waveform capability is standard with the SR785. Use the arbitrary source to playback a section of a captured waveform, play a selected FFT time record or upload your own custom waveform from your computer.

User Math

Create your own measurement in each of the SR785's measurement groups using the math menu. Enter any equation involving RMS averaged, vector averaged or unaveraged time or frequency data, stored files, constants, or a rich array of supplied operations including arithmetic functions, FFT, inverse FFT, jw, d/dw, exp, ln x, and many others. Because all the averaging modes are available as user math operands, non-repeatable



Waterfall Plots

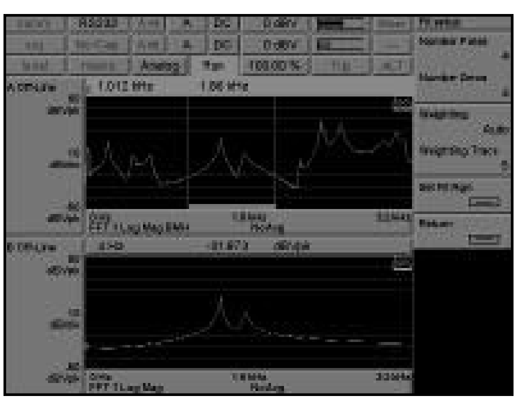
runout measurements (used in analyzing disk drives) can be performed in a single pass by entering the equation MAG(RMS<F1>)-MAG(Vec<F1>). Unlike many other analyzers, the SR785's measurement rate is virtually unaffected when user math is selected. For instance, the function exp(ln(conj(FFT2/FFT1))) can be calculated with a 100 kHz real-time bandwidth.

Waterfall

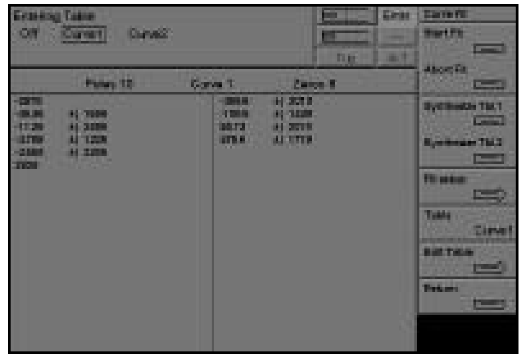
Waterfall plots are a convenient way of viewing a time history of your data. Each successive measurement record is plotted along the z-axis making it easy to see trends in the data. All FFT, octave and order tracking measurements can be stored in the SR785's waterfall buffer memory. You can choose to save all measurements and averaging modes or just the current measurement to conserve memory. Waterfall traces can be stored every n time records for FFT and order tracking measurements. For order tracking measurements new records can be acquired at a specific time interval or change in RPM. In octave measurements, the storage interval is in seconds (as fast as every 4 ms). While displaying waterfall plots, you can adjust the skew angle to reveal important features, or change the baseline threshold to eliminate low-level clutter. Any z-axis slice or x-axis record can be saved to disk or displayed separately for analysis.

Analysis

The SR785 includes a wide variety of built-in analysis features. Marker analysis lets you easily measure the power contained in harmonics, sidebands or within a given band of frequencies. Important information such as THD, THD+N, sideband power relative to a carrier, or total integrated power are calculated in real-time and displayed on the screen. The front/back display feature allows you to display live data from both signal inputs



Original Data with Fit



Curve Fit Tables

on one graph. You can also simultaneously display saved traces and live data. The peak-find marker allows you to quickly locate frequency peaks with the click of a button. The marker statistics feature calculates the maximum, minimum, mean and standard deviation of data in any section of the display. For modal analysis, the cursor can be configured to display the resonant frequency and damping of a single selected mode.

Use data tables to display up to 100 selected data points in tabular format. Limit tables allow you to define up to 100 upper and lower limit segments in each display for GO/NO GO testing. Data and limit table definitions can also be saved to and recalled from disk for quick setup.

Curve Fit and Synthesis

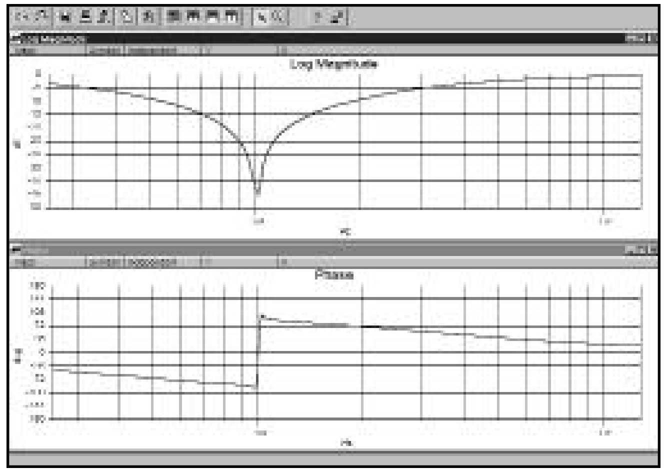
The SR785 has a standard 20-pole, 20-zero curve fitter that can fit frequency domain data from both the FFT and swept-sine measurement groups. Curve models can be displayed in pole/zero, pole/residue and polynomial formats. Synthesis reverses the process: enter a model in any of the above formats and the SR785 synthesizes the required curve. The curve fit/synthesis menu allows you to change gain, delay and frequency scale, set pole and zero locations and instantly see the response of the modeled system.

Output

The SR785's built-in 3.5" 1.44 MB disk drive, GPIB and RS-232 computer interfaces and Centronics printer port combine to allow almost unlimited flexibility in saving, printing, plotting or exporting your measurement data. Displays and setups can be printed or plotted to disk, GPIB, RS-232 or Centronics port in PCX, GIF, PCL (LaserJet® and DeskJet®), dot-matrix, Postscript, HP-GL or ASCII formats. An annotation editor lets you add text, time and date, or filenames to any portion of the plot. For modal analysis applications, nodal degree-of-freedom information is stored with the data files to simplify transfer to external analysis programs.

Data Conversion Utilities

The SR785 contains a complete suite of data conversion utilities for both Windows® and DOS® operating systems. These versatile utilities make data transfer to and from external programs fast and simple. SR785 files can be converted to ASCII for use with spreadsheets, or Universal File Format (UFF) and HP SDF for use with modal analysis programs. SR785 Files can also be converted to MAT file format for use with MATLAB. Conversion from external file types is also supported. Both HP SDF and SR780 files can be converted to SR785 format.



DataViewer Program

DataViewer

The SRS DataViewer is a Windows 95 program that allows you to quickly upload SR785 files into a graphical environment, perform simple editing and cut and paste into other applications. You can add pointers and text, change scaling, and perform simple math operations. Graphs can be saved in PCX, BMP or GIF formats.

SR785 Rear Panel

SR785 Features

Instrument Modes
FFT, Time/Histogram, Correlation, Octave, Swept-Sine, Order Tracking

Frequency Domain Measurements
Frequency Response, Linear Spectrum, Cross Spectrum, Power Spectrum, Coherence, Power Spectral Density

Time Domain Measurements
Time Record, Cross Correlation, Auto Correlation, Orbit

Amplitude Domain Measurements
Histogram, PDF, CDF

FFT Resolution
100, 200, 400, 800 lines

Views
Linear Magnitude, Log Magnitude, dB Magnitude, Magnitude Squared, Real Part, Imaginary Part, Phase, Unwrapped Phase, Nichols, Nyquist, Polar

Units
V, V^2, V^2/Hz, V/√Hz, meters, meters/sec, meters/sec^2, inches, inches/sec, inches/sec^2, mils, g, kg, lbs., N, dynes, pascals, bars, SPL, user-defined engineering units (EUs)

Displays
Single, Dual, Front/Back overlay, Waterfall with Skew, Zoom and Pan, Grid On/Off

Marker Functions
Trace Marker, Dual Trace Linked Marker, Absolute and Relative Marker, Peak Find, Harmonic Marker, Band and Sideband Marker, Waterfall Marker, Frequency-Damping Marker

Averaging
RMS, Vector, Peak Hold, Linear, Exponential, Equal Confidence (Octave), Preview Time Record, % Overlap Averaging, Overload Reject

Triggering
Continuous, Internal, External (Analog or TTL), Source, Auto/Manual Arming, GPIB, RPM Step, Time Step, Pre/Post Trigger Delay

Application Information

Application Note #1, "FFT Spectrum Analyzers," contains a complete discussion of the fundamentals of FFT spectrum analysis and an explanation of many basic terms used when discussing signal analyzers.

Source Outputs
Sine, Two-Tone, Swept-Sine, White/Pink Noise, Burst Noise, Chirp, Burst Chirp, and Arbitrary

Windows
Hanning, Blackman-Harris, Flat Top, Kaiser, Force/Exponential, User-Defined, +/-T/2, +/-T/4, T/2, Uniform

User Math
+, -, *, /, Conjugate, Magnitude/Phase, Real/Imaginary, Sqrt, FFT, Inverse FFT, jw, Log, Exp, d/dx, Group Delay, A, B, C Weighting, x/x-1, Trace 1-5, Vector Average, RMS Average, Peak Hold

Analysis
Harmonic, Band, Sideband, THD, THD + N, Limit Test with Pass/Fail, Data Table, Exceedance, Statistics, Curve Fit and Synthesis

Time Capture
Capture Time Data for later analysis (FFT or Octave). Up to 2 Msamples (8 Msamples optional) of data can be saved.

Storage
3.5", 1.44 Mbyte, DOS formatted disk. Save data, setups and hard copy data.

Hard Copy and Interfaces
Print to Dot Matrix or PCL (LaserJet and DeskJet) printers. Plot to HP-GL or Postscript plotters. Print/Plot on-line (RS-232 serial, Centronics parallel or IEEE-488) or to disk file. EPS, GIF, PCX graphic formats also available for disk storage.

Data Conversion Utility
Data, waterfall and capture files can be converted to ASCII. Data files can also be converted to Universal File Format, SDF format or MATLAB MAT-File Format. SDF and SR780 files can be converted to SR785 format.

DataViewer
Windows 95 based graphics program for viewing SR785 files. Graphs can be pasted to the clipboard or saved in PCX, BMP or GIF formats.

Ordering Information

SR785	**Dynamic Signal Analyzer**	**$10,950**
O780RM	**Rackmount Kit**	**$85**
O780M1	**8 Msample RAM**	**$800**
O78012V	**12 VDC-115 VAC Converter**	**$295**
CT100	**Mobile Instrument Cart**	**$495**
O780UG	**Upgrade SR780 to SR785**	**$2500**

SR785 Specifications

Specifications apply after 30 minutes of warm-up and within two hours of last auto-offset. All specifications are with 400 line FFT resolution and anti-alias filters enabled unless stated otherwise.

Measurement Groups

Groups	FFT analysis, Correlation, Time Histogram, Swept Sine, Order Tracking

Frequency

Range	102.4 kHz or 100 kHz (both displays have the same range).
FFT Spans	195.3 mHz to 102.4 kHz or 191 mHz to 100 kHz. The two displays can have different spans and start frequencies.
FFT Resolution	100, 200, 400 or 800 lines
Real-Time Bandwidth	102.4 kHz (highest FFT span with continuous data acquisition and averaging).
Accuracy	25 ppm from 20 °C to 40°C

Dynamic Range

Dynamic Range	-90 dBfs typical, -80 dBfs guaranteed (FFT and Octave) -145 dBfs typical (Swept-Sine) Includes spurs, harmonics, intermodulation distortion and alias products. Excludes alias responses at extremes of span.
Harmonic Distortion	<-80 dB (single tone in band)
Intermod. Distortion	<-80 dB (two tones in band, each less than -6.02 dBfs)
Spurious	<-80 dBfs
Alias Responses	<-80 dBfs (single tone outside of span, less than 0 dBfs, less than 1 MHz)
Full Span FFT Noise Floor	100 dBfs typical (input grounded, input range >-30 dBV, Hanning window, 64 RMS averages)
Residual DC Response	<-30 dBfs (FFT with Auto Cal on)

Amplitude Accuracy

Single Channel	±0.2 dB (excluding window effects)
Cross Channel	±0.05 dB (dc to 102.4 kHz) (frequency response measurement, both inputs on the same input range, RMS averaged)

Phase Accuracy

Single Channel	±3.0 deg relative to External TTL trigger (-50 dBfs to 0 dBfs, freq <10.24 kHz, center of frequency bin, DC coupled). For Blackman-Harris, Hanning, Flattop and Kaiser windows, phase is relative to a cosine wave at the center of the time record. For Uniform, Force and Exponential windows, phase is relative to a cosine wave at the beginning of the time record.
Cross Channel	±0.5 deg (dc to 51.2 kHz) ±1.0 deg (dc to 102.4 kHz) (frequency response measurement, both inputs on the same input range, vector averaged)

Signal Inputs

Number of Inputs	2
Full Scale Input Range	-50 dBV (3.16 mVpk) to +34 dBV (50 Vpk) in 2 dB steps
Maximum Input Level	57 Vpk
Input Configuration	Single-ended (A) or true differential (A-B)
Input Impedance	1 MΩ + 50 pF
Shield to Chassis	Floating mode: 1 MΩ + 0.01 mF Grounded mode: 50 Ω Shields are always grounded in differential input (A-B)
Maximum Shield Voltage	4 Vpk
AC Coupling	-3 dB rolloff at 0.16 Hz
CMRR	90 dB at 1 kHz (in. range <0 dBV) 80 dB at 1 kHz (in. range <10 dBV) 50 dB at 1 kHz (in. range ≥10 dBV)
ICP Signal Conditioning	Current Source: 4.8 mA Open Circuit Voltage: +26 V
A-Weight Filter	Type 0 Tolerance, ANSI Standard S1.4-1983; 10 Hz to 25.6 kHz
Crosstalk	<-145 dB below signal (input to input and source to inputs, 50Ω receiving input source impedance)
Input Noise	<10 nVrms/√Hz (<-160 dBVrms/√Hz) above 200 Hz

Trigger Input

Modes	Free run, Internal, External, or External TTL
Internal	Level adjustable to ±100% of input scale, positive or negative slope. Minimum Trigger Amplitude: 5% of input range
External	Level adjustable to ±5 V in 40 mV steps, positive or negative slope. Input impedance: 1 MΩ Max input: ±5 V Minimum trigger amplitude: 100 mV
External TTL	Requires TTL level to trigger (low <0.7 V, high >3.0 V)
Post-Trigger	Measurement record is delayed up to 100,000 samples after the trigger.
Pre-Trigger	Measurement record starts up to 8000 samples prior to the trigger.

Tachometer Input

Pulses Per Revolution	1 to 2048
RPM Accuracy	±50 ppm (typical)
Tach Level Range	±25 V, ±5V, TTL
Tach Level Resolution	20 mV @ ±25 V, 4 mV @ ±5 V
Max. Tach Input Level	±40 Vpk
Min. Tach Pulse Width	100 nSec
Max. Tach Pulse Rate	750 kHz (typical)

Transient Capture

Mode	Continuous real-time data recording to memory.
Maximum Rate	262,144 samples/sec for both inputs
Maximum Capture Length	2 Msamples (single input) 8 Msamples with optional memory

Octave Analysis

Standards	Conforms to ANSI standard S1.11-1986, Order 3, Type 1-D.	
Frequency Range	Band centers:	
	Single Channel	
	1/1 Octave	0.125 Hz - 32 kHz
	1/3 Octaves	0.100 Hz - 40 kHz
	1/12 Octaves	0.091 Hz - 12.34 kHz
	Two Channels	
	1/1 Octave	0.125 Hz - 16 kHz
	1/3 Octaves	0.100 Hz - 20 kHz
	1/12 Octaves	0.091 Hz - 6.17 kHz
Accuracy	< 0.2 dB (1 second stable average, single tone at band center)	
Dynamic Range	80 dB (1/3 Octave, 2 second stable average) per ANSI S1.11-1986	
Sound Level	Impulse, Peak, Fast, Slow and Leq per IEC 651-1979 Type 0	

Order Tracking

Delta Order	.0075 to 1
Resolution	up to 400 lines
Amplitude accuracy	±1 dB (typ.)
Displays	Order map (mag and phase), order track (mag and phase), orbit

Curve Fit and Synthesis

Type:	20 poles/20 zeros curve non-iterative rational fraction fit.
Order Selection:	Auto or manual
Output format:	Pole-zero, polynomial, or pole residue

Source Output

Amplitude Range	0.1 mVpk to 5 Vpk
Amplitude Resolution	0.1 mVpk
DC Offset:	<10.0 mV (typical)
Output Impedance	< 5Ω, ±100 mA peak output current.

Sine-Source

Amplitude Accuracy	±1% of setting, 0 Hz to 102.4 kHz 0.1 Vpk to 5.0 Vpk, high impedance load.
Harmonics, SubHarm.	0.1 Vpk to 5 Vpk
and Spurious Signals	<-80 dBc (fundamental < 30 kHz) <-75 dBc (fundamental < 102 kHz)

Two Tone Source

Amplitude Accuracy	±1% of setting, 0 Hz to 102.4 kHz 0.1 Vpk to 5 Vpk, high impedance load.
Harmonics, SubHarm.	< -80 dBc, 0.1 Vpk to 2.5 Vpk

White Noise Source

Time Record	Continuous or Burst
Bandwidth	DC to 102.4 kHz or limited to analysis span.
Flatness	<0.25 dB pk-pk (typical), <1.0 dB pk-pk (max), 5000 rms averages

Pink Noise Source

Bandwidth	DC to 102.4 kHz
Flatness	<2.0 dB pk-pk, 20 Hz - 20 kHz (measured using averaged 1/3 Octave Analysis)

Chirp Source

Time Record	Continuous or Burst
Output	Sine sweep across the FFT span.
Flatness	±0.25 dB, Amplitude = 1.0 Vpk

Swept Sine Source

Auto Functions	Source Level, Input Range and Frequency Resolution
Dynamic Range	145 dB

Arbitrary Source

Amplitude Range	± 5V
Record Length	2 Msamples (playback from Arbitrary Waveform memory or capture buffer). Variable output sample rate.

General

Monitor	Monochrome CRT, 800H by 600V resolution.
Interfaces	IEEE-488, RS232 and Printer interfaces standard. All instrument functions can be controlled through the IEEE-488 and RS232 interfaces. A PC (XT) keyboard input is provided for additional flexibility.
Hardcopy	Print to dot matrix and PCL compatible printers. Plot to HPGL or Postscript plotters. Print/Plot to RS232 or IEEE-488 interfaces or to disk file. Additional file formats include GIF, PCX and EPS.
Disk	3.5 inch DOS compatible format, 1.44 Mbytes capacity. Storage of displays, setups and hardcopy.
Preamp Power	Power connector for SRS preamplifiers.
Power	70 Watts, 100/120/220/240 VAC, 50/60 Hz.
Dimensions	17”W x 8.25”H x 24”D
Weight	56 lbs.
Warranty	One year parts and labor on materials and workmanship.

FFT Signal Analyzer

Model SR780 — 100 kHz Two Channel FFT Network Signal Analyzer

- ***DC to 102.4 kHz Bandwidth***
- ***90 dB Dynamic Range***
- ***16-bit A/D Conversion***
- ***Low Distortion Synthesized Source***
- ***ANSI Standard Octave Measurements***
- ***145 dB Dynamic Range Swept Sine Mode***
- ***2 Msamples Transient Capture Memory***
- ***Internal and External Triggering***
- ***Built-in Hypertext Help***
- ***Graphics Outptut to GIF, EPS, PCX, HPGL***
- ***3.5" MS-DOS Compatible Disk Drive***
- ***GPIB and RS-232 Interfaces Included***

SR780 Overview

The SR780 is a high performance dual-channel FFT analyzer with a dynamic range of 90 dB and a real-time bandwidth of 102.4 kHz. Unlike analyzers that force you to buy an array of extensive options, the SR780 comes standard with swept-sine measurement capability, ANSI standard Octave Analysis, 2 Msamples of transient capture memory, and a low-distortion synthesized source with arbitrary waveform capability.

High Speed

The SR780 uses a state-of-the-art high-speed DSP processor capable of 33,000,000 floating point multiplications per second. This gives the SR780 its high 102.4kHz bandwidth, even when both channels are used. And the floating point DSP ensures that dynamic range and accuracy are never sacrificed even in the most demanding situations, such as the measurement of very small signals in the presence of strong interference.

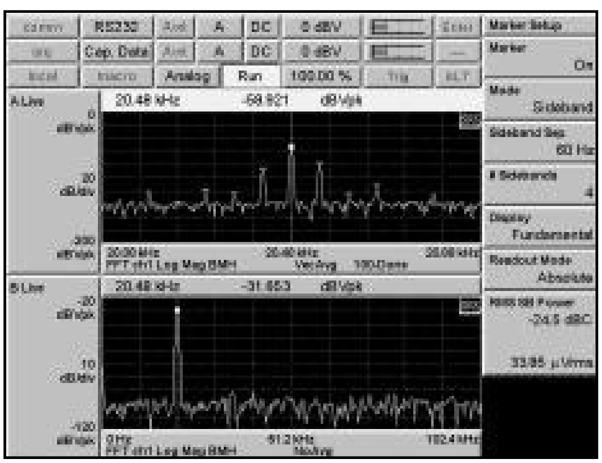
Whether your application involves acoustic measurements, vibration testing, control system analysis, or filter design, the SR780's combination of features, performance, and low cost make it the overwhelming choice in FFT signal analyzers.

SR780 Features

Spectrum Analysis

The SR780 delivers true 2-channel 102.4 kHz FFT performance. Unlike other analyzers, the SR780 doesn’t make you sacrifice 2-channel performance for bandwidth – its fast 32-bit floating point DSP processor gives the SR780 a 102.4 kHz realtime rate with both channels selected. Two precision 16-bit ADCs provide a 90 dB typical dynamic range in FFT mode – enough for the most demanding low-level measurements. Selectable 100 - 800 line analysis optimizes time and frequency resolution and any portion of the 102.4 kHz range can be analyzed with frequency spans down to 191 mHz. Even the sampling rate can be set to 256 kHz or 262 kHz, so frequency spans come out in either a binary (102.4 KHz, 51.2 kHz,...) or decimal (100 kHz, 50 kHz, 25 kHz,...) sequence depending on your requirements.

The SR780’s unique architecture each display function as separate analyzer. Separate frequency spans, starting frequencies, number of FFT lines, and averaging modes can be selected for each display. So it’s no problem to look at a wideband display and zoom in on a specific feature simultaneously.



Simultaneous Dual-Span Analysis of a Signal

Flexible Averaging

A complete set of averaging choices matches any application Choose rms averaging to reduce signal fluctuations, or vector averaging to actually eliminate noise from synchronous signals. Choose linear averaging (stable averaging) for fixed signals, or exponential averaging to track drifting features. Because the SR780’s 102.4 kHz realtime bandwidth lets it take data seamlessly, vector averaging can be selected for any signal that’s repetitive within the time record – no trigger is necessary.

Automatic Unit Conversion

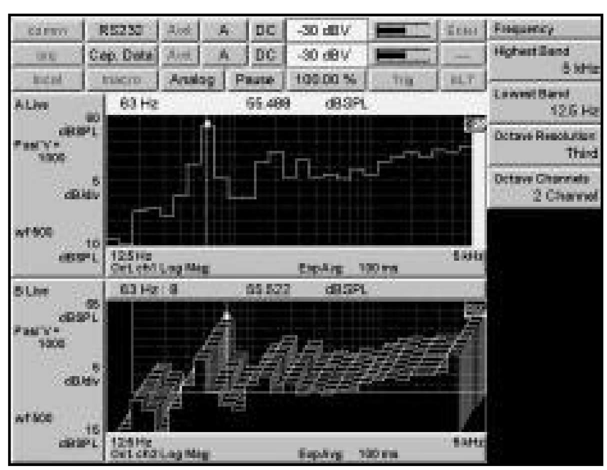
Automatic unit conversion makes translating accelerometer data easy. Enter accelerometer conversion factors directly in V/EU, EU/V or dB(1V/EU). The SR780 will display results in units of meters, inches, mil, g, kg, lbs, N, dynes, pascals, bars, or dBSPL. Accelerometer data is automatically converted to velocity or displacement units. A standard built-in ICP power means an external acceelrometer power supply isn’t necessary.

Octave Analysis

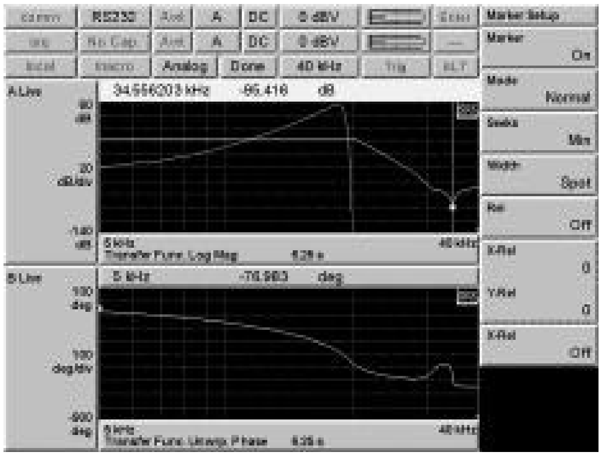
Realtime octave analysis at frequencies up to 40 kHz (single channel), or 20 kHz (2-channel) is standard on the SR780. Octave analysis is fully compliant with ANSI S1.11-1986 (Order 3, type 1-D) and IEC 225-1966. 1/1 octave, 1/3 octave, and 1/12 octave analysis are all available. Switchable analog A-weighting filters as well as built-in user math A, B, and C-weighting functions are all included. Octave averaging choices include exponential time averaging, linear time averaging, peak hold, and equal confidence averaging. IEC 651 Type 0 compliant peak hold, impulse, fast, and slow sound level measurements are all calculated.

Swept-Sine Analysis

Swept-sine analysis for measurements involving high dynamic range or wide frequency intervals is also a standard feature of the SR780. Selectable auto-ranging optimizes the input range at each point in the measurement, providing up to 145 dB of dynamic range. Auto-ranging can be used with source auto-leveling to maintain a constant input or output level at the device under test. To ensure the fastest sweeps possible, auto-resolution can also be selected, providing a variable scan speed tailored precisely to the signal being measured. Choose linear sweeps for high frequency resolution or



Standard 1/1, 1/3 and 1/12 Octave Analysis



Swept-sine Bode Plot of Filter Response

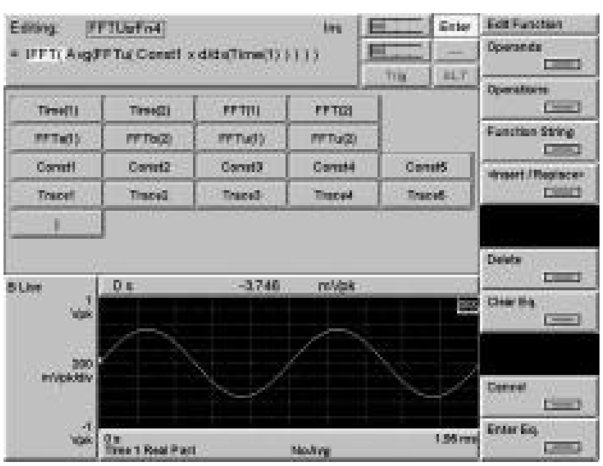
logarithmic sweeps up to 8 decades for the widest frequency coverage.

User Math

All three measurement groups: FFT, octave, and swept-sine, allow you to create your own measurement using the SR780's user math menu. Enter any equation involving time or frequency data, stored files, constants, or a rich array of supplied operations including the arithmetic functions, FFT, inverse FFT, jω, d/dω, exp, ln x, x/(1-x), conjugate, square root, group delay, and many others. Unlike many analyzers, the SR780 maintains its high speed even when user math functions are selected. For instance, the function exp(ln(conj(Average(FFT2/FFT1))) can be calculated with a 50 kHz realtime bandwidth.

Source

The SR780 includes 5 precision source types: low distortion (-80 dBc) single or two-tone sine waves, chirps, white noise, pink noise, or arbitrary waveforms. The chirp and noise sources can both be bursted to provide a source that's active only over a selected portion of the time record for FFT measurements, or to provide an impulsive noise source for acoustic measurements. The digitally synthesized source provides output levels from .1 mV to 5 V and delivers up to 100 mA of current.



User Math Function Entry

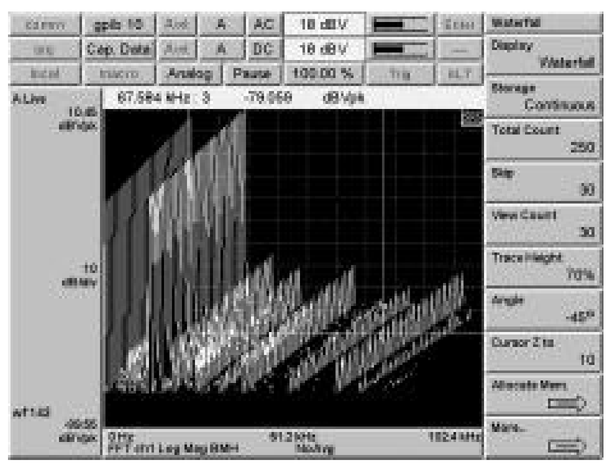
Arbitrary waveform capability is standard on the SR780. Use the arbitrary source to playback a section of a captured waveform, play a selected FFT time record, or upload your own custom waveform from disk or over the remote interface.

Transient Capture

The SR780 comes with 2 Msamples of standard capture memory. Waveforms can be captured at any submultiple of the selected sampling rate to optimize the capture time and sampling rate for your application. Once captured, any portion of the signal can be played back in any FFT or octave measurement. The convenient "AutoPan" feature allows the measurement results to be displayed synchronously with the corresponding portion of the capture buffer to easily identify important features. Optional memory expansion modules lets you expand the SR780's capture depth to up to 8 Msamples – that's almost 30 seconds of capture at the maximum 262 kHz sampling rate.

Waterfall

All Octave and FFT Measurements can be stored in the SR780's two 2k deep waterfall buffers. Waterfall storage is selectable as every nth time record for FFT measurements, or you can select a storage interval in seconds (down to 4 ms) for octave measurements. While displaying waterfalls both the skew angle and baseline



Waterfall Display

threshold are adjustable, allowing you to reveal important features, and eliminate low-level clutter. Any z-axis slice or x-axis record can be saved to disk, displayed separately for individual analysis, or included in a user defined math function.

Analysis

The SR780 includes a wide variety of built-in analysis features. Marker analysis lets you use the marker to easily measure the power contained in the harmonics, sidebands or within a given band of a frequency domain measurement. Important signal information such as THD, THD+N, sideband power relative to carrier, and total integrated power are calculated in realtime and displayed on the screen. Marker statistics quickly calculate the maximum, minimum, mean, and standard deviation of data at any point in the display.

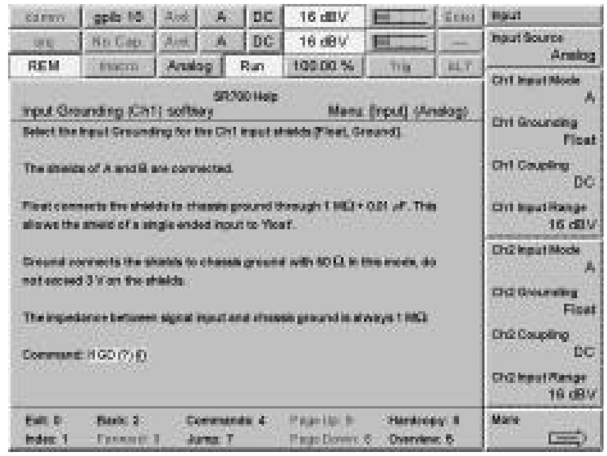
Use data tables to display up to 100 selected data points in a tablular format. Limit tables allow you to define up to 100 upper and lower limit segments in each display for GO/NO GO testing. Data and limit table definitions can also be saved and recalled from disk for quick setup.

Output

The SR780's 3.5" 1.44 MB floppy disk drive, IEEE-488 and RS232 interface ports and Centronics printer port combine to allow almost unlimited flexibility in saving, printing, plotting or exporting your measurement data. Annotated displays can be printed or plotted to the disk, IEEE-488, RS232, or Centronics port in PCX, GIF, PCL (HP LaserJet and DeskJet), Dot-matrix, Postscript, HP-GL or ASCII formats. User generated ASCII files can also be imported with a single keystroke allowing you to create your own displays for use in math functions or to compare with live data. Utilities are even included to translate your HP SDF files into SR780 format.

Ordering Information

SR780	**FFT Network Analyzer**	**$9950**
O780RM	**Rackmount Kit**	**$85**
O780M1	**8 Msample RAM**	**$800**
O78012V	**12 VDC-115 VAC Converter**	**$295**
CT100	**Mobile Instrument Cart**	**$495**



Extensive On-line Hypertext Help

Help

Full, context-sensitive help screens for all SR780 features mean you will rarely have to refer to a printed manual. Hypertext links let you quickly switch between related help pages or instantly reference the remote command corresponding to any SR780 function. Use the help index to quickly locate help on any topic, jump to the online troubleshooting guide, browse a complete listing of the SR780's specifications, or examine a comprehensive description of the SR780's remote commands—all from the front panel.

Application Information

Application Note #1, "FFT Spectrum Analyzers," contains a complete discussion of the fundamentals of FFT spectrum analysis and an explanation of many basic terms used when discussing signal analyzers.

Getting High Dynamic Range with FFT Spectrum Analyzers

To get the maximum dynamic range possible from FFT spectrum analyzers, a thorough understanding of the nature of the signal, and the measurement techniques used by these instruments is essential. To illustrate some of these ideas we'll use the SR780 to measure the transfer function of a lowpass filter. The filter we've chosen has a stopband edge at 20 kHz, a stopband attenuation of ~-80dB and a transmission zero at about 33 kHz with a depth of over 100 dB. The problem is to accurately measure the entire transfer function of the filter, including the 0 dB passband and the structure in the stopband near -100 dB. Our first approach might be to use a periodic chirp, a signal with a flat spectrum, as the input to the filter, and to use the SR780 to measure the transfer function, the ratio of the output spectrum to the input spectrum. The result of such a measurement is shown in figure 1, and the limitations of this simple-minded measurement technique are immediately apparent.

Notice that the stopband of the filter shows up as just a band of noise around 70 dB down from the passband. Why only 70 dB of dynamic range when the noise floor of the SR780 should be about -100 dBV? The answer lies in the nature of our input signal. If we use a periodic chirp with a peak amplitude in the time domain of 1V, we've got to set the analyzer's input range to 0 dBV to avoid overloading the analyzer's front end. However this signal has only about -30dBV of amplitude in each of the FFT's 512 lines. Since we're exciting the filter with only about -30 dBV at each frequency, -100 dBV noise floor shows up only 70 dB below the filter passband.

Averaging, of course, can increase the dynamic range of this kind of measurement. Averaging generally comes in two flavors in FFT signal analyzers: RMS, and vector. RMS averaging simply averages the power found at each point in the spectrum, including the noise power. The result of our measurement after RMS averaging 300 transfer functions is shown in figure 2.

Note that we haven't increased the dynamic range of the result at all. The band of noise at -70 dB is cleaned up considerably, but there's still no hint that that any structure exists in the stopband of our filter transfer function. To see that kind of structure we need an averaging technique that reduces noise, not just smoothes it. This type of averaging is vector averaging. Vector averaging is a synchronous averaging technique that averages the complex value of the FFT at each frequency point. Frequency components that are periodic will reinforce each other on successive averages while noise, which has no fixed phase relationship from one record to the next, will cancel and average towards zero. Figure 3 shows our transfer function measurement after 300 spectra have been vector averaged. Notice how structure in the stopband is beginning to emerge(the transmission zero at 34.5 kHz shows up at -90 dB and the noise floor is about 10 dB below the RMS averaged case.

It's important to realize that for vector averaging to work the phase of the signal has to remain constant from one record to the next. For a signal that is repetitive within the analyzer's time record this means that the analyzer has to be fast enough so that each FFT record represents contiguous data. If the analyzer can't keep up, it will begin each time record at a random point relative to the signal and the signal will appear to have random phase and average to zero. In this case, vector averaging can only be done while using a hardware trigger to guarantee a fixed phase between the FFT time record and the input signal. However, newer FFT analyzers, such as the SR780, are fast enough to keep up with the input data even while taking two FFTs at a100 kHz span. In this case, triggerless vector averaging is possible (for frequencies that are repetitive within a time

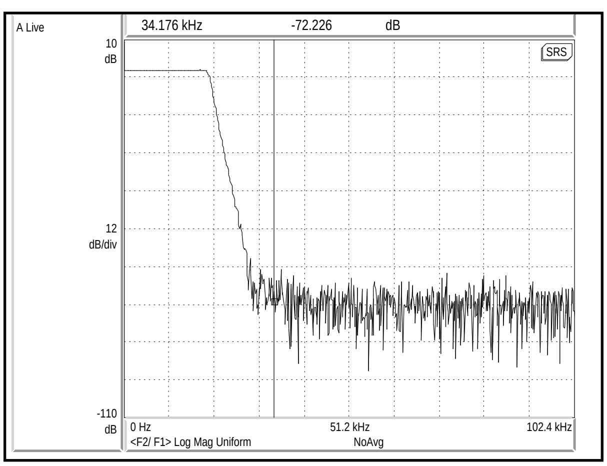


Fig. 1: Unaveraged Transfer Function

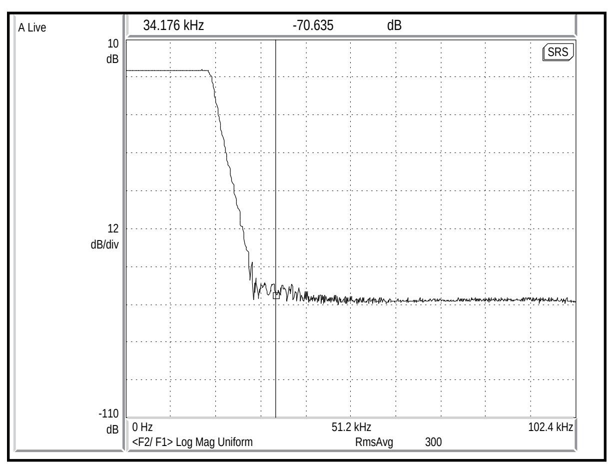


Fig. 2: RMS Averaged Transfer Function

record) since the speed of the FFT analyzer guarantees that each time record will be contiguous with the last.

None of the averaging techniques discussed above however, address the root of the problem with the FFT measurement technique: that an input signal large enough to resolve the stopband will overload the instrument in the passband. It's clear that in this case the FFT's great strength (that it measures all frequencies at once), is a weakness. What we need is a measurement technique that allows us to turn up the instrument's gain in the stopband, where the input signal is small, while leaving it low in the passband to avoid overloads. The swept-sine measurement mode is precisely the tool needed.

In a swept sine measurement the input signal to the device under test is a sine wave whose frequency is swept over the frequency range of interest. Since the source is a sine wave, all of its energy is concentrated at a single frequency, eliminating the 30 dB dynamic range penalty we paid with the periodic chirp signal. In addition because the swept sine technique only looks at one frequency at a time, the input gain of the analyzer can be adjusted at each frequency to provide the optimum input range for that portion of the device's transfer function. Figure 4 shows a swept-sine spectrum of the same filter. The instrument was setup to sweep 400 points from 100 Hz to 102.4 kHz. Each point was averaged for the longer of 4 ms or 10 cycles.

Notice how the zero is resolved cleanly to a depth of -116 dB and the shape of the stopband floor is clearly shown. During this sweep the SR780's input gain autoranged from 0 dBV while the source was sweeping through the passband to -50 dBV at the zero frequency. Of course the added dynamic range of this measurement did not come for free, the sweep took 14s to complete as opposed to 1.2 s to complete 300 vector averages. However even if we had averaged for 14s we still would not have achieved the spectrum in figure 4 with vector averaging because of the dynamic range limitations discussed above.

We can further optimize our swept sine measurement using some of the features found on analyzers such as the SR780. In order to reduce our sweep time we can make some simple observations about the shape of the transfer function. In both the passband and the tail of the stopband, the filter's transfer function changes very slowly, in other words, no information about the shape of the function would be sacrificed by skipping some frequency points. Obviously at the passband edge and near the zero we don't want to skip any frequency points to preserve the depth of the zero. The SR780's autoresolution mode allows the user to set the maximum measured amplitude change which causes the instrument to start skipping points, the minimum amplitude change which causes the unit to stop skipping points and the maximum number of frequencies which can be skipped. By adjusting these parameters it is possible to obtain the spectrum of figure 4 in only 6 s instead of 14 s with only a 2 dB reduction in the depth of the zero. In general, a well specified autoresolution can take up to 80% off the sweep time while preserving the essential features of the signal. Of course to specify autoresolution parameters some advance knowledge of the signal shape is necessary, but for repetitive measurement situations, where this knowledge is available, the autoresolution feature is very useful.

Of course, even with swept-sine measurements careful attention must be paid to issues such as grounding and stray signal coupling to obtain an input and output with much greater than 100 dB separation. However, with care, the SR780 is capable of making measurements with 145 dB of dynamic range using this technique.

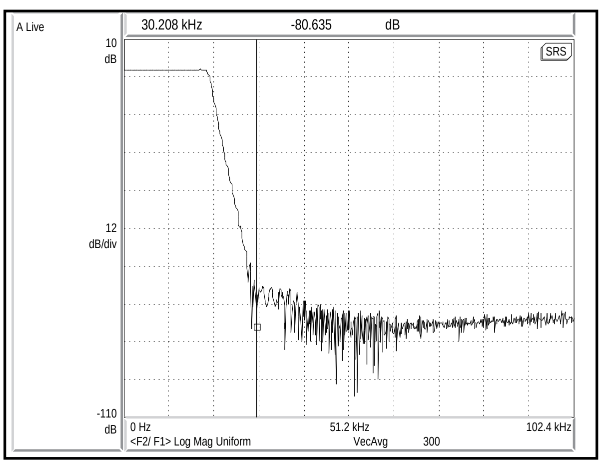


Fig. 3: Vector Averaged Transfer Function

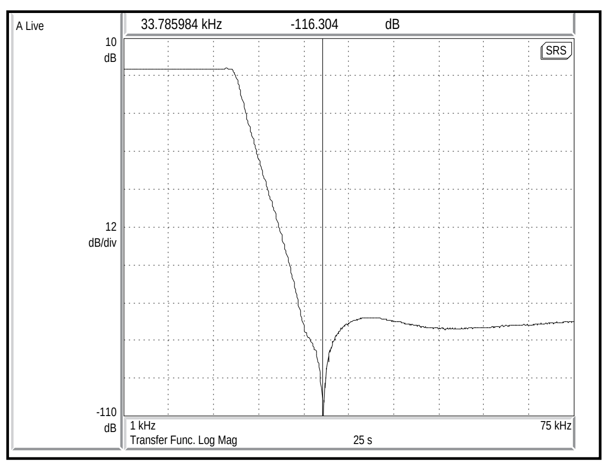


Fig. 4: Swept-Sine Transfer Function

Overview

Specifications apply after 30 minutes of warm-up and within 2 hours of last auto-offset. All specifications are with 400 line FFT resolution and anti-alias filters enabled unless stated otherwise.

Measurements

FFT Group	FFT, Time Record, Windowed Time, Captured Time, Transfer Function, Cross Spectrum, Coherence, Auto and Cross Correlation, Orbit, User Defined
Octave Analysis Group:	1/1, 1/3, 1/12 Octave, Time Capture, User Defined, Leq, Impulse, Total Power
Swept-Sine Group:	Spectrum, Transfer Function, Cross Spectrum, User Defined

Views

All Groups:	Log Mag., Linear Mag., Mag. Squared, Real and Imag., Phase, Unwrapped Phase, Nichols Plot, Nyquist Plot

Frequency

Range	102.4 kHz or 100 kHz (both displays have the same range).
FFT Spans	195.3 mHz to 102.4 kHz or 191 mHz to 100 kHz. The 2 displays can have different spans and start frequencies.
FFT Resolution	100, 200, 400 or 800 lines
Real-Time Bandwidth	102.4 kHz (highest FFT span with continuous data acquisition and averaging).
Accuracy	25 ppm from 20° C to 40° C
FFT Windows:	Hanning, Blackman-Harris, Flattop, Kaiser, Uniform, Force-Exponential, User Defined, ±T/2, ±T/4, 0-T/2

Dynamic Range

Dynamic Range	-90 dBfs typical, -80 dBfs guaranteed (FFT and Octave) 145 dB (Swept Sine) Includes spurs, harmonic and intermodulation distortion and alias products. Excludes alias responses at extremes of span.
Harmonic Distortion	<-80 dB (Single tone in band)
Intermodulation Distortion	<-80 dB (Two tones in band, each <-6.02 dBfs)
Spurious	<-80 dBfs
Alias Responses	<-80 dBfs (Single tone outside of span, <0 dBfs, < 1 MHz)
Full Span FFT Noise Floor	-100 dBfs typical (Input grounded, Input Range > -30 dBV,Hanning window, 64 RMS averages)
Residual DC Response	< -30 dBfs (FFT with Auto Cal On)

Amplitude Accuracy

Single Channel	± 0.2 dB (excluding windowing)
Cross Channel	± 0.05 dB (dc to 102.4 kHz) (Transfer Function measurement, both inputs on the same input range, RMS averaged)

Phase Accuracy

Single Channel	±3.0 deg relative to External TTL trigger (-50 dBfs to 0 dBfs, freq < 10.24 kHz) (Center of frequency bin, DC coupled) For Blackman-Harris, Hanning, Flattop and Kaiser windows, phase is relative to a cosine wave at the center of the time record. For Uniform, Force and Exponential windows, phase is relative to a cosine wave at the beginning of the time record.
Cross Channel	± 0.5 deg (dc to 51.2 kHz) ± 1.0 deg (dc to 102.4 kHz) (Transfer Function measurement, both inputs on the same input range, vector averaged)

Signal Inputs

Number of Inputs	2
Full Scale Input Range	-50 dBV (3.16 mVpk) to +34 dBV (50 Vpk) in 2 dB steps
Maximum Input Level	57 Vpk
Input Configuration	Single-ended (A) or True Differential (A-B)
Input Impedance	1 MΩ + 50 pF
Shield to Chassis	Floating Mode: 1 MΩ+ 0.01 mF Grounded Mode: 50Ω Shields are always grounded in differential input (A-B)
Maximum Shield Voltage	4 Vpk
AC Coupling	-3 dB rolloff at 0.16 Hz
CMRR	90 dB at 1 kHz (In. Range < 0 dBV) 80 dB at 1 kHz (In. Range <10 dBV) 50 dB at 1 kHz (In. Range ≥10 dBV)
ICP™ Signal Cond.	Current Source: 4.8 mA Open Circuit Voltage: +26 V
A-weight Filter	Type 0 Tolerance, ANSI Standard S1.4-1983; 10 Hz to 25.6 kHz
Crosstalk	<-145 dB below signal (Input to Input and Source to Inputs, 50Ω receiving input source impedance)
Input Noise	<10 nVrms/√Hz (< -160 dBVrms/√Hz) above 200 Hz

Trigger Input

Modes	Free run, Internal, External, or External TTL
Internal	Level adjustable to ±100% of input scale. Positive or Negative slope. Minimum Trigger Amplitude: 5% of input range

External	Level adjustable to ±5V in 40 mV steps. Positive or Negative slope. Input Impedance: 1 MΩ Max Input: ±5V Minimum Trigger Amplitude: 100 mV
External TTL	Requires TTL level to trigger (low<0.7V, high>3.0V).
Post-Trigger	Measurement record is delayed up to 8192 samples after the trigger.
Pre-Trigger	Measurement record starts up to 8192 samples prior to the trigger.

Transient Capture

Mode	Continuous realtime data recording to memory.
Maximum Rate	262,144 samples/sec for both inputs
Maximum Cap. Length	2 Msamples (single input) 8 Msamples with optional memory

Octave Analysis

Standards	Conforms to ANSI standard S1.11-1986, Order 3, Type 1-D.
Frequency Range	Band centers: Single Channel 1/1 Octave: 0.125 Hz - 32 kHz 1/3 Octaves: 0.100 Hz - 40 kHz 1/12 Octaves: 0.091 Hz - 12.34 kHz Two Channels 1/1 Octave: 0.125 Hz - 16 kHz 1/3 Octaves: 0.100 Hz - 20 kHz 1/12 Octaves: 0.091 Hz - 6.17 kHz
Accuracy	< 0.2 dB (1 second stable average, single tone at band center)
Dynamic Range	80 dB (1/3 Octave, 2 second stable average) per ANSI S1.11-1986
Sound Level	Impulse, Peak, Fast, Slow and L_{eq} per IEC 651-1979 Type 0

Source Output

Amplitude Range	0.1 mVpk to 5 Vpk
Amplitude Resolution	0.1 mVpk
DC Offset	<10.0 mV (typical)
Output Impedance	< 5Ω, ±100 mA peak output current.

Sine-Source:

Amplitude Accuracy	±1% of setting, 0 Hz to 102.4 kHz 0.1 Vpk to 5.0 Vpk, high impedance load.
Harmonics, SubHarm. and Spurious Signals	0.1 Vpk to 5 Vpk <-80 dBc (fundamental < 30 kHz) <-75 dBc (fundamental < 102 kHz)

Two Tone Source:

Amplitude Accuracy	±1% of setting, 0 Hz to 102.4 kHz 0.1 Vpk to 5 Vpk, high impedance load.
Harmonics, SubHarm.	< -80 dBc, 0.1 Vpk to 2.5 Vpk

White Noise Source:

Time Record	Continuous or Burst
Bandwidth	DC to 102.4 kHz or limited to analysis span.
Flatness	<0.25 dB pk-pk (typical), <1.0 dB pk-pk (max), 5000 rms averages

Pink Noise Source:

Bandwidth	DC to 102.4 kHz
Flatness	<2.0 dB pk-pk, 20 Hz - 20 kHz (measured using averaged 1/3 Octave Analysis)

Chirp Source:

Time Record	Continuous or Burst
Output	Sine sweep across the FFT span.
Flatness	±0.25 dB, Amplitude = 1.0 Vpk

Swept Sine Source:

Auto Functions	Source Level, Input Range and Frequency Resolution
Dynamic Range	145 dB

Arbitrary Source:

Amplitude Range	± 5V
Record Length	2 Msamples (playback from Arbitrary Waveform memory or capture buffer). Variable output sample rate.

12 VDC Converter

Input Voltage	10 VDC to 15 VDC
Output Voltage	115 VAC true RMS ± 5%
Max Cont. Out. Power	200 W
Max Peak Out. Power	300 W
Input No Load Current	.34A

General

Monitor	Monochrome CRT, 800H by 600V resolution.
Interfaces	IEEE-488, RS232 and Printer interfaces standard. All instrument functions can be controlled through the IEEE–488 and RS232 interfaces. A PC (XT) keyboard input is provided for additional flexibility.
Hardcopy	Print to dot matrix and PCL compatible printers. Plot to HPGL or Postscript plotters. Print/Plot to RS232 or IEEE–488 interfaces or to disk file. Additional file formats include GIF, PCX and EPS.
Disk	3.5 inch DOS compatible format, 1.44 Mbytes capacity. Storage of displays, setups and hardcopy.
Preamp Power	Power connector for SRS preamplifiers.
Power	70 Watts, 100/120/220/240 VAC, 50/60 Hz.
Dimensions	17"W x 8.25"H x 24"D
Weight	56 lbs.
Warranty	One year parts and labor on materials and workmanship.

Mobile Instrument Cart

Model CT100 — Mobile Instrument Cart

- ***60 lb. Tray Capacity***
- ***Adjustable Tray Angle***
- ***Heavy Duty Cast Base***
- ***Fits any Full-width Rackmount Instrument***
- ***Ideal for the SR780***

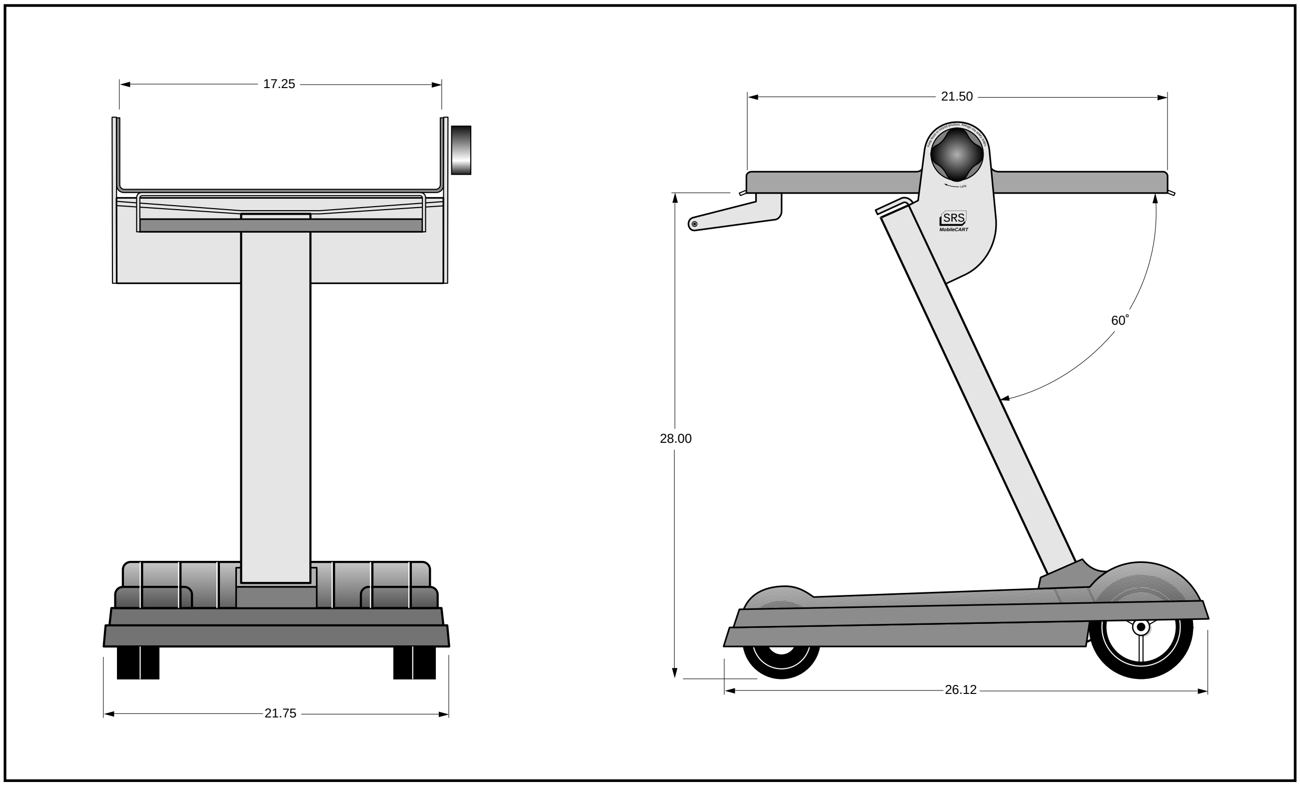
CT100 Overview

The CT100 Instrument Cart is the ideal portability solution for rackmount instrumentation. Unlike other manufacturer's flimsy 'scope carts, the CT100's rugged construction allows you to stack up to 60 pounds of full rack-width instruments in its adjustable tray.

Combined with a car battery and the O78012V DC converter, the CT100 transforms the SR780, or any instrument, into a truly mobile workstation.

CT 100 Specifications

Tray Size	17" x 22" (Accomodates 1 full width rack mount instrument)
Max. Tray Loard	60 lbs.
Max. Base Load	100 lbs.
AngularAdjustment	Horizontal to 60°
Net Weight	58 lbs.



CT100 Dimensions

Ordering Information		
CT100	**Mobile Instrument Cart**	**$495**

FFT Signal Analyzers

Model SR760 — 100 kHz FFT Spectrum Analyzer
Model SR770 — 100 kHz FFT Network Analyzer

- ***DC to 100 kHz Bandwidth***
- ***90 dB Dynamic Range***
- ***16-bit A/D Conversion***
- ***Low Distortion Synthesized Source (SR770)***
- ***PSD and 1/3 Octave Measurements***
- ***Harmonic, Band and Sideband Analysis***
- ***100 kHz Real-time Bandwidth***
- ***Internal and External Triggering***
- ***Linear and Exponential Averaging***
- ***Hardcopy Output to Printers and Plotters***
- ***3.5 inch MS-DOS Compatible Disk Drive***
- ***GPIB and RS-232 Interfaces Included***

SR760/SR770 Overview

The SR760 and SR770 are single channel, 100 kHz FFT Spectrum Analyzers with a dynamic range of 90 dB and a real-time bandwidth of 100 kHz. The SR770 additionally includes a low distortion synthesized source allowing you to measure the transfer functions of electronic and mechanical systems. The instruments' speed and dynamic range of these instruments, coupled with their flexibility and many analysis modes, make them the ideal choice for a variety of applications including acoustics, vibration, noise measurement, and general electronic use.

High Dynamic Range

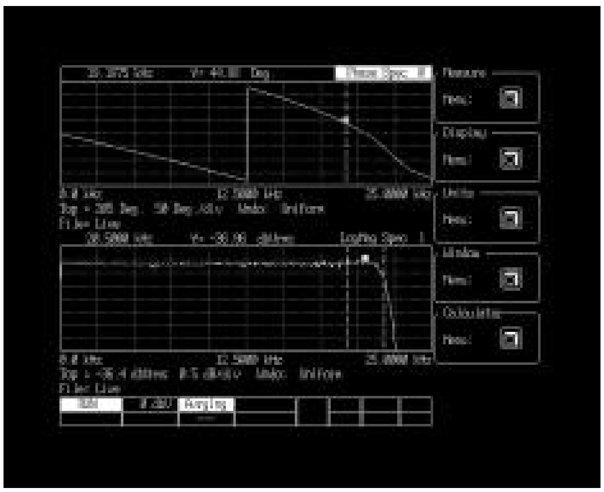
The SR760 and SR770 have a true dynamic range of 90 dB. This means that for a full scale input signal the instruments have no spurious responses larger than –90 dBc (1 part in 30,000). Even signals as small as –114 dBc (1 part in 500,000) may be observed by using averaging. The low front end noise and low harmonic distortion allow you to see signals that are buried in the noise of other analyzers.

At the Speed Limit

The SR760 and SR770 use a pair of high-speed 24-bit digital signal processors (DSP's) to filter, heterodyne, and transform sampled data from its 16-bit analog-to-digital converter. These DSP's can perform 25 million 24-bit multiplications and additions each second. This enormous computing capability allows the analyzers to operate at a real-time bandwidth of 100 kHz. In other words, the SR760 and SR770 process the input signal with no dead time. Your measurements will be done in as little as a tenth of the time of other analyzers, which may typically have real time bandwidths of only 10 kHz.

Easy To Use

The enormous power of the SR760 and SR770 does not come at the expense of ease of use. The simple menu-oriented interface logically groups related instrument functions. Context-sensitive help is available for all keys and menus, and entire instrument setups can be saved to disk and recalled in a single keystroke.



Magnitude and Phase Display

SR760/SR770 Features

Spectrum Measurements

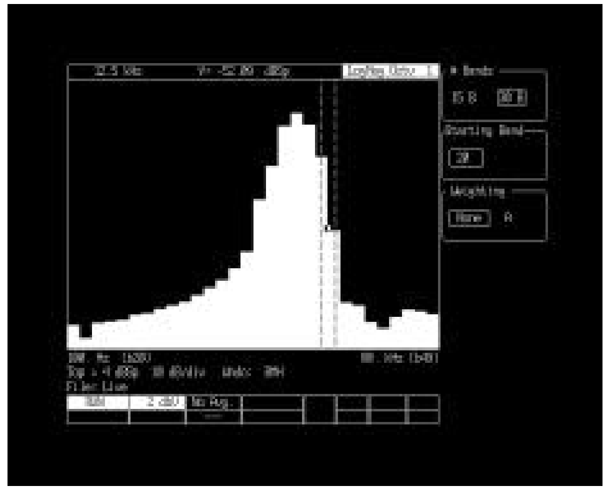
The spectrum, power spectral density, and input time record can be displayed in a variety of convenient linear and logarithmic units including Volts, Vrms, dBV, dBVrms, or user-defined 'Engineering Units' (EU's). The magnitude, phase, and real and imaginary parts of complex signals can all be displayed. Several window functions including Hanning, flattop, uniform, or Blackman-Harris can be chosen to optimize in-band amplitude accuracy or minimize out-of-band sidelobes.

Octave Measurements

The SR760 also computes both the 15 and 30 band 1/3 octave spectra commonly used in acoustics and noise measurement applications. A-weighting compensation is available for octave measurements. Amplitudes are computed for band -2 (630 mHz) through band 49 (80 kHz).

Triggering and Averaging

Flexible triggering and averaging modes let you see

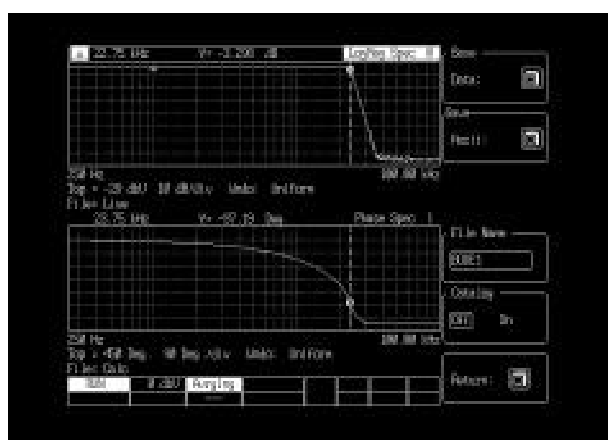


Octave Analysis

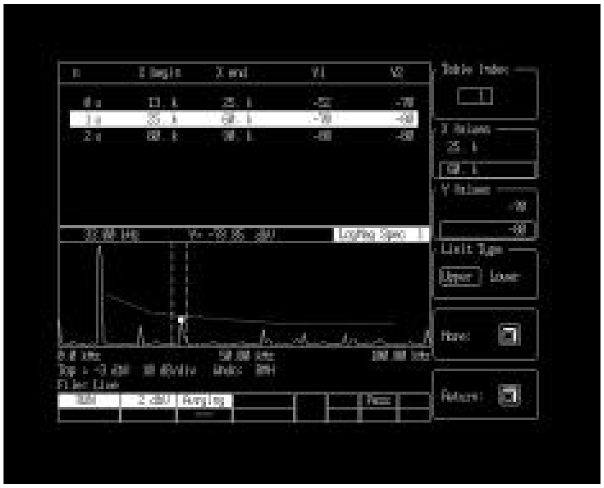
signals as low as 114 dB below full scale. RMS averaging provides an excellent estimate of the true signal and noise levels in the input signal, while vector averaging can be used with a triggered input signal to actually reduce the measured noise level. Both RMS and vector averaging can be performed exponentially, where the analyzer computes a running average (weighting new data more heavily than older data), or linearly, where the analyzer computes an equally weighted average of a specified number of records. Triggering can be used to capture transient events or to preserve spectral phase information. Both internal and external triggering are available with adjustable pre and post-trigger delays.

Synthesized Source

The SR770 includes a very low distortion (-80 dB) synthesized source which can be used to make freqeuncy response measurements. It generates single frequency sine waves, two-tone signals for intermodulation distor-



Bode Plot



Limit Table Display

tion (IMD) testing, pink and white noise for audio and electronic applications, and frequency chirps for transfer function analysis. This direct digital synthesis (DDS) source provides an output level from 100 µV to 1 V, and delivers up to 50 mA of current.

Frequency Response Measurements

With its low distortion DDS source, the SR770 is capable of performing accurate frequency response measurements. The source is synchronized with the instrument's input allowing transfer functions to be measured with 0.05 dB precision. The SR770 measures the magnitude and phase response of control systems, amplifiers, and electro-mechanical systems and displays the resulting Bode plot.

Limit and Data Tables

Sometimes, it's important to keep track of a few key portions of a spectrum. The SR760's and SR770's data tables allow up to 200 selected frequencies to be displayed in a tabular format which can be printed or saved to disk. Automated entry makes it easy to set up data tables for harmonic or sideband analysis. Convenient limit tables allow the entry of up to 100 separate upper or lower limit segments for pass-fail testing. On exceeding a limit, the analyzers can be configured to generate a screen message, an audio alarm, or a GPIB service request.

Analysis Modes

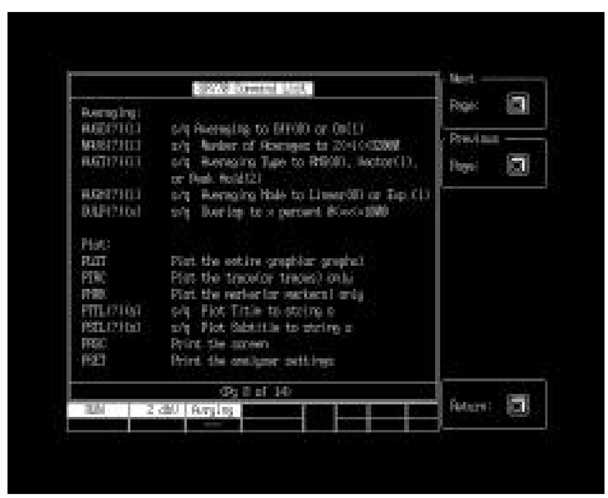
Three built-in analysis modes simplify common measurements. Harmonic analysis computes both harmonic power and THD (Total Harmonic Distortion) relative to a specified fundamental. Sideband analysis lets you compute power in a set of sidebands relative to the carrier

Sideband Analysis

power. And finally, band analysis lets you easily integrate the power in a selected frequency band. All three analysis modes provide clear on-screen markers which make it easy to pick out frequencies of special interest such as harmonics or sidebands.

Finally — a Marker That Works

Too many analyzers have markers that go too slow or too fast, but never seem to end up where you want them. The SR760 and SR770 have a marker that was designed to be fast, responsive, and flexible. The marker can be configured to read the maximum, minimum, or mean of a selected width of display, or can be set to tracking mode to lock on to a moving peak. Delta-mode



On-screen Help

readouts let you easily view frequency or amplitude differences between two peaks. Automated peak-find lets you quickly move between the peaks in a spectrum. And the markers for the upper and lower displays can be linked to easily display similarities or differences in the two spectra.

Math Functions

Data taken with the SR760 and SR770 can be processed with the built-in trace calculator. Basic arithmetic functions such as addition, subtraction, multiplication, division, square roots and logarithms can be performed on traces. Traces can be combined with other on-screen traces, or with traces stored on disks. These calculator functions are quite useful for performing background subtraction or normalization of data.

Flexible Storage and Output

All traces, data tables and limit tables can be stored on the analyzers' 3.5 inch disk drive. The drive uses standard DOS 1.44M (720k for SR760) floppy disks which can be formatted on the analyzer or on your personal computer. Data can be saved in a space-saving binary format, or an easy-to-access ASCII format for off-line analysis. A variety of hardcopy options let you easily view data from the instruments. The screen can be dumped to a dot-matrix printer or a LaserJet compatible laser printer via the standard rear-panel Centronics printer interface. Complete limit and data tables, as well as a summary of the instrument settings can be printed as well. Plotter output is available to any HPGL compatible plotter with an RS-232 or GPIB interface.

Easy to Interface

All functions of the analyzers can be queried and set via the standard RS-232 and GPIB interfaces. A comprehensive set of commands allows flexible control and data transfer to your computer. Data can be quickly transferred in binary format, or more conveniently as ASCII coded numbers. The complete, documented command list is available as a help screen in the instruments for convenient reference while programming.

The Competitive Choice

Spectrum analyzer specifications can be confusing. Is the stated frequency range the real-time bandwidth, or simply the maximum operating frequency? Is the stated dynamic range the spurious-free dynamic range, or simply the A-D converter’s resolution? The SR760 and SR770 have 100kHz real-time bandwidth, as well as 90dB of spurious free dynamic range. Other features are important as well. What is the input range? Does the unit provide a disk for easy storage? Is it easy to get hardcopy output? The SR760 and SR770 give you superior specifications and a broad array of convenient features at a price that is less than 50% of competitive models.

Application Information

Application Note #1, “FFT Spectrum Analyzers,” contains a complete discussion of the fundamentals of FFT spectrum analysis and an explanation of many basic terms used when discussing signal analyzers.

Overview

The SR760 and SR770 Spectrum Analyzers are full featured 100 kHz FFT Spectrum Analyzer with a 90 dB dynamic reserve. Low front-end noise and high speed are coupled with numerous analysis modes (PSD, Harmonic, Band, Limit testing, etc.) to make them useful in a number of applications such as acoustic analysis, audio electronics, noise measurements, and general circuit characterization. Interfacing with the instruments is simple with standard GPIB (IEEE-488) and RS-232 Interfaces, as well as an external printer port and a 3.5" DOS format floppy disk drive.

Frequency

Measurement Range	476 µHz to 100 kHz, baseband and zoomed
Spans	191 mHz to 100 kHz in a binary sequence.
Center Frequency	Anywhere within the 0 to 100 kHz measurement range.
Accuracy	25 ppm from 20°C to 40°C.
Resolution	Span/400
Window Functions	Blackman-Harris, Hanning, Flattop and Uniform.
Real-time Bandwidth	100 kHz

Signal Input

Number of Channels	1
Input	Single-ended or true differential
Input Impedance	1 MΩ, 15 pf
Coupling	AC or DC
CMRR	90 dB at 1 kHz (Input Range < -6 dBV) 80 dB at 1 kHz (Input Range < 14 dBV) 50 dB at 1 kHz (Input Range ≥ 14 dBV)
Noise	5 nVrms/√Hz at 1 kHz (-166 dBVrms/√Hz typical), 10 nVrms/√Hz (-160 dBVrms/√Hz) maximum

Amplitude

Full Scale Input Range	-60 dBV (1.0 mVpk) to +34 dBV (50 Vpk) in 2 dB steps.
Dynamic Range	90 dB typical
Harmonic Distortion	No greater than -90 dB from DC to 50 kHz. (Input Range ≤ 0 dBV) No greater than -80 dB to 100 kHz
Spurious	Input range ≥ -50 dBV: No greater than -85 dB below full scale below 200 Hz. No greater than -90 dB below full scale to 100 kHz.
Input Sampling	16 bit A/D at 256 kHz
Accuracy	± 0.2 dB ± 0.003% of full scale (excluding windowing effects).
Averaging	RMS, Vector and Peak Hold. Linear and exponential averaging up to 64k scans.

Trigger Input

Modes	Continuous, internal, external, or TTL
Internal Level	Adjustable to ±100% of input scale. Positive or Negative slope.
Min. Trigger Amplitude	10% of input range.
External Level	±5V in 40 mV steps. Positive or Negative slope. 10 kΩ Impedance
Min. Trigger Amplitude	100 mV.
External TTL	Requires TTL level, (low<0.7V, high>2V)
Post-Trigger	Measurement record is delayed by 1 to 65,000 samples (1/512 to 127 time records) after the trigger. Delay resolution is 1 sample (1/512 of a record).
Pre-Trigger	Measurement record starts up to 51.953 ms prior to the trigger. Delay resolution is 3.9062 µs.
Phase Indeterminacy	< 2°

Display Functions

Display	Real, imaginary, magnitude or phase spectrum.
Measurements	Spectrum, power spectral density, time record and 1/3 octave.
Analysis	Band, sideband, total harmonic distortion and trace math.
Graphic Expand	Display expand up to 50x about any point.
Harmonic Marker	Displays up to 400 harmonics.
Data Tables	Lists Y values on up to 200 X points.
Limit Tables	Detects data exceeding up to 100 user defined upper and lower limit trace segments.

Source (SR770 Only)

Amplitude Range	0.1 mVpk to 1.0 Vpk
Amplitude resolution	1 mVpk (output>100 mVpk) 0.1 mVpk (output<100 mVpk)
DC offset	<10.0 mV (typical)
Output Impedance	<5Ω, 50 mA peak output current
Sine Source:	
Frequency Range	DC to 100 kHz
Resolution	15.26 mHz
Amplitude accuracy	±1% (0.09 dB) of setting
Spectral purity	-80 dBc, f<10 kHz -70 dBc, f>10 kHz (harmonics and subharmonics) <-100 dB full scale (spurious)
TwoTone Source:	
Frequency Range	DC to 100 kHz
Resolution	15.26 mHz
Amplitude accuracy	±1% (0.09 dB) of setting
Spectral purity	-80 dBc, f<10 kHz -70 dBc, f>10 kHz (harmonics and subharmonics) <-100 dB full scale (spurious)
White Noise Source:	
Frequency Range	DC to 100 kHz (all spans)
Flatness	<1.0 dBpp (rms averaged spectra)
Pink Noise Source:	
Frequency range	DC to 100 kHz (all spans)
Flatness	<4.0 dBpp (using 1/3 octave analysis)

Chirp Source:

Output	Equal amplitude sine waves at each frequency bin of the current span.
Flatness	<0.05 dBpp (typical) <0.2 dBpp (max)
Phase	AutoPhase function calibrates to current phase spectrum.

General

Monitor	Monochrome CRT. 640H by 480V resolution. Adjustable brightness and position.
Interfaces	IEEE-488, RS-232 and Printer interfaces standard. A PC keyboard input is provided for additional flexibility.
Hardcopy	Screen dumps and table and setting listings to dot matrix and HP LaserJet compatible printers. Data plots to HP-GL compatible plotters (RS-232 or IEEE-488).
Disk	3.5" DOS compatible format, 1.44Mbyte capacity. (SR760 is 720 kbyte) Stores data and instrument configuration.
Power	60 W, 100/120/220/240 VAC, 50/60 Hz
Dimensions	17" W x 6.25" H x 18.5" D
Weight	36 lbs.
Warranty	One year parts and labor on any defects in material or workmanship.

SR760/SR770 Rear Panel

Ordering Information

SR760	**FFT Spectrum Analyzer**	**$ 4950**
SR770	**FFT Network Analyzer**	**$ 6500**
O760RM	**Rack Mount**	**$ 85**
O760H	**Carrying Handle**	**$ 100**

Synthesized Function Generator

Model DS345 — 30 MHz Function & Arbitrary Waveform Generator

- ***Frequency Output from 1 μHz to 30.2 MHz***
- ***16,300 Point Arbitrary Waveforms***
- ***Sine, Square, Ramp and Noise Output***
- ***10 V (peak-peak) Output into 50Ω***
- ***1 μHz Frequency Resolution***
- ***Phase Continuous Frequency Sweeps***
- ***AM, FM, and Phase Modulation***
- ***Optional RS-232, GPIB interfaces***

DS345 Overview

The DS345 is a full featured 30 MHz synthesized function generator. Using an innovative Direct Digital Synthesis (DDS) architecture, the DS345 provides outputs of high spectral purity, outstanding frequency and phase agility, and remarkably versatile modulation and arbitrary waveform capability.

Versatile Outputs

The standard waveforms–sine, square, ramp and triangle, are all available with 1 μHz frequency resolution. You can generate sine and square waves up to 30.2 MHz or ramps and triangles up to 100 kHz. All functions can be swept over their full frequency range both linearly and logarithmically. There is even a wideband 10 MHz white noise source. With direct digital synthesis, frequency changes are fast (only 25 ns) and are made with no loss of phase continuity.

Arbitrary Waveforms

DDS also gives the DS345 the ability to generate fast, high resolution arbitrary waveforms. Arbitrary waveforms up to 16k points long, with 12 bits of vertical resolution, can be stored and output at sample rates of up to 40 Msamples/s. Powerful PC software to aid in designing and downloading arbitrary waveforms is also available.

Complex Modulation Patterns

The DS345 has unparalleled modulation capabilities. All functions may be modulated in frequency, amplitude, or phase by an internal digitally synthesized modulation source. The internal source provides standard modulation patterns or can be programmed for arbitrary modulation waveforms. Output signals can also be amplitude modulated by an external source.

The Competitive Choice

The DS345 provides all the features of traditional function generators, sweep generators, and arbitrary waveform generators with no compromise in performance or ease of use. And because it uses Direct Digital

Synthesis, the DS345 can provide this performance at a price well below that of conventional function generators.

DS345 Features

Functions and Outputs

The DS345 synthesizes high accuracy sine and square waves up to a frequency of 30 MHz. Triangle and ramp waveforms up to 100 kHz can also be generated. The frequency resolution is 1 µHz in all cases. Frequencies can be simply entered from the front panel using the numeric keypad, or by modifying the existing entry with cursor keys. In addition to the standard function waveforms, the unit also provides a wideband (10 MHz) white noise source.

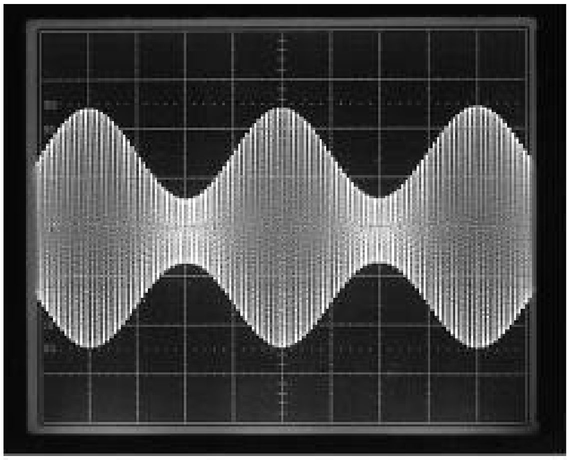
Both the function output and a TTL SYNC output are available through floating front-panel BNC connectors. Both outputs have 50Ω output impedance and may be floated by ±40V relative to earth ground. The amplitude of all function outputs is adjustable from 10 mVpp to 10 Vpp with 3 digit resolution. A convenient feature allows the specification and display of amplitudes in either Volts, Volts RMS, or dBm. In addition, standard TTL and ECL output levels can be selected.

Additional useful connectors are provided on the rear panel. A trigger input is used to trigger arbitrary waveforms, modulation patterns, sweeps, and bursts, while a TTL trigger output is provided to allow synchronization of external devices to sweeps and bursts. A sweep output generates a 0-10 V ramp synchronous with frequency sweeps. The sweep marker outputs allow specified portions of a frequency sweep to be highlighted on a plot or oscilloscope display.

The 10 MHz input allows the DS345 to be synchronized to any external timebase. When used in conjunction with another DS345's 10 MHz clock output, it allows the two units to operate in a fixed phase relationship. Two such synchronized DS345's can have their relative phase set between 0° and ±7200° with 0.001° phase resolution.

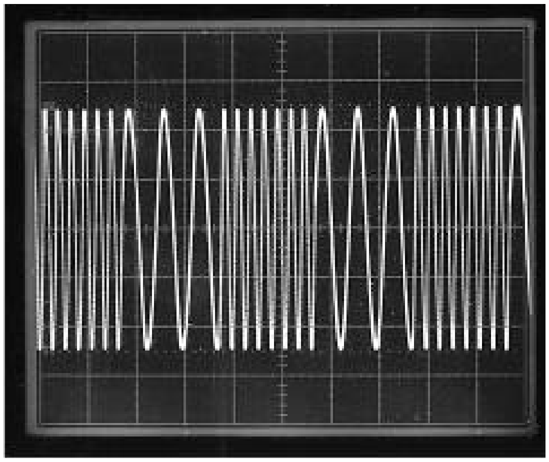
Modulation

The DS345 allows a wide variety of flexible modulation options. It contains an internal digitally synthesized modulation generator which can modulate any of its standard waveforms except noise. The modulation waveform can be chosen to be a sine, square, triangle, ramp, or an arbitrary modulation waveform can be specified using the optional GPIB or RS-232 interface. Modulation rates from 1 mHz to 10 kHz can be selected with the internal modulation generator.



Amplitude Modulation of Sine by Sine

The internal modulation generator can provide amplitude modulation, frequency modulation, and phase modulation. When using amplitude modulation (AM), modulation depths of ±100% can be selected with 1% resolution. Negative values of modulation correspond to Double Sideband Suppressed Carrier (DSBSC) modulation. To select frequency modulation, you enter a modulation ‘span.’ The output frequency will vary between the center frequency plus and minus one half the span. FM spans can be selected with 1 mHz resolution. Finally, with phase modulation, you choose a phase span between 0° and 7200° with 0.001° resolution. The output phase will vary between 0° and the selected phase span. In all three cases, the rear-panel modulation output provides a 0-5 Volt signal whose amplitude is proportional to the modulation strength.



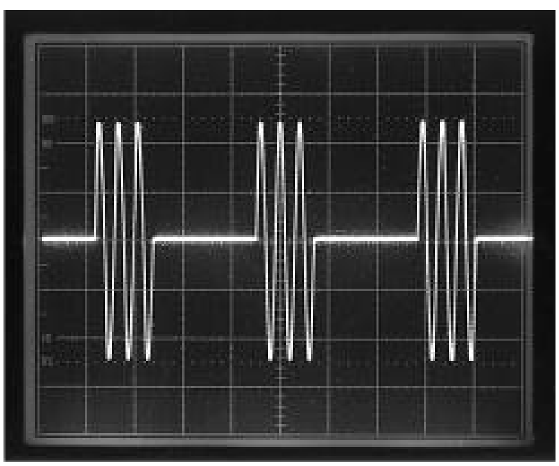
Frequency Shift Keying of Sinewave

External Amplitude Modulation

In addition to the internal modulation generator, the output waveform can be amplitude modulated by an external signal applied to the rear-panel AM input. This input is always active–even when other modulation types are turned on. Applying 5 Volts at this input causes the output to assume its maximum value, 0 Volts will turn the output off, and -5 Volts will invert the output. The bandwidth of the rear-panel AM input is 15 kHz.

Burst Modulation

You can generate tone bursts of any output function except noise. In the burst mode, the DS345 will output an exact number of complete waveform cycles after receiving a trigger. By adjusting the phase, you can control where in the waveform the burst begins. While using the burst mode, the maximum frequency for sine and square waves is 1 MHz, while triangles and ramps are limited to 100 kHz. Burst mode may be used with arbitrary waveforms at any frequency. Up to 30,000 complete waveform cycles may be specified in a burst.



Three Cycle Burst of Sine

Flexible Sweeps

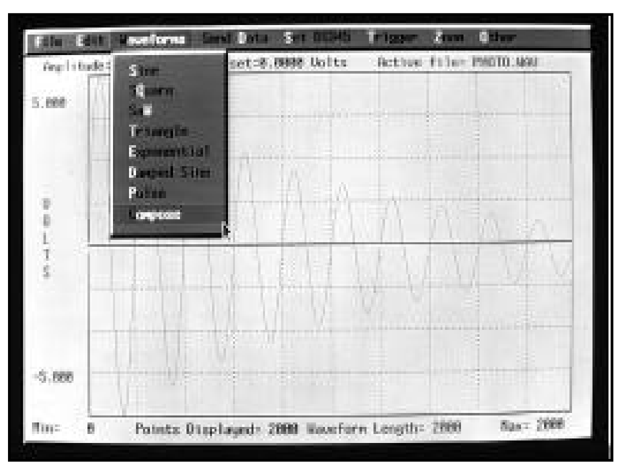
The DS345 can frequency sweep any of its function outputs (except noise). You can sweep up or down in frequency using linear or log sweeps. And unlike conventional function generators, there are no annoying discontinuities or band-switching artifacts when sweeping through certain frequencies. The DS345's DDS architecture inherently allows it to perform smooth, phase-continuous sweeps over it's entire frequency range. Sweeps can either be configured as single-shot, repetitive, or bi-directional. When sweeping bi-directionally, the DS345 will sweep from the start to the stop frequency, then from the stop to the start frequency, and repeat.

Two sweep marker frequencies can be specified. When the sweep crosses either of the marker frequencies, a TTL transition is generated at the rear-panel MARKER output to allow synchronization of external devices to the sweep.

Arbitrary Waveform Capability

The DS345 isn't just a function generator. It's also a full-featured arbitrary waveform generator. You can enter arbitrary waveforms in two formats–as points or as vectors. In the point format the DS345 stores a list of up to 16k output amplitude values. Each amplitude value is specified as a 12-bit integer. In the vector format, data is stored as a list of up to 6144 x,y pairs. Each pair specifies an address in waveform RAM (the x value) and a 12-bit amplitude (the y value). The DS345 computes the values between the specified pairs by linear interpolation. With either format, data in the waveform RAM can be played back at up to 40 Msamples/s, or at any integer submultiple of 40 Msamples/s.

Since composing complex arbitrary waveforms at the keyboard can be a tedious task, Arbitrary Waveform Composer (AWC) software is provided when you order the DS345/01(GPIB and R-S232 interface) option. AWC is a simple, menu-based program which lets you create and edit arbitrary waveforms on the screen, store them, and download them to the DS345. The AWC software lets you download these waveforms to the DS345 or a disk file via the computer interface. AWC runs on all PC compatible computers with 640K of memory and DOS 2.1 or greater. All major graphics modes (CGA/EGA/VGA/HGC) are supported.



Arbitrary Waveform Composer Software

What is Direct Digital Synthesis?

Direct digital synthesis (DDS) has had a dramatic impact on the "best approach" to bench-top function generators. Over the last few years improvements in LSI logic, fast random access memories (RAM), and digital-to-analog converters (DACs) have made DDS the technology of choice for this application.

There are three major components to a DDS: a phase accumulator, a sine look-up table, and a DAC. The phase accumulator computes an address for the sine table (which is stored in RAM). The sine value is converted to an analog value by the DAC. To generate a fixed frequency sine wave, a constant value (called the Phase Increment) is added to the phase accumulator with each clock. If the phase increment is large, the phase accumulator will step quickly through the sine look-up table, and so generate a high frequency sine wave.

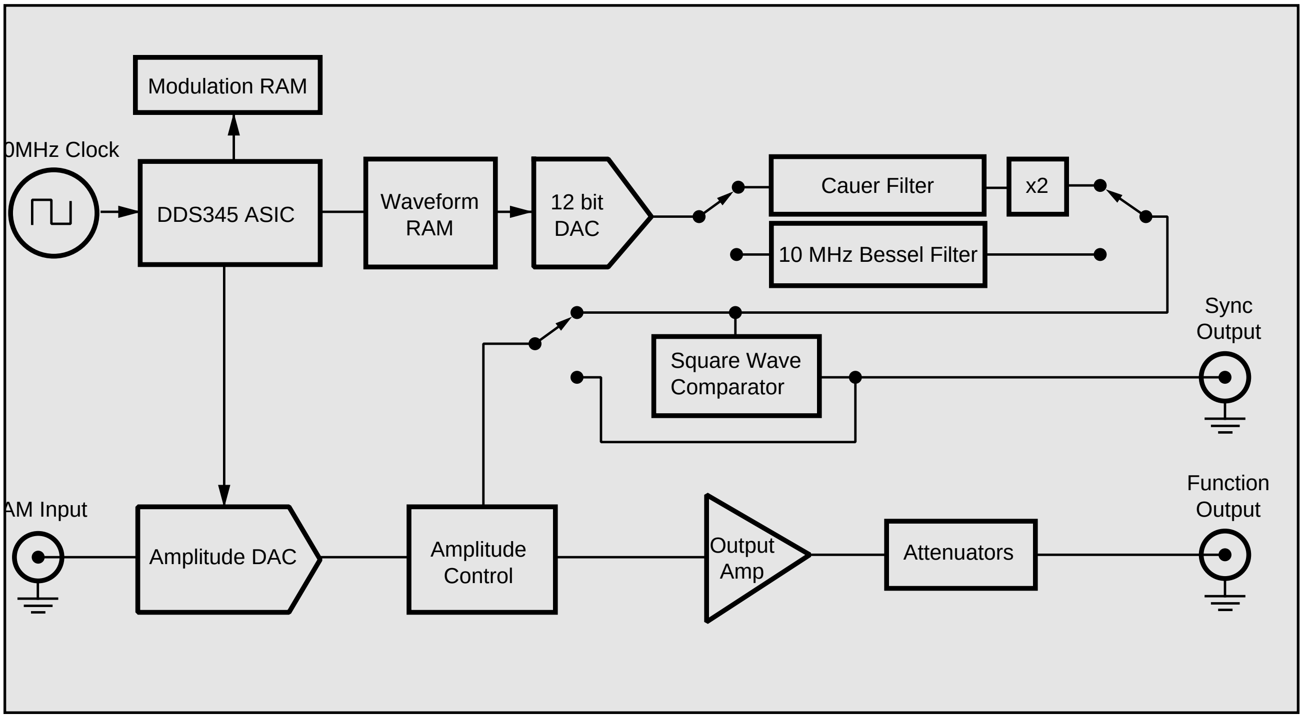
One might think that to generate a "clean" sine wave you would need hundreds or thousands of points in each cycle of the sine wave. In fact, you need about three. Of course a three step approximation to a sine wave hardly looks like a sine wave, but if you follow the DAC with a very good low-pass filter, all the high frequency components are removed, leaving a very clean sine wave indeed.

The frequency resolution of the DDS is given by the number of bits in the phase increment and phase accumulator. Lots of bits provide very high frequency resolution. The DS345 uses a 48-bit phase accumulator for a frequency resolution of 1 part in 10^{14}. This provides 1 µHz resolution at all frequencies from 1 µHz to 30 MHz.

The maximum frequency depends on how fast you can add the 48-bit phase increment to the phase accumulator. Using a highly pipe-lined architecture these additions can be performed at 40 MHz. This allows direct digital synthesis to 15 MHz. A frequency doubler is used to reach 30 MHz.

For agile frequency and phase modulation, it is necessary to change the phase increment values quickly. To do this, the phase accumulator may switch between two 48-bit phase increment values in 25 ns, and each of these 48-bit registers may be loaded in less than 1 us. During frequency modulation one register is used while loading the other.

A 1.2 micron CMOS gate array is used to reduce size, power and cost in the DS345. The gate array does the 48-bit additions at 40 MHz, and operates as a processor, modifying its control registers up to 10 million bytes per second–eliminating a traditional bottleneck which prevents rapid modulation of conventional direct digital synthesizers. The gate array also handles the trigger and counting logic for the arbitrary waveform functions.



Overview

The DS345 is a full-featured synthesized function generator that can generate standard and arbitrary waveforms from 1 µHz to 30.2 MHz. The standard output functions (sine, square, triangle and ramp) can all be AM, burst, FM and phase modulated. Arbitrary waveforms and phase continuous sweeps are all simply implemented by the direct digital synthesis. The instrument's tremendous versatility and accuracy make it attractive for a wide range of applications.

Frequency Range

	Freq	Resolution
Sine	30.2 MHz	1 µHz
Square	30.2 MHz	1 µHz
Ramp	100 kHz	1 µHz
Triangle	100 kHz	1 µHz
Noise	10 MHz	(Gaussian Weighting)
Arbitrary	10 MHz	40 MHz/n sample rate

Output

Source Impedance	50Ω
Grounding	Output may float up to ±40V (AC + DC) relative to earth ground.

Amplitude

Range	0.01 to 10 Vpp into 50Ω 20 Vpp into high impedance
Resolution	3 digits (DC offset = 0 V)

Sine Wave Accuracy (0V DC Offset):

	1mHz-100 kHz	100kHz-20 MHz	20MHz-30.2 MHz
5-10 Vpp	±0.2 dB	±0.2dB	±0.3dB
0.01-5 Vpp	±0.4dB	±0.4dB	±0.5dB

Square Wave Accuracy:

	1mHz-100 kHz	100kHz-20 MHz	20MHz-30.2 MHz
5-10 Vpp	±3%	±6%	±9%
0.01-5 Vpp	±5%	±8%	±12%

Triangle, Ramp, Arbitrary	±3%	> 5Vpp
Accuracy	±5%	< 5Vpp

DC Offset

Range	±5V (limited such that |Vac peak| +|Vdc| < 5 V)
Resolution	3 digits (Vac = 0)
Accuracy	1.5% of setting + 0.2 mV (DC only) ±0.8 mV to ±80 mV depending on AC and DC settings

Sine Wave

Spurious Components	<-55 dBc (non-harmonic)
Phase Noise	<-50 dBc in a 30 KHz band centered on the carrier, exclusive of discrete spurious signals
Subharmonic:	<-50 dBc
Harmonic Distortion	Harmonically related signals will be less than:

Level	Frequency Range
<-60 dBc	DC to 100 kHz
<-45 dBc	.1 to 1 MHz
<-30 dBc	1 to 30 MHz

Square Wave

Rise/Fall Time	< 15 nS (10 to 90%), at full output
Asymmetry	< 1% of period + 4 nS
Overshoot	< 5% of peak to peak amplitude at full output

Ramps, Triangle and Arbitrary Waveforms

Rise/Fall Time	45 nS (10 MHz Bessel Filter)
Linearity	±0.5% of full scale output
Settling Time	< 1 µs to settle within 0.1% of final value at full output

Arbitrary Waveforms

Sample Rate	40 MHz/N, N = 1 to 2^{34}-1.
Memory Length	8 to 16,300 points
Resolution	12 bits (0.025% of full scale)

Phase

Range	±7199.999° with respect to arbitrary starting phase
Resolution	0.001°

Amplitude Modulation

Source	Internal (sine, square, triangle, or ramp) or External
Depth	0 to 100% AM or DSBSC
Rate	0.001Hz to 10 kHz internal, 15 kHz max external
Distortion	<-35dB at 1 kHz, 80% depth
DSB Carrier	<-35dB typical at 1 kHz modulation rate (DSBSC)
External Input	±5 V for 100% modulation, 100 kΩ impedance, 15 kHz BW.

Frequency Modulation

Source	Internal (sine, square, triangle, ramp or arbitrary)
Rate	0.001 Hz to 10 kHz
Span	1 mHz to 30.2 MHz (100 kHz for triangle,ramp)

Phase Modulation

Source	Internal (sine, square, triangle, ramp)
Rate	0.001 Hz to 10 kHz
Span	±7199.999°

Frequency Sweep

Type	Linear or Log, phase continuous
Waveform	Up, down, up-down, single sweep
Time	0.001 s to 1000 s
Span	1 mHz to 30.2 MHz (to 100 kHz for triangle,ramp)
Markers	Two markers may be set at any sweep point (TTL output)
Sweep Output	0 - 10 V linear ramp signal, synchronized to sweep

Burst Modulation

Waveform	Any waveform except noise may be burst modulated
Frequency	Sine, square to 1 MHz; triangle, ramp to 100 kHz; arbitrary to 40 MHz sample rate
Count	1 to 30,000 cycles/burst (1ms to 500 s burst time limits)

Trigger Generator

Source	Single, Internal, External, Line
Rate	0.001 Hz to 10 kHz internal (2 digit resolution)
External	Positive or Negative edge, TTL input
Output	TTL output

Standard Timebase

Accuracy	±5 ppm (20 to 30° C)
Aging	5 ppm/year
Input	10 MHz/N ± 2 ppm. N = 1 to 8. 1V pk-pk minimum input level.
Output	10 MHz, >1 Vpp sine into 50 Ω

Optional Timebase

Type	Ovenized AT-cut oscillator
Stability	< 0.01ppm, 20 - 60°C
Aging	< 0.001ppm/day
Short Term	< 5 x 10^{-11} 1s Allan Variance

General

Interfaces	Optional RS-232 (300 to 9600 Baud, DCE) and GPIB with DOS based Arbitrary Waveform Software. All instrument functions are controllable over the interfaces.
Dimensions	8.5" x 3.5" x 13" (W x H x L)
Weight	10 lbs
Power	50 W, 100/120/220/240 Vac 50/60 Hz
Warranty	One year parts and labor on any defects in material or workmanship.

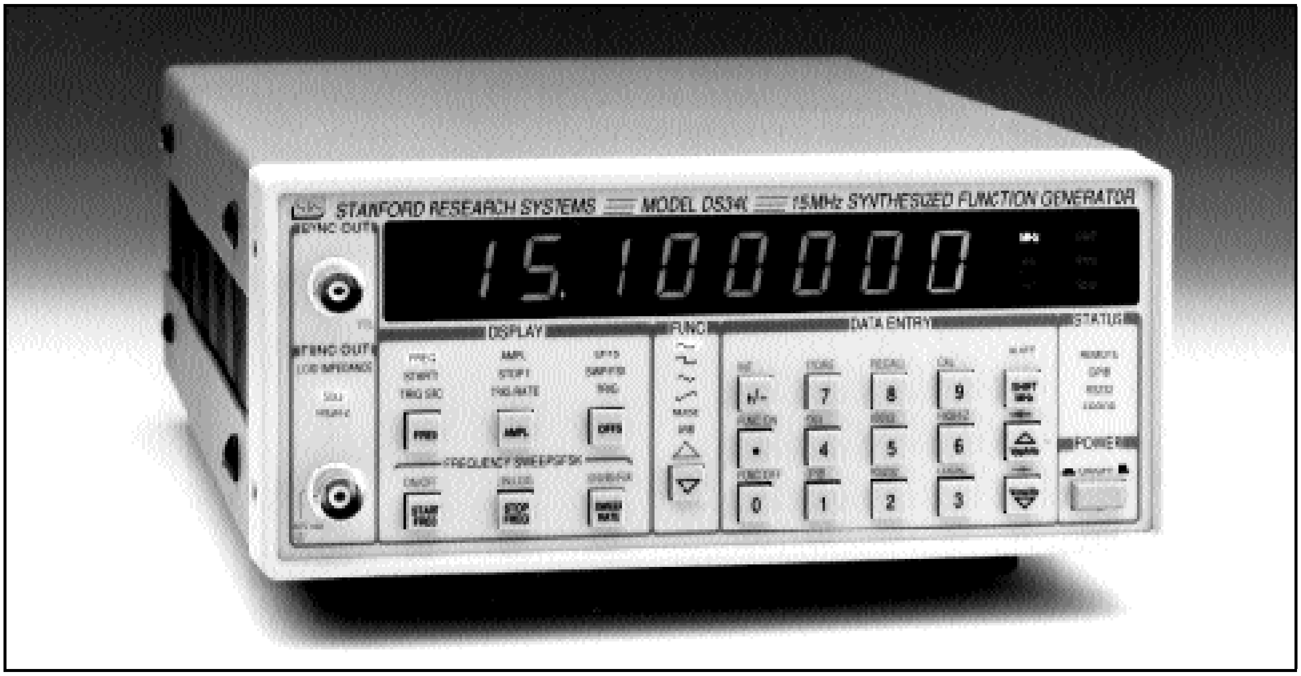
DS345 Rear Panel

Ordering Information

DS345	**Function Generator**	**$ 1595**
Option 01	**GPIB and RS-232 Interfaces**	**$ 495**
Option 02	**High Stability Timebase**	**$ 650**
O345RMS	**Single Rackmount**	**$ 85**
O345RMD	**Double Rackmount**	**$ 85**

Synthesized Function Generator

Model DS340 — 15 MHz Function and Arbitrary Waveform Generator



- ***Frequency Output from 1 µHz to 15.1 MHz***
- ***Sine, Square, Ramp and Noise Output***
- ***16,300 Point Arbitrary Waveforms***
- ***Linear and Log Frequency Sweeps***
- ***1 µHz Frequency Resolution***
- ***FSK Modulation***
- ***Optional RS-232, GPIB interfaces***
- ***Optional 2ppm TCXO Timebase***

DS340 Overview

The DS340 is a general-purpose 15 MHz function generator based on Direct Digital Synthesis (DDS). The DS340 also has the ability to output arbitrary waveforms up to 16,300 points long. A combination of features, performance, and low cost make the DS340 ideal for a variety of test and measurement applications.

A Function Generator and a Sweep Generator

You can generate sine and square waves up to 15.1 MHz, and ramps and triangles up to 100 kHz. Frequency resolution is 1 µHz for all functions. The DS340 also includes a 10 MHz gaussian white-noise generator.

All functions can be swept logarithmically or linearly in a phase-continuous fashion over the entire frequency range of the instrument. A rear-panel SWEEP output provides a trigger output at the start of a sweep to allow synchronization of external devices. Both uni-directional and bi-directional sweeps can be selected, as well as single-shot and repetitive sweeps.

Arbitrary Waveforms

Up to 16,300 12-bit arbitrary waveform points can be downloaded to the DS340's waveform memory via the optional GPIB or RS-232 interfaces. PC software for composing, editing, and downloading arbitrary waveforms is provided when the GPIB and RS-232 interfaces are ordered. The waveform memory can be played back at sample rates up to 40 Msamples/s.

FSK Modulation

Both internal and external FSK modes allow the output frequency to be rapidly toggled between two preset values. FSK toggling can be done at a fixed rate of up to 50 kHz, or externally. The external FSK input is sampled every 25 ns.

GPIB and RS232

GPIB and RS-232 interfaces are available providing complete instrument control by an external computer.

Frequency Range

	Freq	Resolution
Sine	15.1 MHz	1 µHz
Square	15.1 MHz	1 µHz
Ramp	100 kHz	1 µHz
Triangle	100 kHz	1 µHz
Noise	10 MHz	(Gaussian Weighting)

Output

Source Impedance	50Ω
Grounding	Output may float up to ±40V (AC+DC)

Amplitude

Range	50 mVpp to 10 Vpp into 50Ω 100 mVpp to 20 Vpp into high Z
Resolution	3 digits (DC offset = 0V)
Offset	±5 VDC (50Ω) ±10VDC (High Z)
Offset Resolution	10 mV (50Ω) 20 mV (High Z)
Accuracy	0.1 dB (sine output)

Sine Wave

Spurious Response	<-65 dBc
Harmonic distortion:	
DC to 20 kHz	<-70 dBc
20 kHz to 100 kHz	<-60 dBc
100 kHz to 1 MHz	<-50 dBc
1 MHz to 15 MHz	<-40 dBc
Phase noise	<-55 dBc (30 kHz band centered on carrier)

Square Wave

Rise/Fall time (10-90 %)	17.5±5 ns
Asymmetry	3 ns+1%
Overshoot	<2 %

Ramps and Triangles

Rise/Fall time(10-90%)	45±10 ns
Linearity	±0.1% of full scale
Settling Time	200 ns (0.5% of final value)

Arbitrary Waveforms

Sample Rate	40 MHz or integer submultiples
Waveform Length	8 to 16,300 points
Vertical Resolution	12 bits
Rise/Fall time(10-90%)	45±10 ns

FSK Modulation

Modes	Internal, External
Max Rate	50 kHz Internal External FSK input sampled every 25 ns

Sweeps

Type	Linear and Logarithmic
Span	Linear, Full Frequency Range Log, 6 decades
Sweep Rate	0.01 Hz to 1 kHz

Timebase Accuracy

Standard	±5 ppm (20-30 °C)
Optional	TCXO, 2ppm stability, 2 ppm aging (20-50 °C)

Connectors

Front Panel:	
	FUNC Function Output
	SYNC TTL level into 50Ω
Rear Panel:	
	FSK /TRIG input TTL, triggers sweep, ARB, or FSK shift
	SWP/FSK out TTL
TRIG	TTL

General

Interfaces	Optional RS-232 (300 to 9600 Baud, DCE) and GPIB with DOS based Arbitrary Waveform Software. All instrument functions are controllable over the interfaces.
Non-volatile memory	Up to 9 sets of instrument settings may be stored and recalled
Dimensions	8.5" x 3.5" x 13" (W x H x L)
Weight	8.5 lbs
Power	32 W, 100/120/220/240 Vac 50/60 Hz
Warranty	One year parts and labor on any defects in material or workmanship.

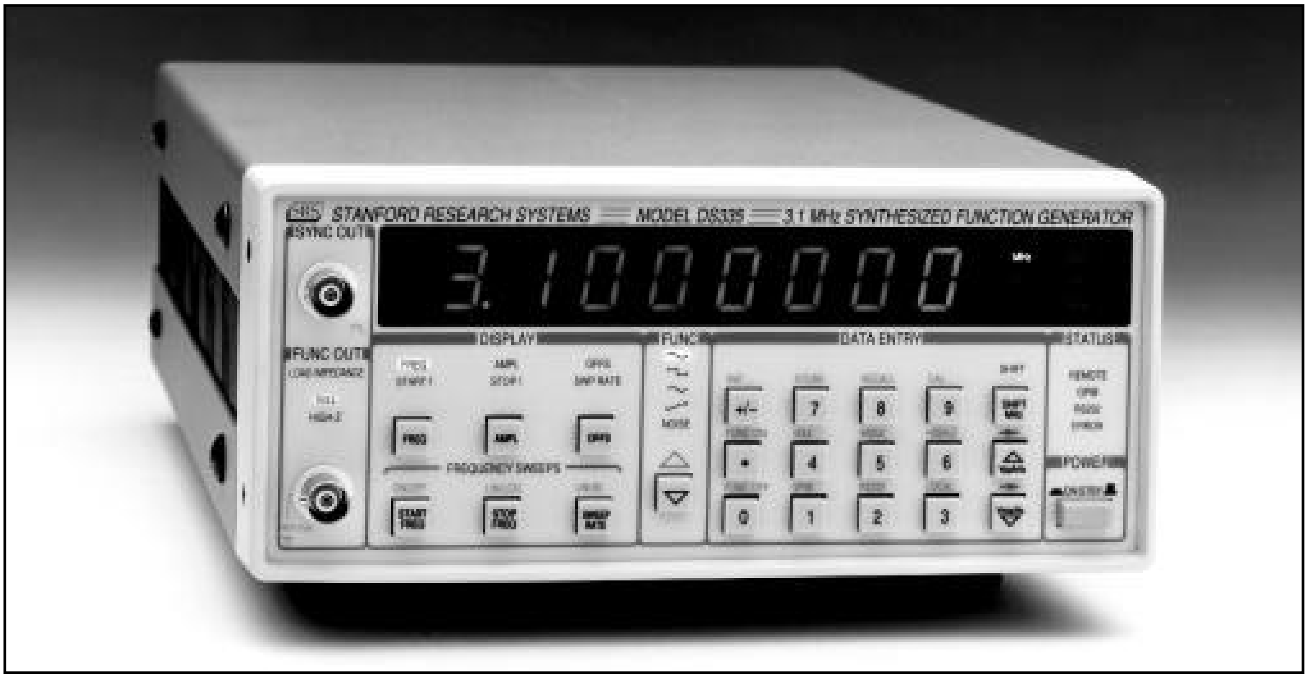
DS340 Rear Panel

Ordering Information

DS340	**Function Generator**	**$ 1195**
Option 01	**GPIB and RS-232, AWC software**	**$ 495**
Option 02	**TCXO Oscillator**	**$ 350**
O345RMS	**Single Rackmount**	**$ 85**
O345RMD	**Double Rackmount**	**$ 85**

Synthesized Function Generator

Model DS335 — 3 MHz Function Generator



- ***Frequency Output from 1 µHz to 3.1 MHz***
- ***Sine, Square, Ramp and Noise Output***
- ***10 Vpp into 50 Ω, 20 Vpp into High-Z***
- ***Linear and Log Frequency Sweeps***
- ***1 µHz Frequency Resolution***
- ***FSK Modulation***
- ***Optional RS-232, GPIB interfaces***
- ***Optional 2 ppm TCXO Timebase***

DS335 Overview

The DS335 is a simple, low-cost, 3 MHz function generator designed for general benchtop or ATE applications. Based on a Direct Digital Synthesis (DDS) architecture, the DS335 includes features not normally found in function generators in this price range.

A Versatile Function Generator

Basic functions include sine, square, (up to 3.1 MHz) and ramps and triangles (up to 10 kHz). You can set the frequency of any function with 1 µHz resolution. Frequency accuracy is 5 ppm with the standard timebase and 2 ppm with the optional TCXO timebase. A 3.5 MHz gaussian white-noise generator is also provided.

All functions can be swept logarithmically or linearly in a phase-continuous fashion over the entire frequency range. A rear-panel SWEEP output marks the beginning of a sweep to allow synchronization of external devices. Both uni-directional and bi-directional sweeps can be selected, as well as single-shot and repetitive sweeps.

FSK Modulation

Both internal and external FSK modes allow the output frequency to be rapidly toggled between two preset values. Toggling is done either at a fixed, internal rate of up to 50 kHz, or externally via a rear-panel input.

Low Distortion Outputs

Outputs have the low phase noise inherent to DDS. Wideband amplifiers maintain good pulse response and provide low distortion. The result is an output capable of driving 10 Vpp into a 50Ω load with only a 15 ns risetime, or 20 Vpp into a high-impedance load.

GPIB and RS-232

Both GPIB and RS-232 interfaces are available to provide complete control via an external computer. All instrument functions can be set and read via the computer interfaces.

Frequency Range

	Freq	Resolution
Sine	3.1 MHz	1 µHz
Square	3.1 MHz	1 µHz
Ramp	10 kHz	1 µHz
Triangle	10 kHz	1 µHz
Noise	3.5 MHz	(Gaussian Weighting)

Output

Source Impedance	50Ω
Grounding	Output may float up to ±40V (AC + DC) relative to earth ground.

Amplitude

Range	50 mVpp to 10 Vpp (50Ω) 100 mVpp to 20 Vpp (High Z)
Resolution	3 digits (DC offset = 0V)
Offset	±5 VDC (50Ω) ±10 VDC (High Z)
Offset Resolution	10 mV (50Ω) 20 mV (High Z)
Accuracy	0.1 dB (sine output)

Sine Wave

Spurious Response	<-65 dBc
Harmonic distortion:	
DC to 20 kHz	<-70 dBc
20 kHz to 100 kHz	<-60 dBc
100 kHz to 1 MHz	<-50 dBc
1 MHz to 3 MHz	<-40 dBc
Phase noise	<-60 dBc (30 kHz band centered on carrier)

Square Wave

Rise/Fall time (10-90 %)	15±5 ns
Asymmetry	3 ns + 1%
Overshoot	<5 %

Ramps and Triangles

Rise/Fall time(10-90%)	100±20 ns
Linearity	±0.1% of full scale
Settling Time	200 ns (0.5% of Final Value)

FSK Modulation

Modes	Internal, External
Max Rate	50 kHz Internal External FSK input sampled every 25 ns

Sweeps

Type	Linear and Logarithmic
Span	Linear, Full Frequency Range Log, 6 decades
Sweep Rate	0.01 Hz to 1 kHz

Timebase Accuracy

Standard	±5 ppm (20-30 °C)
Optional	TCXO, 2ppm stability, 2 ppm aging (20-50 °C)

Connectors

Front Panel:	
FUNC	Function Output
SYNC	TTL level into 50Ω
Rear Panel:	
FSK input	TTL
Sweep/FSK out	TTL

General

Interfaces	Optional RS-232 (300 to 9600 Baud, DCE) and GPIB. All instrument functions are controllable over the interfaces.
Non-volatile memory	Up to 9 sets of instrument settings may be stored and recalled
Dimensions	8.5" x 3.5" x 13" (W x H x L)
Weight	8 lbs
Power	22 W, 100/120/220/240 Vac 50/60 Hz
Warranty	One year parts and labor on any defects in material or workmanship.

DS335 Rear Panel

Ordering Information

DS335	**Function Generator**	**$ 995**
Option 01	**GPIB and RS-232 Interfaces**	**$ 395**
Option 02	**TCXO Oscillator**	**$ 350**
O345RMS	**Single Rackmount**	**$ 85**
O345RMD	**Double Rackmount**	**$ 85**

Synthesized Function Generator

Model DS360 — Ultra-low Distortion Function Generator

- ***Frequency Output from 1 mHz to 200 kHz***
- ***Sine, Square, White and Pink Noise***
- ***20 μVpp to 40 Vpp Output Range***
- ***Linear and Log Frequency Sweeps***
- ***1 mHz Frequency Resolution***
- ***<-100 dBc Distortion (to 20 kHz)***
- ***25 ppm Frequency Accuracy***
- ***BNC, XLR, and Banana Connectors***
- ***Balanced and Unbalanced Outputs***
- ***Standard RS-232, GPIB interfaces***

DS360 Overview

The DS360 is the world's first ultra-low distortion synthesized function generator. Unlike conventional low distortion generators which use RC oscillators, the DS360 uses a precision digitally synthesized source, providing you with precise frequency control, sine, two tone, square wave, and noise waveforms, computer control capability, **and** distortion as low as -100 dBc. Because of this unique combination of features, the DS360 is ideally suited for a variety of audio testing applications where a low distortion source is required.

Function and Sweep Generator

Unlike many low distortion oscillators which provide only a sine output, the DS360 has the output flexibility associated with conventional synthesized function generators. Sine waves and squarewaves from 1 mHz to 200 kHz can be generated as well as white and pink noise waveforms. Standard two tone output signals include SMTPE, DIM and CCIF. Three standard output connectors, BNC, XLR, and banana jacks, allow connection to virtually any type of test setup.

Computer Interfaces Included

Unlike RC oscillators, the DS360 includes standard GPIB and RS-232 computer interfaces allowing you to automate test procedures from any laboratory computer. The interfaces allow you to set and read the status of any instrument function.

DS360 Features

Functions

The DS360 synthesizes low distortion sine, square, and two tone waveforms with frequencies from 1 mHz to 200 kHz. The frequency resolution is 1 mHz or 6 decimal digits. The frequency accuracy is 25 ppm over the entire frequency range — far better than conventional analog ultra-low distortion oscillators. Frequencies can be simply entered on the front panel keypad or by modifying the existing entry with the knob. In addition to the standard function waveforms the DS360 includes a noise generator capable of providing white, pink, and bandlimited white noise. Two tone outputs can be specified as a combination of two sine waves or a sine wave plus a square wave to create standard SMPTE, DIM, and CCIF two tone signals.

Outputs

Outputs can be selected as either balanced or unbalanced. The front panel provides two floating BNC connectors, an XLR jack and four banana jacks to simplify interconnections in almost any application The unbalanced outputs can provide amplitudes from 2 µVrms to 5.09 Vrms into a 50Ω load, 2 µVrms to 7.071 Vrms into a 600Ω load, or 4 µVrms to 14.14 Vrms into a high impedance load. The balanced outputs provide twice the output amplitude range of the unbalanced outputs. Output amplitudes can be specified in Vpp, Vrms, dBV, dBm, or dB relative to a user specified level.

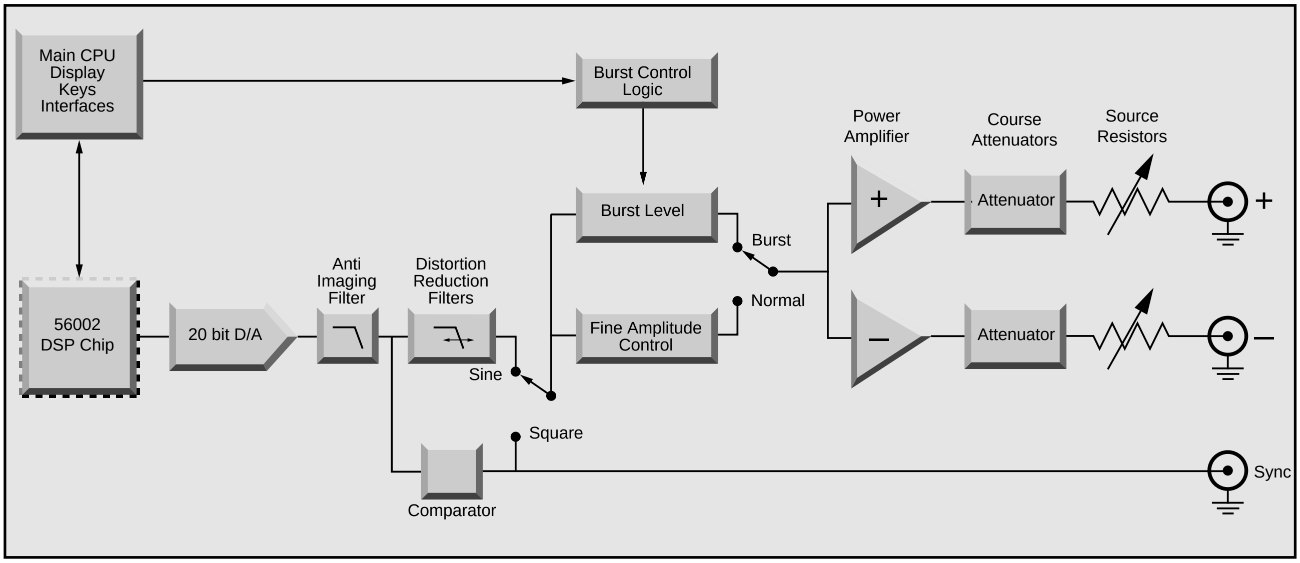
Sweeps and Bursts

The DS360, unlike single frequency analog low distortion oscillators, is also a flexible, low distortion, precision sweep generator. Sweeps can be programmed anywhere within the instrument's frequency range with sweep lengths ranging from 0.3 ms to 100s. Sweeps can be single shot or repetitive, and the instrument can sweep both up and down in frequency. The DS360's synthesized architecture gives it excellent flatness characteristics while sweeping: .5% to 20 kHz, and 1% to 100 kHz. Sweeps may be internally or externally triggered, and a rear panel TTL sweep marker is provided to synchronize external equipment to the beginning of the sweep.

The DS360 also includes a number of flexible burst options. Burst length can be specified in number of cycles from .5 to 65535, or the burst rate can be specified with an external TTL signal. The burst may be internally or externally triggered. Burst signals switch only at zero crossings to minimize high frequency artifacts.

Computer Control

The DS360 includes standard RS232 and GPIB computer interfaces. All instrument functions can be set and queried via the computer interface.



DS360 Block Diagram

Output Waveforms

Sine Wave:	
Frequency	0.001 Hz to 200.000 kHz
THD, Unbalanced	-100 dB to 20 kHz
	-85 dB to 100 kHz
THD, Balanced	-97 dB to 20 kHz
	-82 dB to 100 kHz
Square Wave:	
Frequency	0.001 Hz to 200 kHz
Rise Time	800 ns
White Noise:	
Bandwidth	DC to 200 kHz
Flatness	< 1.0 dB pk-pk, 1Hz to 100kHz
Crest Factor	11 dB
Pink Noise:	
Bandwidth	10 Hz to 200 kHz
Flatness	< 2.0 dB pk-pk, 20 Hz - 20 kHz (measured using 1/3 octave analysis)
Crest Factor	12 dB
Bandwidth Limited Noise:	
Bandwidths	100 Hz, 200 Hz, 400 Hz, 800 Hz, 1.6 kHz, 3.2 kHz, 6.4 kHz, 12.8 kHz, 25.6 kHz, 51.2 kHz, 102.4 kHz
Center Frequency	0 Hz to 200.0 kHz, 200 Hz increments
Flatness (in band)	< 1.0 dB pk-pk
Crest Factor:	
Base Band (0Hz Freq)	12 dB
Non Base Band	15 dB
Two-Tone:	
Type	Sine-Sine, Sine-Square
Sine Frequency	0.001 Hz to 200.000 kHz
Square Frequency	0.1 Hz to 5.0 kHz
Square Resolution	2 digits
SFDR	>90 dB
Sine or Square Burst:	
ON Cycles	1/2, 1 to 65534 cycles
Repetition Rate	1 to 65535 cycles
Triggering	Internal, External, Single, Externally Gated
OFF Level	0.0% - 100.0% (of ON Level)
White or Pink Noise Bursts:	
ON Time	10µs - 599.9s
Repetition Time	20µs - 600s
Triggering	Internal, External, Single, Externally Gated
OFF Level	0.0% - 100.0% (of ON Level)
Sine or Square Sweeps:	
Type	Linear or Logarithmic
Range	0.001 Hz to 200.000 kHz
Rate	0.01 Hz to 3.1 kHz
Resolution	2 digits
Flatness	±0.1 dB (1%)

Frequency

Resolution	6 digits or 1 mHz, whichever is larger (unless otherwise specified)
Accuracy	25 ppm (0.0025%) from 20° to 40° C

Output Amplitude

Unbalanced Outputs:	
50Ω Load	5.0 µVpp - 14.4 Vpp
600Ω Load	5.0 µVpp - 20.0 Vpp
Hi-Z Load	10.0 µVpp - 40.0 Vpp
Balanced Outputs:	
50Ω Load	10 µVpp - 28.8 Vpp
150Ω Load	10 µVpp - 28.8 Vpp
600Ω Load	10 µVpp - 40.0 Vpp
Hi-Z Load	20 µVpp - 80.0 Vpp
Resolution:	
Vpp or Vrms	4 digits or 1µV, whichever is greater
dBm or dBV	0.1dB
Accuracy	± 0.1 dB (1%)

Noise

Broadband Noise (into Hi-Z load, 1.0 kHz):	
40 Vpp - 1.26 Vpp	<150 nV/√Hz
1.26 Vpp - 126 mVpp	<10 nV/√Hz
126 mVpp - 12.6 mVpp	<7.5 nV/√Hz
<12.6 mVpp	<4 nV/√Hz

Offset

Unbalanced Output:	
50Ω Load	0 - ± 7.4 VDC
600Ω Load	0 - ±10.0 VDC
Hi-Z Load	0 - ±20.0 VDC
Balanced Output	Not Active
Resolution	3 digits
Accuracy	1%

Outputs

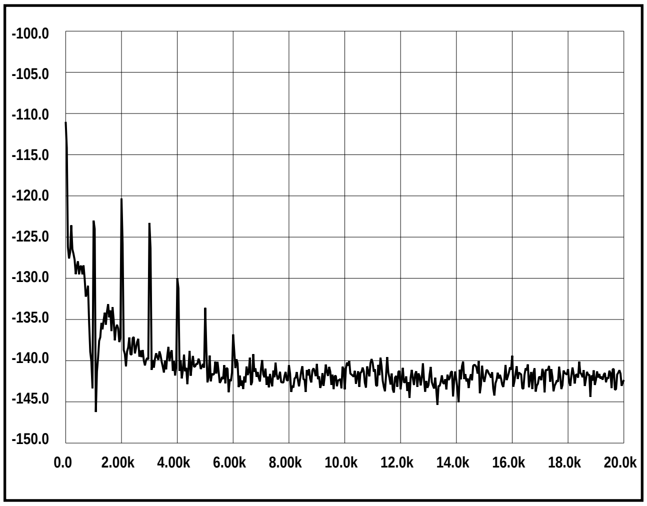
Configuration	Balanced and Unbalanced
Connectors	Floating BNCs, banana plugs and XLR Jack
Source Impedance:,	
Balanced Output:	50Ω± 3%, 150Ω ± 2%, 600Ω ± 1%, Hi-Z (50Ω ± 2Ω)
Unbalance Outputs:	50Ω ±3%, 600Ω ±1%, Hi-Z (25Ω ±1Ω)
Maximum Floating Voltage	± 40 VDC

Other Outputs

Sync	TTL squarewave (same frequency and phase as output)
Burst Out	TTL pulse marks burst (TTL high for ON time)
Trigger/Gate In	TTL pulse starts sweep or burst. TTL HI activates gated burst.
Sweep	TTL pulse marks beginning of sweep
Digital Outputs	AES-EBU (balanced XLR) and SPDIF/EIAJ (RCA and optical)

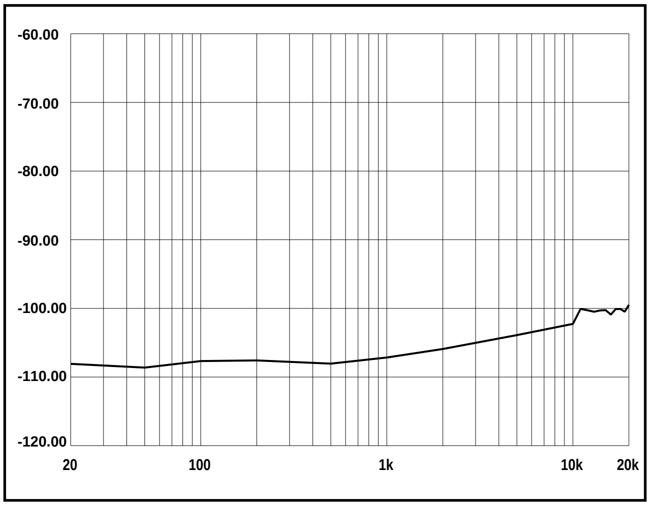
General

Computer Interface	GPIB and RS-232 standard.
Size	17"W x 3.5"H x 16.25"D
Weight	17 lbs.
Power	50 W, 100/120/220/240 VAC, 50,60 Hz
Warranty	One year parts and labor on any defects in material or workmanship.



THD vs. Frequency (Hz)

Residual distortion for a 1 kHz, 28 Vrms (balanced) sinewave after passing through a non-distorting notch filter to attenuate the fundamental



THD+N vs. Frequency (Hz)

DS360 Rear Panel

Ordering Information

DS360	**Ultra-low Distortion Function Generator**	**$ 2395**
O360RM	**Rack Mount**	**$ 85**

Digital Delay Generator

Model DG535 — Four Channel Digital Delay Generator

- ***Four Independent Delay Channels***
- ***Two Fully Defined Pulse Outputs***
- ***5 picosecond Edge Resolution***
- ***50 picosecond RMS Jitter***
- ***Fully Adjustable Amplitude and Offset***
- ***Delays up to 1000 seconds***
- ***1 MHz Maximum Trigger Rate***
- ***Internal, Burst, Line or External Triggering***
- ***Standard GPIB Interface***
- ***Optional Outputs: ±35 V, 100 ps Rise/Fall***

DG535 Overview

The DG535 is a very precise delay and pulse generator providing four precision delays or two independent pulses with 5 picoseconds resolution. Trigger to output jitter is less than 50 picoseconds. The high accuracy, precision, wide range, and low jitter of the DG535 makes it widely used in laser timing systems, automated testing, and precision pulse applications.

Delays and Triggers

The DG535 has four delay outputs and two pulse outputs. Each delay can be set from 0 to 1000 seconds relative to the trigger with 5 ps resolution. The two pulse outputs are defined by pairs of delay outputs. The first delay specifies the leading edge of the pulse, while the second delay defines the trailing edge or pulse width.

You can trigger the DG535 internally from 1 mHz to 1 MHz with four-digit frequency resolution. External, single-shot and burst mode triggers are also supported. For power control applications, the DG535 can be synchronized to the line voltage.

Flexible Outputs

Each output amplitude can be independently adjusted from -3 to 4 V with 10 mV resolution. For convenience, several preset output levels are also provided: TTL, ECL, and NIM. For applications requiring higher output voltages, optional rear-panel outputs provide the same precise timing at output amplitudes up to ±35 V.

Simple to Use

The DG535 is easy to use. All delay values can be entered numerically, or by convenient cursor keys. A GPIB interface is standard, allowing complete instrument control by an external computer. For quick configuration, the instrument stores up to nine complete setups in non-volatile memory.

DG535 Features

The DG535 Digital Delay and Pulse Generator can provide four precisely timed logic transitions, or two precisely controlled pulses. The four digitally controlled time intervals may be programmed from the front panel or via the GPIB Interface. Front panel BNCs provide high slew rate outputs for TTL, ECL, NIM or continuously adjustable levels. The outputs may be set to drive either 50Ω or high impedance loads. The high accuracy (1 ppm), precision (5 ps), wide range (0 to 1000 s), and low jitter (50 ps) make the DG535 the ideal solution to many difficult timing problems.

Delay Outputs

There are four delay output BNCs: A, B, C and D. The logic transitions at the outputs of A, B, C and D can be delayed by up to 1000 seconds in 5 ps increments relative to the T0 pulse. The T0 pulse, which marks the beginning of a timing cycle, is generated by the internal rate generator or in response to an external trigger. Insertion delay between the external trigger and the T0 pulse is 85 ns.

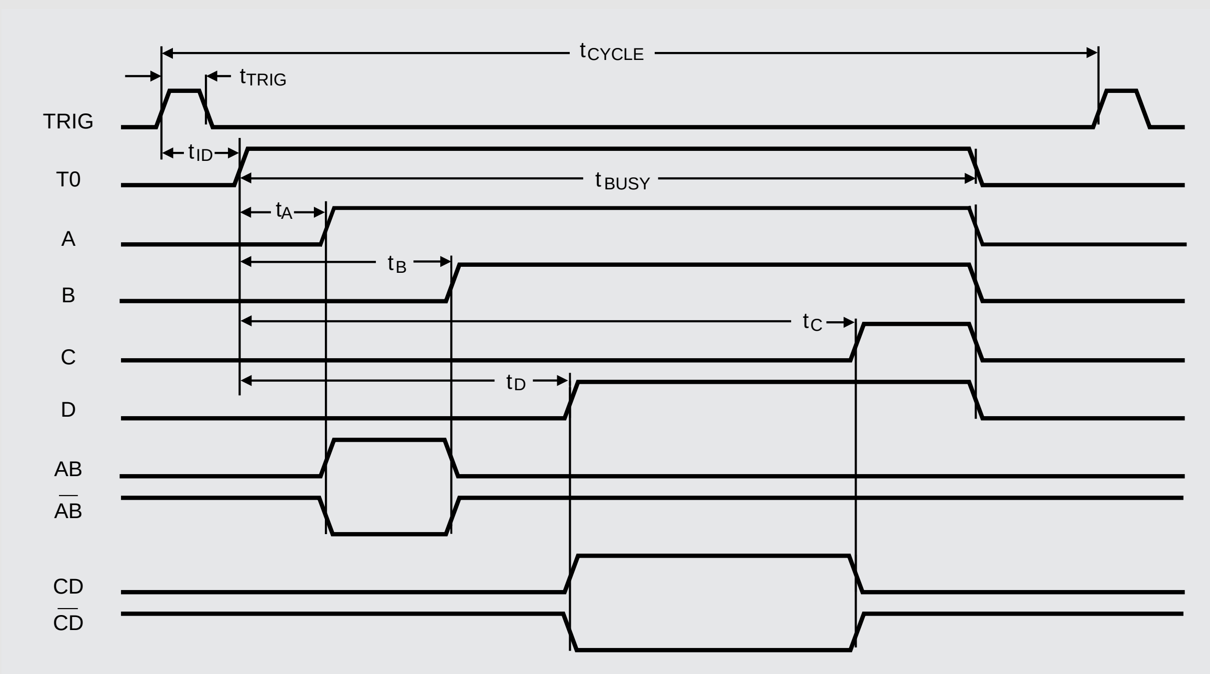
Delays for each channel may also be "linked" to another channel. For instance, you can specify the delays of the 4 channels as:

A=	**T_0**	**+**	**.00125000**
B=	**A**	**+**	**.00000005**
C=	**T_0**	**+**	**.10000000**
D=	**C**	**+**	**.00100000**

In this case, when the A delay is changed, the B output will move with it. This is useful, for instance, when A and B specify a pulse, and you want the pulse width to remain constant as the delay of the pulse is changed.

Each timing cycle is initiated by either an internal or an external trigger. The T0 output is asserted approximately 85 ns following the trigger. The other outputs, A, B, C and D, are asserted relative to T0 after their programmed delays. The delays may be specified with respect to T0 , or may be linked to another channel's delay. All of the delay outputs return low about 800 ns after the last delay is asserted. The pulse outputs, AB and CD, go high for the time interval between their corresponding delay channels.

T_{trig}	**> 5 ns**
t_{cycle}	**> 1 us + longest delay**
t_{ID}	**< 85 ns**
t_{BUSY}	**< 800 ns + longest delay**
$t_{A,B,C,D}$	**0 to 999.999 999 999 995 s**



Regardless of how the delay is specified each delay output will stay asserted until 800 ns after all delays have timed out. The delays will then become unasserted, and the unit will be ready to begin a new timing cycle.

Pulse Outputs

In addition to the four delay outputs there are four pulse output BNCs: AB, -AB, CD and -CD. The AB pulse output is asserted when the A delay times out and unasserted when the B delay times out. For instance, in the previous example, a 50 ns pulse would appear at the AB output and a 1 ms pulse at CD. Pulses as short as 4 ns (FWHM) can be generated in this manner. The complementary outputs (-AB and -CD) provide a pulse with identical timing and inverted amplitude.

Flexible Output Control

Each delay and pulse output has an independently adjustable offset and amplitude which can be set between -3 V and 4 V with 10 mV resolution. The maximum transition for each output is limited to 4 V. In addition, you can also separately select 50Ω or High Impedance termination for each output. For convenience, preset levels corresponding to standard logic families can also be selected. TTL, NIM, and ECL levels can all be selected with a single keypress.

Optional Inputs and Outputs

For applications requiring higher voltages, a rear-panel high voltage option is available. This option provides five rear-panel BNCs which output an amplified 1 µs pulse at the transition time of the front-panel T0, A, B, C, and D outputs. The high voltage option does not affect the function or the timing of the front panel outputs. The amplitude of the rear-panel outputs is 8x the corresponding front-panel output, and the outputs are designed to drive 50Ω loads. Since these outputs can only drive an average current of 0.8 mA, charging and discharging the cable capacitance may be the most important current limiting factor to consider when using them (assuming a high impedance load). In this case, the average current is: $I = 2Vtf / Z$, where V is the pulse step size, t is the length of the cable in time (5 ns/meter for RG-58), f is the pulse repetition rate, and Z is the cable's characteristic impedance (50Ω for RG-58) An opitional trigger inhibit input allows you to enable or disable external triggering with an external TTL signal.

Rear Panel of DG535 With Optional Outputs

Internal and External Timebases

Both internal and external references may be used as the timebase for the DG535. The internal timebase can be either the standard 25 ppm crystal oscillator timebase, or the optional 1 ppm Temperature Compensated Crystal Oscillator (TCXO). The internal timebase is available as a 1 Vpp square wave on the rear panel BNC. This output is capable of driving a 50Ω load, and can be used to provide a master timebase to other delay generators. Any external 10.0 MHz (±1%) reference signal with a 1 Vpp amplitude can also be used as an external timebase. For instance, the FS700 LORAN-C Frequency Standard can be used with the DG535 to provide a timebase with 1 part in 10^{12} long term stability.

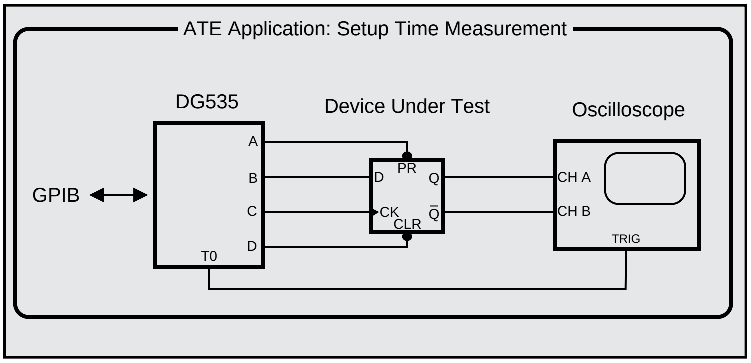
Easy to Use, Easy to Program

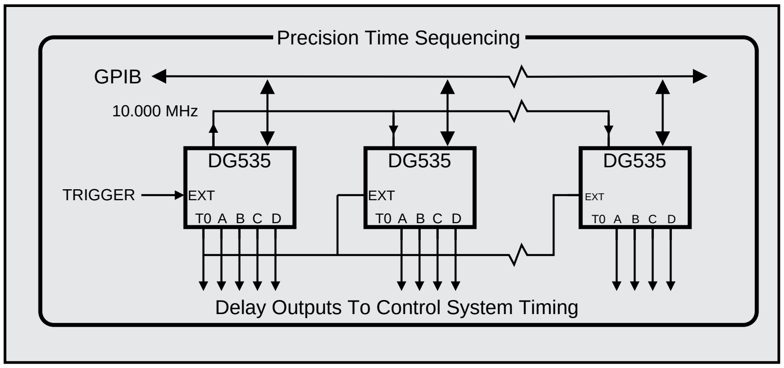
All instrument functions can be accessed through a simple, intuitive, menu-based interface. Delays can be entered with the numeric keypad, in either fixed-point or exponential notation, or by using the cursor keys to select and change individual digits. The 20 character backlit LCD display makes it easy to view delay settings in all lighting conditions.

The DG535 comes standard with a GPIB (IEEE-488) interface. All instrument functions can be queried and set via the interface. You can even display the characters the DG535 has received over the interface on the front-panel LCD display. This can be enormously valuable when debugging programs which send commands to the instrument.

ATE Applications

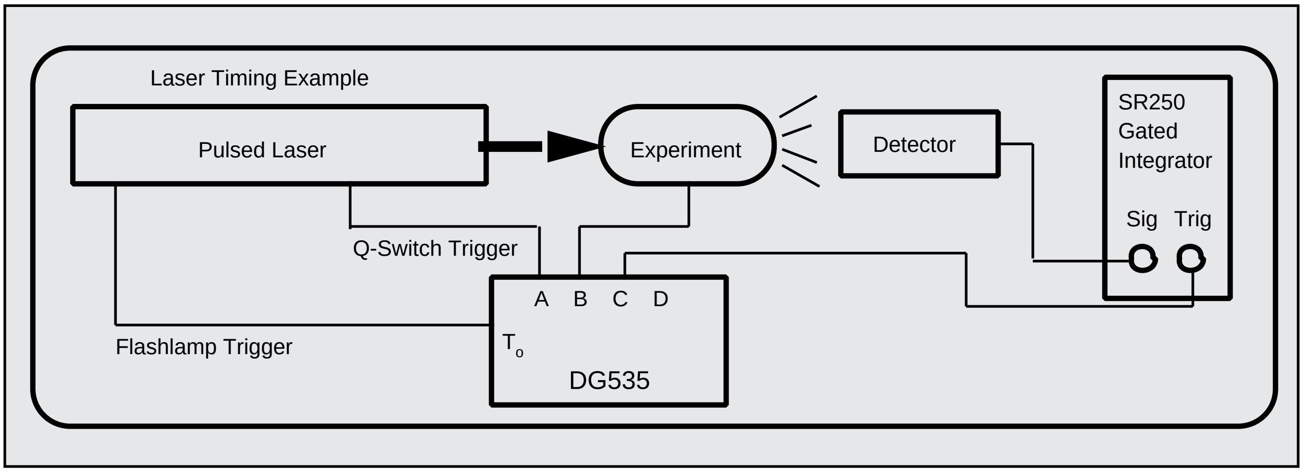
The DG535's versatility, precision, and accuracy recommend it for a wide variety of test and measurement tasks. In this example, the DG535 is used to measure the setup times for the data, preset, and clear inputs to a flip-flop. The measurements may be made with picosecond resolution. The logic thresholds for the device under test may be measured using the DG535's adjustable output levels. All measurements may be controlled from the front panel or by a computer via the standard GPIB interface.





Precision Time Control

A single DG535 can provide four transitions for precise system timing. Several DG535's may be used if more channels are needed. The 10 MHz reference may be daisy-chained between units so that each DG535 in an experiment uses the same timebase. All of the units may be controlled over the same GPIB bus. The flexible output levels and simple architecture of the pulse/delay generators make it simple and easy to rapidly reconfigure test systems.



Laser Timing Applications

The DG535's 4 independent outputs make it ideal for laser timing applications. In the above example the T_0 output of the DG535 fires the flashlamp of a pulsed laser. The DG535's internal rate generator controls the repetition rate of the laser and the overall experimental repetition rate. The A delay output controls the firing of the laser Q-Switch. The B delay output can be used to synchronize some aspect of the experiment to the laser pulse, e.g. the application of a voltage pulse, or the triggering of a discharge. Finally, the C delay is used in this example to trigger the gated integrator looking at the detector output. Note that both the B and C delays in this example can be specified relative to the A delay. In this way, as the laser pulse is moved by changing the A to T0 delay, the experimental trigger and the gated integrator trigger will stay fixed relative to the laser pulse.

Option 04A 100 ps Risetime Module
Option 04B 100 ps Falltime Module
Option 04C Bias Tee

Fast Rise Time and Fall Time Modules

External in-line modules are available to reduce the rise or fall time of the DG535 outputs to 100 ps. These modules use step recovery diodes to speed up the rise time (option 04A) or the fall time (option 04B). The bias tee (option 04C) allows these modules to be used with the optional rear-panel outputs to produce steps up to 15 V. Applications include time domain reflectometry measurements, pulse response measurements of fast amplifiers, testing high speed digital circuits, or use as a low jitter trigger source in high EMI environments.

The devices consist of a step recovery diode and matching network mounted in an in-line package. The units provide a fast, low distortion step into a 50Ω line with adjustable amplitudes from 0.5 V to 2.0 V, and up to 15 V when used with high voltage rear panel outputs.

Operation

For step amplitudes of less than 2.0 V the fast transition time units should be attached directly to the front panel of the DG535.

The dc offset is critical to the operation of the device: the offset is used to forward bias the step recovery diode (SRD) prior to the pulse output from the DG535. When the pulse from the DG535 begins, the stored carriers in the SRD maintain the conduction in the diode, shunting the output pulse to ground. When the stored carriers are depleted (about 3 ns after the start of the pulse), the diode abruptly stops conduction, creating a very fast transition time step at the output.

The offset must be increased when the output amplitude is increased and adjusted for the best output pulse shape. If the offset is set too high, the output step will overshoot: if the offset is too small, the output step will undershoot the final value.

Output Steps Up To 15 Volts

The fast rise time and fast fall time units may be used with the high voltage rear panel outputs (Option 02) to generate step sizes up to 15 V. A bias tee, Option 04C, is required for this mode of operation.

The high voltage rear panel outputs are AC coupled. Therefore some accommodation must be made to provide a DC current to forward bias the SRD prior to the output pulse. This current is applied via a "bias tee" (Option 04C) which passes the bias current through an inductor to the diode. The same inductor prevents the pulse from the rear panel output from passing to the bias source. A front panel output may be used as the bias source.

Overview

Option 04A: Fast Risetime

Output Amplitude	+0.5 to 2.0 Vdc
Output Offset	-0.8 Vdc, typical
Transition Time:	
Rise (20/80%)	100 ps, max.
Fall (20/80%)	2000 ps, max.
Pulse Aberrations	
Foot	4%, typ.
Ring	±5%, typ.

Option 04B: Fast Falltime

Output Amplitude	-0.5 to -2.0 Vdc
Output Offset	+0.8 Vdc, typ.
Transition Time:	
Rise (20/80%)	2500 ps, max.
Fall (20/80%)	100 ps, max.
Pulse Aberrations	
Foot	4%, typ.
Ring	±5%, typ.

The DG535 Digital Delay and Pulse generator can provide four precisely timed logic transitions, or two precisely controlled pulse outputs. The four digitally controlled time intervals maybe programmed from the front panel or via the GPIB interface with 5 picosecond resolution. Triggers are provided externally or from the internal rate generator at rates up to 1 MHz. The multiple trigger modes include internal, external, line, burst and single shot capabilities. The output can drive up to 4 V into 50 Ω impedances. An optional timebase improves the unit's accuracy and optional high voltage outputs provide up to 35 V from back panel BNCs.

Delays

Channels	Four independent delay outputs
Range	0 to 999.999,999,999,995 seconds
Resolution	5 ps
Accuracy	1500 ps + Timebase Error x Delay
Timebase	Standard: 25 ppm crystal oscillator Optional: 1 ppm TCXO (Opt. 03) External: User provided 10.0 MHz reference
RMS Jitter:	
T0 to any output	< 50 ps + delay x 10^{-8}
Ext. Trig to any output	< 60ps + delay x 10^{-8}
Trig Delay	External Trigger to T0 output : 85 ns

External Trigger

Rate	DC to 1/(1 µs + longest delay)
Threshold	±2.56 Vdc
Slope	Trigger on rising or falling edge
Impedance	1 MΩ + 40 pF or 50 Ω

Internal Rate Generator

Rate	Single Shot, .001 Hz to 1.000 MHz, or Line
Resolution	Four digits, 0.001 Hz below 10 Hz
Accuracy	Same as timebase
Jitter	1:10,000
Settling	< 2 seconds for any rate change
Burst Mode	2 to 32766 pulses per burst at integer multiples (4 to 32767) of the trigger period

Outputs

Load	50 Ω or high impedance
Risetime	3 ns Typical
Slew Rate	1 Volt / ns
Overshoot	< 100mV + 10% of pulse amplitude
Levels	TTL: 0 to 4 Vdc, normal or inverted ECL:-1.8 to -0.8 Vdc, normal or inverted NIM: -0.8 to 0.0 Vdc, normal or inverted VAR: Adjustable offset and amplitude between -3 and +4 Vdc with 10 mV resolution. 4V maximum transition.
Accuracy	50 mV + 3% of pulse amplitude
Option 02	Rear panel 1 µs pulses corresponding to T0, A, B, C, D outputs with amplitudes x8 the front outputs.

Computer Interface

IEEE488. All instrument functions and settings may be controlled over the interface bus. Interface queue can be viewed from front panel.

General

Display	Electroluminescent backlit LCD
Dimensions	14" D x 8.5" W x 4.75" H
Weight	10 lbs
Power	70 W from 100, 120, 220, or 240 V, 50/60 Hz
Warranty	One year parts and labor on any defects in material or workmanship.

Ordering Information

DG535	**Delay Generator**	**$ 3850**
Option 02	**±35 V Outputs**	**$ 650**
Option 03	**1 ppm TXCO Timebase**	**$ 350**
Option 04A	**Fast Risetime Module**	**$ 250**
Option 04B	**Fast Falltime Module**	**$ 250**
Option 04C	**Bias Tee**	**$ 100**
Option 05	**Dual Rack Mount**	**$ 150**
Option 06	**Trigger Inhibit Input**	**$ 250**

Accuracy, Jitter, and Drift

Accuracy, jitter, and drift are three terms often used when discussing delay generators and time measurement equipment. Here we'll discuss what these terms mean, and how they relate to the performance of the DG535.

Jitter

Various noise sources in the DG535 modulate the time delay for the outputs causing jitter. Some of these noise sources are common to all channels, others are independent. The distribution of the pulses around the desired time can be approximated by a Gaussian distribution:

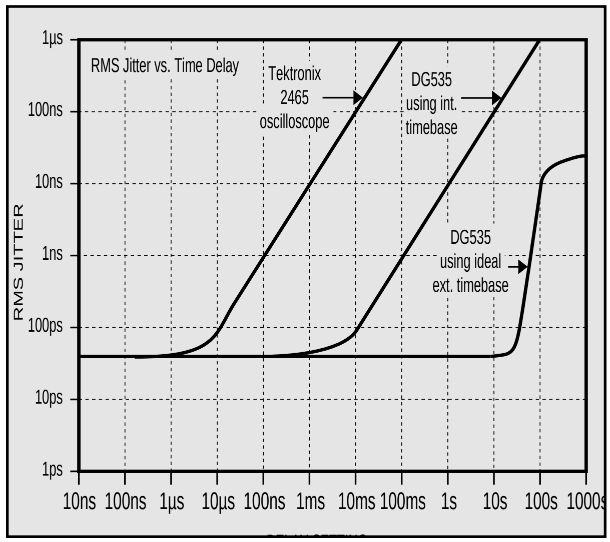
$$p(t) = \frac{1}{\sigma\sqrt{2\pi}} e^{-(t-T)^2/2\sigma^2}$$

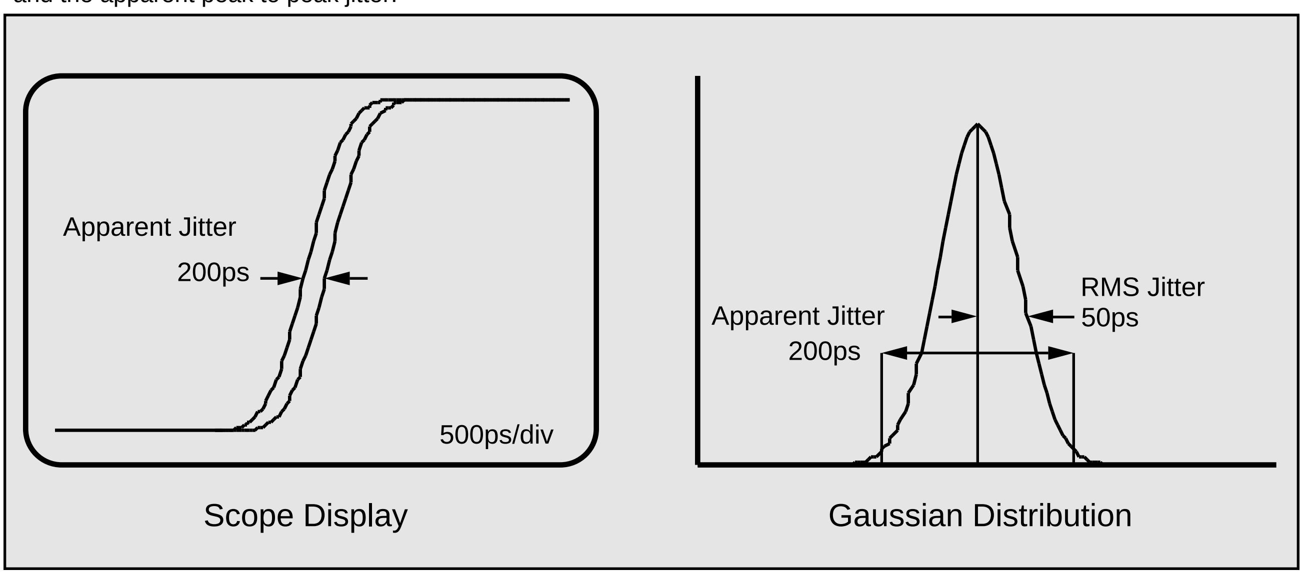
where:

p(t)	=	probability of pulse occurring at time t
T	=	time delay for the output (mean value)
σ	=	standard deviation of the distribution

The RMS jitter of the output is defined as σ, the standard deviation of the pulse delay distribution. In the DG535, the RMS jitter is a function of the delay setting. For delays less than 100 µs, the RMS jitter is 50 ps. At greater delays the jitter is about 10^{-8} of the delay setting. Note that on an oscilloscope display you will not see the RMS jitter but will instead observe the peak to peak jitter which is about 4 times worse. The diagram below illustrates the relation between the RMS jitter and the apparent peak to peak jitter.

Another factor to consider when trying to observe jitter on an oscilloscope is the jitter in the scope's timebase. For a good 350 MHz scope, the timebase jitter is typically 25 ps RMS + 10 ppm of the timebase setting. To show how this affects the apparent jitter seen on a scope, the diagram below plots the RMS jitter vs. delay for a Tektronix 2465 oscilloscope, the DG535 with it's standard internal timebase, and the DG535 with an ideal, jitter-free timebase. Note that for delays above 1µs, the DG535's jitter is swamped by that of the scope timebase. The bottom line is that it is difficult to accurately measure small amounts of jitter with an oscilloscope. The only accurate way to measure jitter is with a time interval counter, such as the SR620.





Accuracy

The accuracy of a delay generator is defined as the difference between the mean value of the pulse probability distribution and the nominal front panel delay setting. In the DG535, the error in the time delay between any two outputs can be expressed as:

Error = 1.5 ns + Timebase Error x Δt

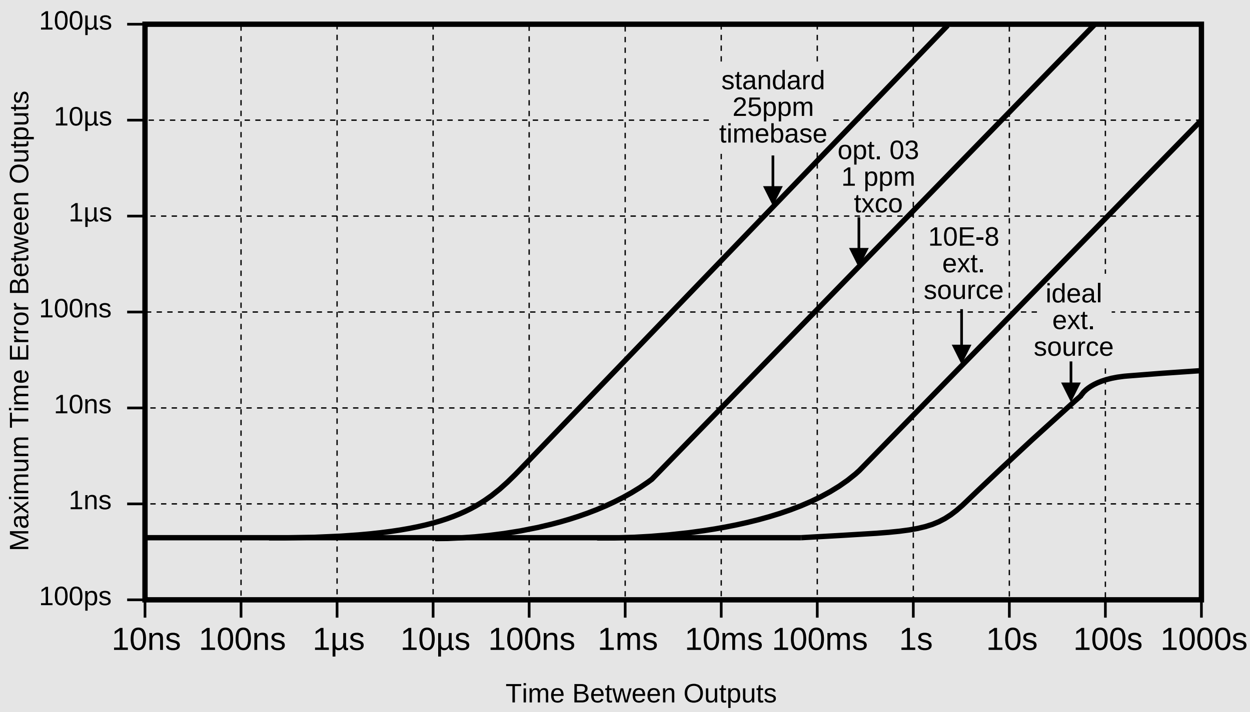
where Δt is the time delay between the two outputs. This error is exclusive of time shifts due to slew rates at the outputs. In other words, to accurately measure the error between two outputs they should be set to the same output levels and be driving the same load impedance. The timebase error term depends on which timebase is being used:

Timebase	Error
Standard	<25 ppm, 0-50 °C
Option 03	< 1 ppm, 0-50 °C
External	Source spec + 0.0002 ppm

For a time delay of 1.0 ms, this implies an absolute error of ±25 ns, ±2.5ns, and ±1.5ns respectively for the standard, optional and external timebases (assuming a 0.01 ppm external source specification). If A=100.000 µs and B=100.010 µs, the error with respect to T0 could be as large as 4.0 ns with the standard timebase, however the error of A with respect to B will be less than 1.5 ns. A graph showing the maximum time error as a function of time delay is shown below. The four curves show the time error for the standard, optional, 0.01 ppm external, and a perfect timebase. The excess error for time delays longer than 1 second on the ideal external source curve is due to drift in the analog jitter compensation circuits.

Drift

The drift of the timebase over several hours is substantially less (x10 to x100 less) than the absolute timebase error. The major factor in the timebase drift is the instrument's temperature: after the instrument is warm, the timebase drift is about 0.5 ppm/°C for the standard timebase, and about 0.05 ppm/°C for the optional timebase. The drift between several delay generators in the same experiment may be eliminated by daisy-chaining the reference output from one DG535 to the reference input on the other unit.



Time Interval Counter

Model SR620 — 25 ps, 1.3 GHz Universal Counter/Timer

- ***25 ps Single-Shot Time Resolution***
- ***1.3 GHz Maximum Frequency Measurement***
- ***1 nHz Frequency Resolution***
- ***0.001 Degree Phase Resolution***
- ***Statistical Analysis & Allan Variance***
- ***Sample Size from 1 to 1 million***
- ***Graphical Output to XY Scopes***
- ***Hardcopy to Printers and Plotters***
- ***GPIB and RS-232 Interfaces***
- ***Optional Ovenized Timebase***

SR620 Overview

The SR620 Time Interval Counter performs practically all the time and frequency measurements required in the laboratory or ATE environment. The instrument's high single-shot timing resolution, low jitter, and outstanding flexibility make it the counter of choice for almost any application.

Time Measurements

You can measure time intervals with 25 ps rms resolution, making the SR620 one of the highest resolution counters available. Time intervals up to ±1000 seconds can be measured with 50 ps relative accuracy, and 500 ps absolute accuracy.

Measurement Flexibility

The SR620 also measures frequency (up to 1.3 GHz), pulse-width, rise and fall time, period, and phase. Frequency can be measured with gates from 1 µs to 500 s, making the SR620 suitable for applications as varied as the measurement of short-term phase locked loop jitter, to the study of the long-term drift of atomic clocks. Up to 11 digits of frequency resolution for a 1 s measurement are provided. All measurement modes are supported by a wide variety of flexible arming and triggering options.

Complete Statistical Calculations

Statistics are automatically calculated on sample sizes from one to one million. The mean, standard deviation, Allan variance, minimum or maximum, can all be displayed on the front panel or quickly dumped to your computer via the standard GPIB and RS-232 interfaces.

Graphics and Hardcopy

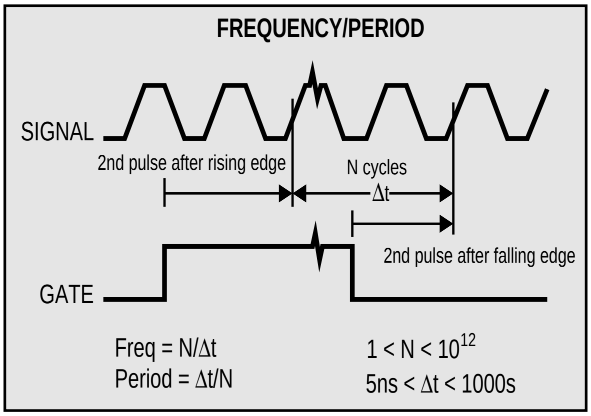
Unlike conventional counters that only have numeric displays, the SR620 provides live, graphical displays of measurement results. Histograms and stripcharts can be displayed on any oscilloscope with an X-axis input, or output to HP-GL compatible plotters and dot-matrix printers. Convenient on-screen cursors and display annotation facilitate the analysis of measured data.

SR620 Features

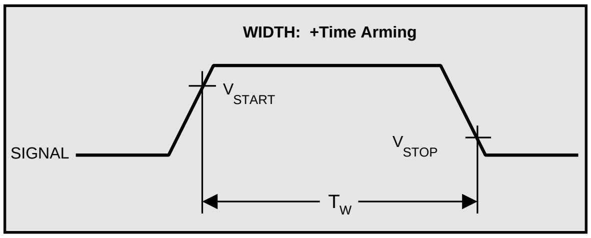
Measurement Modes

Time: In its most basic measurement mode, the SR620 measures the time interval between a start and a stop pulse. Either of the SR620's two inputs, or its REF output may be selected as the source of start and stop pulses. Internal and external gating signals can be used to holdoff the acceptance of either start or stop pulses. The SR620 can make both positive time measurements, (in which the stop pulse follows the start pulse), or negative time measurements (in which the stop pulse occurs before the start pulse).

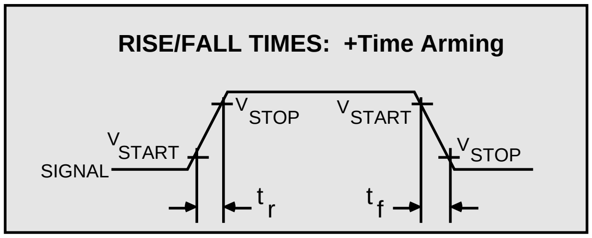
Frequency: The SR620 measures frequency by the reciprocal frequency counting technique. In other words, the instrument measures the time interval for some integer number of input cycles, then computes frequency by dividing the number of cycles by the time interval. Since no fractional cycle measurements are involved (as would be the case if the instrument measured the number of cycles in a fixed time interval), extremely high frequency resolution can be achieved–up to 11 digits for a 1 second gate. The diagram below illustrates this method of computing frequency. Fixed and scanning internal gates are provided, or the gate can be triggered externally, or delayed by an adjustable amount from an external trigger.



Pulse Width: The width of pulses at either input can be measured. Separate start and stop voltages can be selected for pulse width measurements. Resolution and accuracy are the same as time measurement mode.

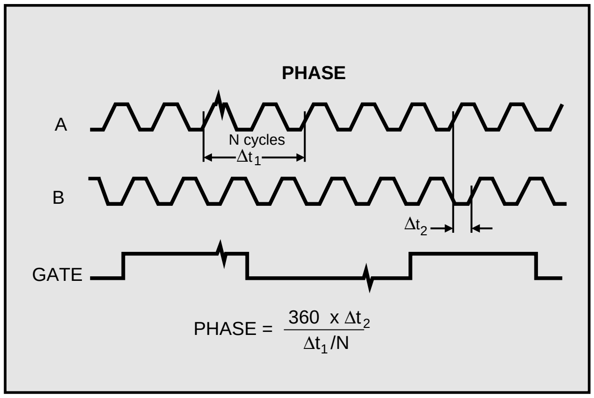


Transition Time: Rise and fall times of either input may be measured. Start and stop thresholds may be set between ±5 V with 10 mV resolution. The 350 MHz input bandwidth allows measurements of rise and fall times as small as 1 ns.



Period: The SR620 can also measure the period of waveforms. Period is measured similarly to frequency, but the reciprocal of frequency is computed and displayed.

Phase : The phase angle between signals on the A and B input can be measured with 0.001 degree resolution. You can measure the phase of signals (at the same frequency) from 0.001 Hz to 100 MHz in frequency. The counter actually makes two measurements: a frequency measurement of one channel, and a time measurement of the delay of the second channel with respect to the first. The phase is then computed as shown below.



Event Counting: The SR620 will also count transitions (events) at either of its inputs. As with all the other modes, event counting may be gated internally or externally, and both the voltage threshold and slope for a transition are adjustable. Event rates up to 300 MHz can be counted, with up to 12 digits of output. The unit also has a ratio mode which will compute the ratio of the number of events counted on the A and B inputs.

Reference Output

A precision, 1 kHz, 50% duty cycle square wave is available at the front panel REF output. The REF output can be used as a source of start or stop pulses for any of the SR620's measurement modes. For instance, the length of a cable connected between REF and the B input can be precisely determined by measuring the time delay between REF and B.

Statistical Calculations

The SR620 can make measurements on a single-shot basis, or calculate the statistics of a set of measurements. Sample sizes from 1 to one million can be selected. The SR620 will automatically calculate the mean, standard deviation or Allan variance, minimum, and maximum for each set of measurements.

Histograms and Strip Charts

The ability to generate live, graphical, output makes the SR620 stand out among time interval counters. Graphical output of measurement data is available in three formats: a histogram showing the distribution of values within a set of measurements, a stripchart of mean values from successive measurements, or a stripchart of jitter (standard deviation or Allan variance) values from successive measurements. A new histogram or stripchart point is generated after each set of measurements is completed. Up to 250 stripchart points, or histogram bins, can be displayed. Data for all three graphs is continuously saved, so you can view any of the graphs by cycling through selections on the front panel.

Both histograms and stripcharts can be displayed on any oscilloscope with an X-axis input (see the pictures below), or can be plotted on an HP-GL compatible plotter or dot-matrix printer. Convenient cursors allow you to read the value of any data point on the histogram or stripchart. Autoscale and zoom features make it simple to display all, or any portion, of the graphs.

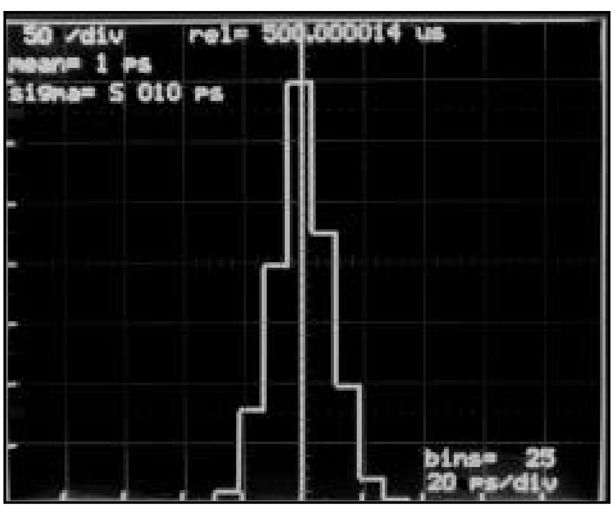
Built-in Autocalibration

A sophisticated, built-in, autocalibration routine nulls insertion delays between start and stop channels, and compensates for the differential nonlinearites inherent in analog time-measurement circuitry. The autocalibration routine takes about two minutes to perform and should be run every 1000 hours of operation.

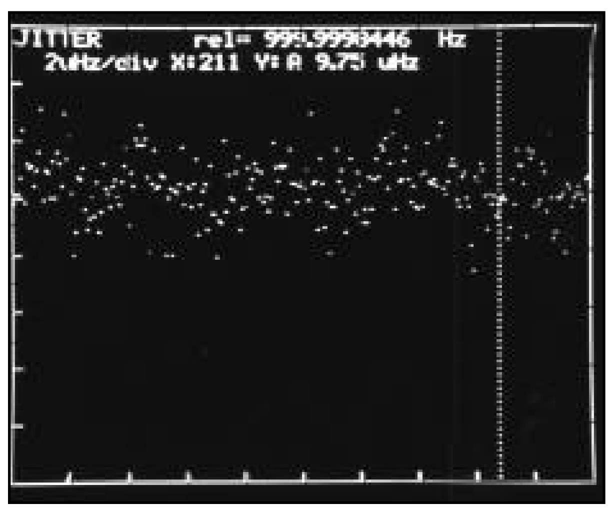
Choice of Timebases

The choice of timebase affects both the resolution and accuracy of measurements made with the SR620. (See Application Note #2 for a detailed discussion of this issue.) The SR620 lets you choose between an economical standard timebase with an aging coefficient of $1x10^{-6}$ /year, or an optional ovenized-oscillator timebase with only $5x10^{-10}$/day aging, and about an order of magnitude better short-term stability than the standard timebase. A rear-panel input lets you connect any external 5 MHz or 10 MHz source as a timebase.

Configuration Menus and Scanning



X-Y Oscilloscope Display of Histogram



X-Y Oscilloscope Display of Stripchart

Many measurement control parameters are typically set once and then rarely re-adjusted. These include the communications setup (GPIB and RS-232), the graphic output controls, the scan and D/A controls, and the self calibration menus. The SR620 lets you access these functions through convenient front-panel configuration menus. By scrolling through the different choices in each configuration menu, users can quickly and easily configure the instrument for their individual test conditions.

Computer Interfaces

Standard GPIB (IEEE-488) and RS-232 interfaces allow remote control of the instrument. All instrument functions and configuration menu settings are accessible via the interfaces. A fast binary dump mode outputs up to 1400 measurements per second to a computer. A parallel printer port allows direct printing from the instrument. Standard IEEE-488.2 communications are supported, and plotter outputs are provided in HP-GL format making interfacing simple and easy. For convenient debugging, the last 256 characters passed over the interfaces can be viewed on the front panel.

Rear Panel Features

Two rear-panel DVMs make measurements of dc voltages with 0.3% accuracy (up to 20 Volts). These values may be read via the interfaces or displayed directly on the front panel. Two rear panel outputs continuously provide voltages proportional to the mean and the jitter of the measurement sample. These 0 to 10V outputs can drive strip chart recorders, or can be set to provide fixed or scanned output voltages.

Application Information

Application Note #2, “Making Measurements with the SR620 Time Interval Counter,” contains a more detailed description of some of the SR620's measurement modes, as well as a discussion of accuracy and resolution issues.

Overview

SR620 Rear Panel

The SR620 is a full-featured time interval and frequency counter, capable of measuring frequency with 11 digits of resolution, or time intervals as small as 25 picoseconds. It also measures pulse width, rise/fall time, period, phase and event counts. Statistics are automatically calculated and reported on samples as large as 1 million, including: sample mean, maximum, minimum, and standard deviation or Allan variance. Standard GPIB and RS-232 interfaces control all front panel functions, and printers or plotters can be directly connected to the SR620.

Timebase

	Standard	Option 01
Frequency	10.000 MHz	10.000 MHz
Type	TCVCXO	Ovenized VCXO
Aging	$1x10^{-6}$/yr	$5x10^{-10}$/day
Allan Variance (1s)	$2x10^{-10}$	$5x10^{-11}$
Stability 0-50° C	1 ppm	0.005 ppm
Settability	0.01 ppm	0.001 ppm
External	User may supply 5 or 10 MHz timebase. 1 Volt nominal.	

Time Interval, Width, Rise and Fall Times

Range	-1000 to +1000 s in +/- TIME mode; -1 ns to +1000 s in all other modes
Trigger Rate	0 to 100 MHz
Display LSD	4 ps single sample, 1 ps with averaging
Resolution:	
Standard Timebase	$(((25 \text{ ps typ } [50 \text{ ps max}])^2 + (0.2 \text{ ppb x Interval})^2) / N)^{1/2}$ rms
Option 01	$(((25 \text{ ps typ } [50 \text{ ps max}])^2 + (0.05 \text{ ppb x Interval})^2) / N)^{1/2}$ rms (N = Sample size)
Error	< ±(500 ps typ[1 ns max] + Timebase Error x Interval + Trigger Error)
Relative Error	< ±(50 ps typ[100 ps max] + Timebase Error x Interval)
Arming Modes	+TIME Stop is armed by Start +TIME EXT Ext arms Start +TIME EXT HOFF Leading EXT edge arms Start, trailing EXT edge arms Stop. ±TIME Armed by Start/Stop pair ±TIME CMPL Armed by Stop/Start pair ±TIME EXT Armed by EXT input edge EXT arming may be internally delayed or scanned with respect to the EXT input in variable steps. The step size may be set in a 1,2,5 sequence from 1 µs to 10 ms. The maximum delay is 50,000 steps.
Display	16 digit fixed point with 1 ps LSD
Sample Rate	For a sample size of N, the total sample time is N x(800 µs + measured time interval) + Calculation time. The calculation time occurs only after N measurements are completed and varies from zero (N=1, no graphics, binary) to 5 ms (N=1, no graphics) to 10 ms (display mean or std dev) to 60 ms (histo)

Frequency

Range	0.001 Hz to 300 MHz via comparator inputs. 40 MHz to 1.3 GHz via internal UHF prescalers. RATIO A/B range: 10^{-9} to 10^{3}
Error	< ± ((100ps typ [350 ps max])/Gate + Timebase Error) x Frequency
Gates	External, 1 period, 1 µs to 500 s in 1,2,5 sequence. Gates may be externally triggered with no delay. Gates may be delayed relative to an EXT trigger. The delay from trigger may be set from 1 to 50,000 gate widths.
Display	16 digit fixed point with LSD = Freq x 4 ps/Gate . 1 µHz maximum resolution (1nHz with x1000 for frequencies <1MHz)

Period

Range	0 to 1000 seconds. RATIO A/B range: 10^{-9} to 10^{3}
Error	< ±((100 ps typ [350ps max])/Gate + Timebase Error) x Period
Gates	Same as frequency
Display	16 digit fixed point with LSD = 1 ps (1 fs with x1000 for periods < 1 second)

Phase

Definition	Phase = 360 x (Tb - Ta) / Period A
Range	-180 to +180 degrees (0 to 100 MHz frequency)
Resolution	(25 ps x frequency x 360 + 0.001) degree
Gate	0.01 seconds (1 period min.) for period measurement and 1 sample for time interval measurement. Period may also be measured using externally triggered internal gates as in frequency mode.
Error	< ±(1 ns x Frequency x 360 + 0.001)°

Counts

Range	10^{12} RATIO A/B range: 10^{-9} to 10^{3}
Count Rate	0 to 300 MHz
Gates	Same as frequency
Display	12 digits

Inputs

Threshold	-5.00 to +5.00 Vdc with 10 mV resolution
Accuracy	15 mV + 0.5% of setting
Sensitivity	see graph on next page
Autolevel	Threshold set between peak input excursions. (f>10 Hz, duty cycle >10^{-6})
Slope	Rising or falling edge
Impedance	(1 MΩ + 30 pf) or 50Ω. 50Ω termination has SWR < 2.5:1 from 0-1.3 GHz
Coupling	AC or DC. Ext is always dc coupled.

Input Noise	350 µVrms typical
Bandwidth	300 MHz BW provides 1.2 ns risetime
Prescaler	see graph below
Protection	100 V. 50Ω terminator is released if input exceeds ±5 Vpeak.

REF Output

Frequency	1.00 kHz (Accuracy same as timebase)
Rise/Fall	2 ns
Amplitude	TTL: 0 to 4 Vdc (2 Vdc into 50Ω) ECL: -1.8 to -0.8 Vdc into 50Ω

DVM Inputs

Full Scale	±1.999 or ±19.99 Vdc
Type	Sample & hold with successive approximation converter
Impedance	1 MΩ
Accuracy	0.3% of full scale
Speed	Formatted response in approximately 5 ms.

D/A Outputs

Full Scale	±10.00 Vdc
Resolution	5 mV
Impedance	< 1Ω
Default	Voltage proportional to Mean & Deviation
Accuracy	0.3% of full scale

Graphics

Scope	Two rear panel outputs to drive x-y scope
Displays	Histograms and strip charts of mean & jitter
X-axis	-5 to +5 V for 10 division deflection
Y-axis	-4 to +4 V for 8 division deflection
Resolution	250 (H) x 200 (V) pixels
Hardcopy	Centronics port to dot matrix printers. RS-232, IEEE-488 to HP-GL compatible digital plotters.

Interfaces

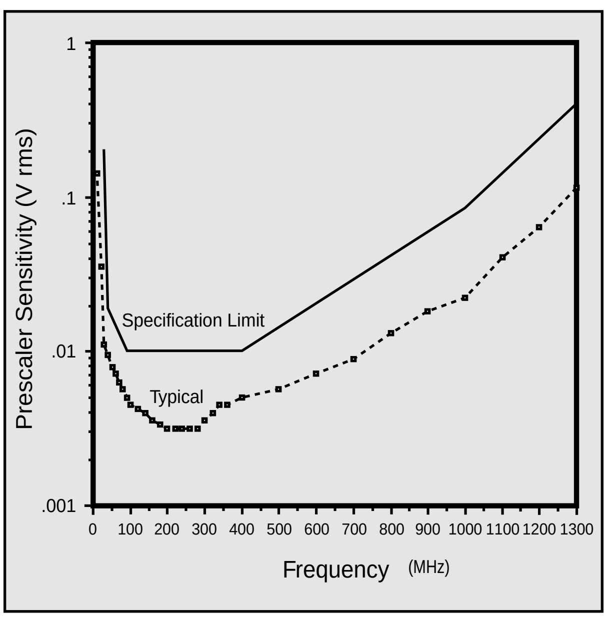
RS-232C	300 to 19.2 kBaud. All instrument functions may be controlled. PC compatible serial cable.
GPIB	IEEE-488 compatible interface. All instrument functions may be controlled.
Speed	Approximately 150 ASCII formatted responses per second.1400 binary responses per second.

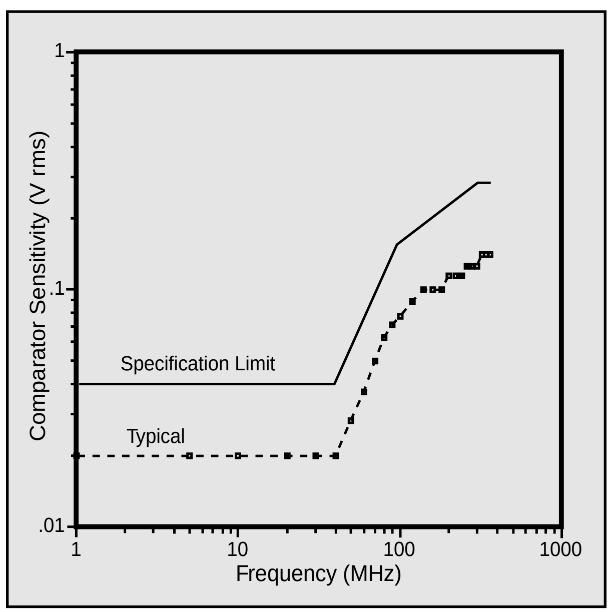
General

Operating	0 to 50° C
Power	100, 120, 220 or 240 VAC +5% -10%. 50/60 Hz. 70 W.
Dimensions	14" D x 14"W x 3.5"H. Rack mounting hardware included.
Weight	11 lbs
Warranty	1 year parts and labor on all defects in materials and workmanship

Ordering Information

SR620	**Time Interval Counter**	**$ 4950**
Option 01	**Ovenized Timebase**	**$ 950**





Frequency Counter

Model SR625 — 2 GHz Rubidium Stabilized Frequency Counter

- ***12 Digit Frequency Resolution***
- ***Rubidium Atomic Timebase***
- ***4x10^{-11} /day Drift***
- ***2 GHz Direct Prescaler Input***
- ***Short 10 Minute Warmup Period***
- ***Portable Operation***
- ***10 MHz Rb Timebase Output***
- ***GPIB and RS-232 Interfaces***

SR625 Overview

The SR625 Frequency Counter is a NIST traceable frequency counting standard for calibarating base stations, transmitters and many other types of communication systems. It combines the high resolution and wide variety of features found in the SR620 counter with the atomic accuracy of a Rubidium timebase.

Low Drift, High Accuracy

The SR625 Frequency Counter consists of a SR620 Time Interval Analyzer plus a high accuracy Rubidium timebase and 2 GHz input prescaler. The combination of the SR620 and the prescaler allow direct frequency measurements up to 2 GHz with twelve digits of resolution in a 100 second gate. The Rubidium timebase ensures both excellent short term stablity (only 3.16x10^{-11} 10s Allan Variance) and low long term drift (4x10^{-11}/day).

Simple, Portable Operation

The SR625's warmup time is less than ten minutes, making it ideal for field applications. An additional back panel output provides a Rubidium stablized 10 MHz signal which can be used to drive other test equipment such as synthesizers or spectrum analyzers. The standard GPIB and RS232 interfaces allow for complete control and data acquisition from any laboratory computer. The SR625's performance makes it the standard for remote applications or laboratory calibration.

Overview

The SR625 is a high-accuracy, high-precision Rubidium stabilized frequency counter. The SR625 combines a SR620 Time Interval Counter with a Rubidium timebase and a fast 2.2 GHz prescaler. The following specifications relate to the 2.2 GHz prescaler and the Rubidium timebase only. Please see the section on the SR620 for general specifications relating to the SR620 Time Interval Analyzer.

Rubidium Timebase

Frequency	10 MHz	
Accuracy at Shipment	± 5×10^{-11}	
One Day Stability	4×10^{-11}/day	
Long Term Drift	<5×10^{-11}/month	
	<5×10^{-10}/year	
Short Term Stability:		
	1s Allan Variance	1×10^{-10}
	10s Allan Variance	3.16×10^{-11}
	100s Allan Variance	1×10^{-11}
Warmup Interval	10 minutes to meet short term stability specification	
Power Consumption	70W (at warmup), 100/120/208/240V	
Output	10 MHz, 1 Vpp sinewave	

Prescaler

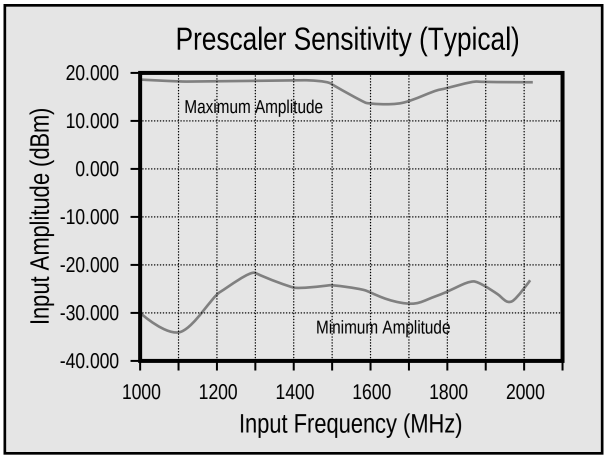
Frequency Ratio	10/1
Input Impedance	50Ω
Max. Input Level	+23 dBm
Input Freq. Range	50 MHz to 2 GHz
Input Sensitivity	See Graph

Output

Output Load	50Ω
Ouput Amplitude	700 mVpp square wave
Output Offset	500 mV

General

Size	3"H x 17"W x 14.5"D
Weight	15 lbs.



Ordering Information

SR625	**Frequency Counter**	**$14,850**
O625RF	**Upgrade SR620 to SR625**	**$10,350**

LORAN-C Frequency Standard

Model FS700 — 10 MHz High Stability Frequency Standard

- ***1 x 10^{-12} Long-term Stability***
- ***NIST Traceable***
- ***10^{-10} Short Term Stability (Optional 10^{-11})***
- ***Four 10 MHz Reference Outputs***
- ***Guaranteed Reception Throughout Most of the Northern Hemisphere***
- ***Phase Detector with Stripchart Output***
- ***0.01 Hz to 10 MHz TTL Output***

FS700 Overview

Many complex electronic systems require a stable, highly accurate timebase. Communication, automatic test and measurement, and precision time-measurement systems all require accurate frequency standards. Traditionally, users have looked to atomic clocks (cesium or rubidium) for high stability and accuracy. With the FS700 LORAN-C Frequency Standard, cesium-clock stability is now available at a fraction of the cost of atomic standards. The FS700 serves as a NIST traceable frequency reference in the U.S., Europe and Asia.

Cesium Stability

More than 50 LORAN transmitters are maintained throughout the Northern Hemisphere by the US Coast Guard. The timing of their transmissions is controlled by cesium clocks located at each transmitter site. The FS700 extracts the timing information from the transmitted signal and uses it to frequency-lock its own highly stable oscillator. The result is a 10 MHz output signal with the same stability as the cesium clock used to generate the LORAN transmissions.

Four 10 MHz outputs are available at the rear panel of the instrument, in addition to an adjustable frequency front-panel source. These outputs may used as timebase inputs to other laboratory instruments, such as frequency counters, synthesizers, spectrum analyzers, or pulse generators. The SR710 distribution amplifier can help you distribute the 10 MHz reference signal throughout your facility.

Phase Detector

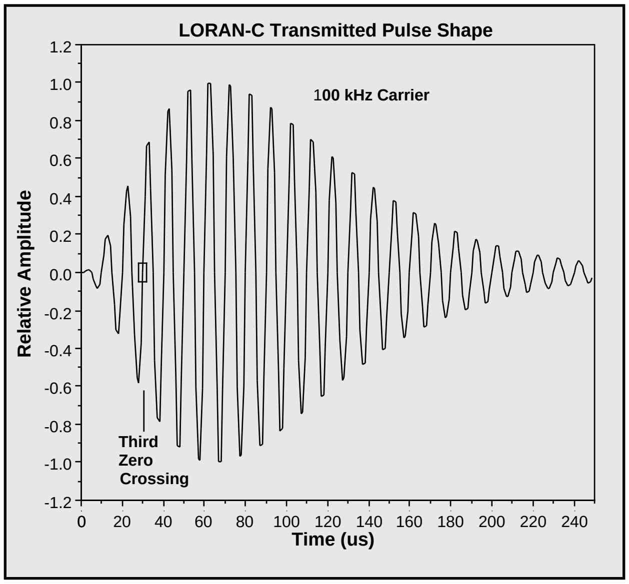
A built-in phase detector measures the phase shift between an external timebase and the internal frequency source. The FS700 can easily calibrate precision oscillators between 100 kHz and 10 MHz, providing a visual display as well as a voltage output proportional to phase difference. You can use the analog output to drive recorders for long-term frequency-stability testing or to phase-lock other sources to the FS700.

NIST Traceable

The timing information in the LORAN signal is constantly monitored by NIST. Detailed reports of the accuracy of each station and projected station downtime are available. With the FS700, you not only get the accuracy of a single atomic standard, but an entire network of redundant, continuously monitored standards, all at a price far below that of even a single atomic clock.

LORAN Operation

LORAN stations are divided into groups called chains, each of which broadcast LORAN pulses at a common repetition rate, known as the GRI (Group Repetition Interval). Once every GRI, each station in the chain transmits a group of eight LORAN pulses each having the pulse shape shown below:



It is the third zero-crossing of these pulses which are accurately controlled by a 10 MHz cesium standard at the LORAN transmitter station. The zero crossings of the 10 MHz standard are precisely synchronous with the zero-crossings of the received 100 kHz LORAN carrier signal in the ratio of 100:1. Thus, by locking to the third zero-crossing of the 100 kHz pulses, the FS700's internal oscillator maintains the long term stability of the transmitter's standard.

To operate the FS700 you need only enter the GRI for the chain nearest your location. The FS700 automatically adjusts its receiver gain and begins searching for LORAN pulses. You can set the FS700 to search for the station with the largest detected signal in the chain, or you can select a particular station. Once LORAN pulses have been identified, the FS700 begins an initial frequency lock to the selected station to remove any initial gross frequency offsets. When this is accomplished, the FS700 searches for the position of the third zero-crossing, and locks to it. The entire search process takes between 15 and 40 minutes depending on the signal-to-noise ratio of the selected station. Front panel readout of the signal-to-noise ratio for any station is available.

Continuously Monitored Output

Once locked, the FS700 minimizes the frequency difference between its internal oscillator and the LORAN transmission by use of a firmware frequency-locked-loop (FLL). Frequency-locked-loops have much better phase-noise and short term stability than phase locked loops, which can introduce large instantaneous frequency offsets while attempting to maintain zero phase difference. While locked, the FS700 continuously monitors the received signal for error conditions. If errors are found, the unit suspends frequency updates for 20 minutes. If the error condition is cleared by that time, the FLL is resumed, otherwise the FS700 can either wait for the selected station to come back on-line, or search for a new station.

Frequency Outputs and Phasemeter

In addition to the four 10-MHz rear-panel sine outputs, the FS700 has a front-panel adjustable-frequency TTL output Frequencies between 0.01 Hz and 10 MHz, in a 1, 2.5, 5 sequence, can be selected. The frequency source has the same accuracy as the 10 MHz outputs. A built-in phasemeter allows comparison of the frequency source with an external input. You can either display a bar-graph of the phase difference between the two signals, or the FS700 will compute the frequency difference between them by looking at the accumulated phase difference over a specified time interval. A front panel connector supplies a voltage output proportional to the measured phase, which can be used to drive recorders, or to phase-lock other sources to the FS700.

Antennas and Filters

The FS700 comes complete with an 8-foot long remote active whip antenna, which is easily mounted on the roof of a building. The antenna is capable of driving up to 1000 feet of coaxial cable to the receiver. The weatherproof antenna base contains an FET preamplifier, and a 100 kHz bandpass filter. An optional lightning protection module can be inserted between the antenna and the receiver to provide protection in adverse weather conditions. Six adjustable notch filters in the receiver allow the user to reject interfering signals for optimum reception.

FS710 Distribution Amplifier

Installations requiring remote outputs will benefit from the optional FS710 AGC distribution amplifier. This unit provides seven 10 MHz outputs from a single input at distances of up to 1 mile from the FS700.

Some Questions About LORAN-C

Is the FS700 a NIST traceable frequency standard, and how is traceability provided?

According to NIST Special Publication 250-29:

"LORAN-C is an NBS traceable frequency source because its signals are received by both the user and NBS. NBS continuously monitors the LORAN-C stations and compares the transmitted phase to the nation's frequency standard."

A monthly bulletin (Time and Frequency Bulletin) documents the availability of LORAN transmissions during the previous month and any transmission downtimes. It also details any projected outages in the near future. A daily bulletin of the accuracy of each LORAN station is also available through NIST.

Isn't WWVB a better standard because it emanates from NIST in Colorado?

No. WWVB is inferior to LORAN-C because of its transmission characteristics. WWVB systems receive a skywave which has been reflected off the ionosphere. Since the position of the ionosphere is a variable, WWVB is about an order of magnitude less stable than the ground wave received LORAN-C signal. WWVB does, however, provide UTC ticks which makes it a better choice for low accuracy time transfer (as opposed to high frequency stability). For a frequency standard with optimum long term stability, LORAN-C is clearly the best choice for both performance and cost.

What about system stability?

LORAN-C has a 99.5% availability target. Usually, when stations go down, they are absent for less than 20 minutes, during which the FS700 oven stabilized oscillator will maintain the unit within frequency specifications. The FS700 may be set to seek other stations in case the primary selection is unavailable.

No single frequency source can be trusted completely. For example, cesium clocks deplete in 3 to 7 years, virtually assuring a reference failure. This inherent failure mechanism can only be handled by comparison with other references, as is done at the LORAN stations. For critical applications, redundancy is mandatory.

What limits the resolution of my frequency measurements?

The FS700 will compute and display frequency offsets of other sources by measuring phase change between the source and the LORAN locked reference, and dividing by the duration of the measurement. Because of the 3 degree quantization resolution of the phase comparator, the resolution of frequency offsets will be limited to about one part in 10^{+9} for a one second measurement, or about one part in 10^{+11} for a 100 second measurement. Longer measurements are limited by the stability of the timebase.

The front panel BNC phase output is an analog output and therefore has no quantization error, so it may be used in more demanding applications. It has a resolution of 1 degree per 10 mV.

How can I get better short term stability?

Use option 01 (an SC Cut oven stabilized crystal oscillator), run with longer time constants, and select the strongest station in your area. Several customers are using the FS700 to stabilize rubidium references with a time constant of several hours. This combination will provide the best combination of short and long term stability with NIST traceability.

How can I get more detailed information?

More detailed information may be obtained from:

United States Naval Observatory
Time Services Division
34 Massachusetts Ave.
Washington, DC 20390
(202) 653-1507

LORAN-C station information may also be obtained from:

United States Coast Guard Headquarters
Washington, D.C. 20593
(202) 267-0283

Overview

The FS700 LORAN-C Frequency Standard produces a highly stable and accurate 10 MHz output by locking its oven stabilized crystal oscillator to LORAN-C radio transmissions. The FS700 system consists of a receiver containing amplifiers, filters, data acquisition circuitry, and a remote antenna. The long term stability of the FS700 is that of the transmitted standard, an atomic cesium source. An internal phase meter circuit allows for the precise frequency calibration of external oscillators. An optional distribution amplifier provides multiple outputs of the 10 MHz reference.

Receiver

Sensitivity	Will lock with signal-to-atmospheric noise level of -10dB or better
LORAN Output	Filtered and gain controlled antenna signal, typically 6V peak-to-peak
Station Search	All available stations pre-programmed, Auto-Seek finds and tracks strongest station
Notch Filters	6 adjustable -30 dB notch filters, 3 at 40 - 90 kHz, 3 at 110 - 220 kHz

Frequency

Frequency Stability:	
Long Term	10^{-12}, The same as LORAN-C transmitter Cesium Clock.
Short Term	10^{-10}, standard oscillator 10^{-11}, low phase noise option
Outputs	Four 10 MHz Outputs. 1 Vpp sine into 50 Ω
LOCK Output	rear-panel TTL indicates receiver lock.
Front Panel Output	TTL level output from .01 Hz to 10 MHz in a 1, 2.5, 5, sequence.

Internal Oscillator

	Standard	Option 01
Frequency	10.000 MHz	10.000 MHz
Type	AT Cut Ovenized	SC Cut Ovenized
Aging	$5x10^{-10}$/day	$5x10^{-10}$/day
Allan Variance (1s)	$5x10^{-11}$	$5x10^{-12}$
Stability 0-50° C	0.005 ppm	0.005 ppm
Phase Noise		-125dBc @ 10Hz
	-130 dBc @ 100 Hz	-155dBc @ 100Hz
		-165dBc @ 1kHz

Phasemeter

Frequency Output	0.01 Hz to 10 MHz in 1, 2.5, 5 sequence, TTL level. Can be 50Ω terminated.
Oscillator Input	1 kΩ, 0.5 V peak-to-peak minimum level. 5.0 Volts max.
Phase Output	0.01 V/degree, 0 to ±360°. Output proportional to phase difference between OSC IN and FREQUENCY OUTPUT for frequencies between 100 kHz and 10 MHz.

Computer Interface

GPIB	IEEE-488 compatible interface. All instrument functions may be controlled. GPIB interface is standard.
RS-232	Optional 300 to 19,200 baud DCE serial interface. All instrument functions may be controlled.

Antenna

Type	100 kHz active antenna, with 40 kHz bandwidth bandpass filter in base.
Height	102"
Material	Fiberglass Whip
Base	2" dia. x 7.5", PVC
Mounting	3/4" FIPT
Output	50Ω nominal, female BNC
Environmental	-40 TO 60°C, 0-100% RH
Lightning Protection Module:	
Surge	18,000 A IEEE 8/20 waveform (based on ANSI C62.1)
Frequency Range	DC to 30 MHz
Throughput Energy	< 16 µJ (based on 1 KV/nS waveform)
Insertion Loss	<.25 dB

General

Operating	0 to 50° C
Power	100, 120, 220 or 240 VAC +5% -10%. 50/60 Hz. 50 Watts.
Dimensions	17" x 17" x 3.5". Rack mounting hardware included.
Weight	14 lbs
Warranty	One year parts and labor on any defects in material or workmanship

Ordering Information

FS700	**LORAN Frequency Standard**	**$ 2950**
Option 01	**Low Phase-Noise Oscillator**	**$ 450**
Option 02	**RS-232 Interface**	**$ 350**
O700ANT	**Replacement Antenna**	**$ 250**
O700LNG	**Lightning Module**	**$ 100**
FS710	**Distribution Amplifier**	**$ 1000**

West Coast USA 99400 µs	Fallon, Nevada, USA George, Washington, USA Middletown, California, USA Searchlight, Nevada, USA
Canadian West Coast 59900 µs	Williams Lake, BC, Canada Shoal Cove, Alaska, USA George, Washington, USA Port Hardy, BC, Canada
North Central USA 82900 µs	Havre, Montana, USA Baudette, Minnesota, USA Gillette, Wyoming, USA Williams Lake, BC, Canada
South Central USA 96100 µs	Boise City, Oklahoma, USA Gillette, Wyoming, USA Searchlight, Nevada, USA Las Cruces, New Mexico, USA Raymondville, Texas, USA Grangeville, Louisiana, US
Great Lakes 89700 µs	Dana, Indiana, USA Malone, Florida, USA Seneca, New York, USA Baudette, Minnesota, USA Boise City, Oklahoma, USA
Southeast USA 79800 µs	Malone, Florida, USA Grangeville, Louisanna, USA Raymondsville, Texas, USA Jupiter, Florida, USA Carolina Beach, North Carolina
Northeast USA 99600 µs	Seneca, New York, USA Caribou, Maine, USA Nantucket, Massachuetts, USA Carolina Beach, North Carolina, Dana, Indiana, USA
Canadian East Coast 59300 µs	Caribou, Maine, USA Nantucket, Mass., USA Cape Race, Newfoundland, Fox Harbor, Labrador, Canada
Newfoundland Coast 72700 µs	Comfort Cove, Canada Cape Race, Canada Fox Harbor, Canada
Bo 70010 µs	Bo Norway Jan Mayen, Norway Berlevag, Norway
Ejde 90070 µs	Ejde, Faeroe Island, Denmark Jan Mayen, Norway Bo, Norway Vaerlandet, Norway Loop Head, Ireland
Leassay 67310 µs	Leassay, France Soustons, France Loop Head, Ireland Sylt, Germany
Sylt 74990 µs	Sylt, Germany Lessay, France Vaerlandet, Norway
Mediterranean Sea 79900 µs	Sellia Marina, Italy Lampedusa, Italy Estartit, Spain
Saudi Arabia North 70300 µs	Afif, SA Salwa, SA Al Hamasin, SA Ash Shaykh Humayd, SA Al Muwassam, SA
Saudi Arabia South 88300µs	Al Khamasin, SA Salwa, SA Afif, SA Ash Shaykh Humayd, SA Al Muwassam, SA
Western Russia 80000 µs	Bryansk, Russia Petrozavodsk, Russia Solnim, Russia Simferopol, Ukraine Syzran, Russia
Eastern Russia 79500 µs	Aleksandrovsk, Russia Petropavllo Russia Ussuriysk, Russia Kurilsk, Russia Ohotosk, Russia
East Asian 99300 µs	Pohang, Korea Kwang-Ju, Korea Gesashi, Okinawa Niijima, Japan Ussuriisk, Russia
China North Sea 74300 µs	Rongcheng, PRC Xuancheng, PRC Helong, PRC
China East Sea 83900 µs	Xuancheng, PRC Raoping, PRC Rongcheng, PRC
China South Sea 67800 µs 67300 µs	Hexian, PRC Raoping, PRC Chongzuo, PRC
Northwest Pacific 89300 µs	Niijima, Japan Gesashi, Okinawa Miamitorishima, Japan Tokatibutto, Hokkaido, Japan Pohang, Korea
Russia-American 59800 µs	Petropavlo, Russia Attu, Alaska Aleksandrovsk, Russia
North Pacific 99900 µs	Saint Paul, Pribilof Is, Alaska Attu, Alaska Point Clarence, Alaska Narrow Cape, Kodiak Is, Alaska
Gulf of Alaska 79600 µs	Tok, Alaska Narrow Cape, Kodiak Is, Alaska Shoal Cove, Alaska Port Clarence, Alaska
Bombay 60420 µs	Dhrangadhr, India Veraval, India Billamora, India
Calcutta 55430 µs	Balasore, India Diamond Harbour, India Patpur, India

LORAN-C Frequency Standard

Model FS-710 — 7 Channel AGC Distribution Amplifier

- ***Distributes FS700 Output***
- ***7 Output Channels***
- ***Drives up to One Mile of Cable***
- ***30 dB Gain***

FS710 Overview

For installations of the FS700 which require many outputs at a remote location from the instrument, the FS710 AGC Distribution Amplifier offers 7 independent outputs from one input. WIth an AGC circuit capable of adding up to 30 dB of gain, this amplifier can be used to provide the FS700's timing signal at locations as far as a mile from the FS700.

Ordering Information

FS710	**Distribution Amplifier**	**$ 1000**

FS710 Specifications

Input

Frequency	10 MHz ± 100 kHz
Type	Insulated BNC, Transformer Coupled
Level	50 mVpp to 5 Vpp
VSWR	<1.2 at 10 MHz

Output

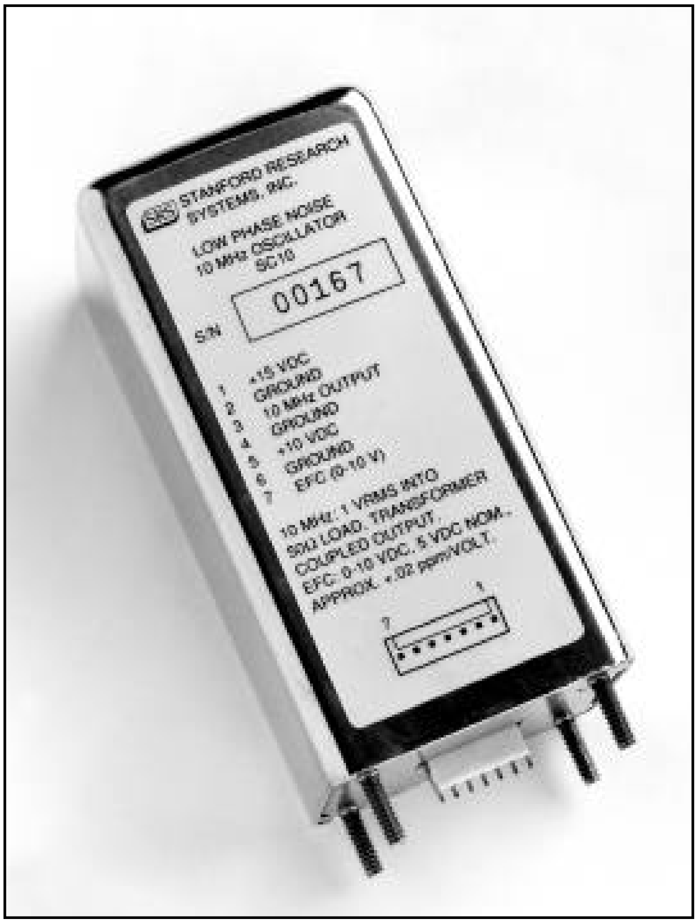
Type	7 Local Grounded BNCs
Level	1Vpp into 50Ω, ±10% 2Vpp into 10 kΩ, ±10%
Output VSWR	<1.2 at 10 MHz
Distortion	< -25 dBc

General

Power	100/120/220/240 VAC, 3W 50/60 Hz
Dimensions	7.75" x 7.5" x 2"
Weight	3 lbs.

High Stability Oscillator

Model SC10 — 10 MHz High Stability Ovenized Quartz Oscillator



- ***SC Cut Crystal for Lowest Phase Noise and 1-10s Allan Variance***
- ***Low Aging (<2x10^{-10}/day)***
- ***+15 or +24 Vdc Operation***
- ***Flexible Electronic Frequency Control***
- ***Accessibility to all components***
- ***Documented Circuit Schematic and Calibration Procedures***

SC10 Overview

The SC10 is a high stability ovenized 10 MHz quartz oscillator combining excellent phase noise, Allan variance, and aging characterstics. Using an SC cut crystal for lowest phase noise characteristics, and an innovative "electronic double oven" temperature controller to minimIze temperature gradients, the SC10 achieves a 1s Allan variance of 2x10^{-12}, and an aging rate of only 2x10^{-10} making it ideal for virtually any precision timing application.

Convenient Options

A number of options can be specified to match the SC10's performance to your requirements. +15 or +24 Vdc operation can be specified, and output is available on SMA, SMB, and SMC connectors, or on a single pin. Aging , noise, temperature stability and operating temperature range can all be separately specified in one of three grades so you only pay for the performance you need. The electronic fine tuning (EFC) is available with a number of tuning ranges and slopes.

Reliable Operation

Unlike the oscillators we have seen from several manufacturers, with a hand-built, short-run appearance, SC10 was designed from the beginning with reliable, repeatable, documented design and calibration procedures. All components are accessible to facilitate servicing.

Grade Dependent Specifications:

	J	K	A
Frequency	10MHz	10MHz	10MHz
Aging	$<1x10^{-9}$/day	$<5x10^{-10}$/day	$<2x10^{-10}$/day
Allan Var. (1s)	$<1x10^{-11}$	$<5x10^{-12}$	$<2x10^{-12}$
Phase Noise:			
10Hz	<-120 dBc/Hz	<-125 dBc/Hz	<-130 dBc/Hz
100 Hz	<-150 dBc/Hz	<-150 dBc/Hz	<-150 dBc/Hz
1 kHz	<-158 dBc/Hz	<-158 dBc/Hz	<-158 dBc/Hz
10 kHz	<-158 dBc/Hz	<-158 dBc/Hz	<-158 dBc/Hz
Temp. Range	0° to 50°C	-20° to 50°C	-55° to 75°C
Temp. Stability (over T range)	$<2x10^{-9}$	$<1x10^{-9}$	$<5x10^{-10}$
Power:			
Warmup	8W	8W	12W
25°C	3W	3W	3W

Output

Output Level	1 Vrms into 50Ω (+13 dBm)
Output Accuracy	±5%
Output Waveform	Sinewave
Harmonic Distortion	<-60 dBc

Tuning

Mechanical Tuning Range	>±3 Hz
EFC Range and Slope:	
Option 1	0 to 10 V, 5V nominal, +0.5 Hz/V
Option 2	0 to 10 V, 5V nominal, -0.5 Hz/V
Option 3	-10 to 10V, 0V nominal, +0.25 Hz/V
Option 4	-10 to 10V, 0V nominal, -0.25 Hz/V
Option 5	-5 to 5V, 0V nominal, +0.5 Hz/V
Option 6	-5 to 5V, 0V nominal, -0.5 Hz/V
Option 7	0 to 6V, 3V nominal, +0.75 Hz/V
Option 8	0 to 6V, 3V nominal, -0.75 Hz/V

General

Output Connector	Pin, SMA, SMB, or SMC
Supply Voltage	+15 Vdc or +24 Vdc
Size	2"x2"x4"
Weight	1 lb

Ordering Information

SC10-VS--E-T-S-N-A-CON **$250 (Base Price)**

VS	**15 for +15 Vdc operation, 24 for +24 Vdc operation**
E	**1 - 8 specifying the EFC range and slope. (See Specifications)**
T	**J, K, or A per the required temperature range**
S	**J, K, or A per the required stability vs. ambient temperature**
N	**J, K, or A per the required noise level (Allan variance and Phase noise)**
A	**J, K, or A per the required daily aging rate**
CON	**PIN, SMA, SMB, or SMC for pin, SMA, SMB or SMC 10 MHz connectors**

Price Modifiers

Multiply price by:
1.0 for each J grade option specified
1.2 for each K grade option specified
1.4 for each A grade option specified

Add $10 for SMA, SMB, or SMC connectors

For order quantities of:	**Multiply price by:**
1-4	**x1.5**
5-9	**x1.4**
10-24	**x1.3**
25-49	**x1.2**
50-99	**x1.1**
>100	**x1.0**

LCR Meters

Models SR715/SR720 — Inductance, Capacitance, and Resistance Meters

- ***0.05% Basic Accuracy (SR720), 0.2% (SR715)***
- ***Five Digit Display of L, C, R and Q or D***
- ***Test Frequencies to 100 kHz***
- ***Up to 20 Measurements per Second***
- ***Binning and Limits for Part Sorting***
- ***Standard RS-232 Interface***
- ***External Capacitor Bias up to 40 V***
- ***Optional GPIB and Handler Interface***

SR715/SR720 Overview

The SR715/720 family of LCR Meters quickly and accurately measures passive components with as little as 0.05% error. These easy-to-use instruments are simple to setup, adjust and calibrate. These factors, combined with low cost, make these meters perfect for applications such as incoming inspection, quality control, automated test, and general benchtop use.

Flexibility

Five different test frequencies, three preset voltage levels, internal and external capacitor biasing, and three different measurement speeds allow unmatched flexibility to create your own test environment. The convenient automeasure mode simplifies measuring unknown parts by automatically calculating the best part model and range for any component.

Calibration

Stray impedances are removed from measurements with a simple, fast, null calibration procedure. The built-in Kelvin fixture handles radial lead components, and adaptors are included for axial components. Optional fixtures include SMD tweezers, Kelvin clips, and BNC adapters.

Automation

Built-in binning sorts your components into overlapping or sequential bins. Up to nine complete instrument configurations may be saved to non-volatile memory for quick setup. For easy integration with computerized test environments an RS-232 interface is standard, IEEE-488 and handler interfaces are optional.

SR715/SR720 Features

Test Fixtures

The SR715/720's Kelvin fixture uses two wires to carry the test current and two independent wires to sense the voltage across the device under test (DUT). This prevents the voltage drop in the current-carrying wires from affecting the voltage measurement. Radial leaded components are simply inserted into the test fixture, one lead in each side. Axial leaded devices require the use of the standard axial fixture adapters. Surface mount (SMD) devices or components with large or unusually shaped leads can be measured with SMD tweezers or Kelvin clips. The tweezers and clips attach directly to the SR715/720 test fixture. An optional BNC fixture adapter allows you to connect a remote fixture or other equipment through one meter of coaxial cable.

Display

The five digit LED display shows measured values, entered parameters, instrument status, and user messages. When making normal measurements, the major parameter (L, C, or R) is shown on the left display and the appropriate minor parameter (Q, D or R) is shown on the right display (see the following SRS Tech Note for a discussion of these parameters). The number of displayed digits and the location of the decimal points are automatically adjusted according to the selected range and resolution. Indicator LEDs show the parameter (L, C, or R and Q, D, or R) being displayed. Measurements can be displayed as nominal values, deviations, or as percentage deviation. Units of the major parameter (Ohms, Farads, or Henrys) are indicated by LEDs located between the two displays. The minor parameter is either dimensionless (Q and D) or has the units of Ohms (R). Status information relating to the remote interfaces is conveniently displayed on the front panel.

Parameter Selection

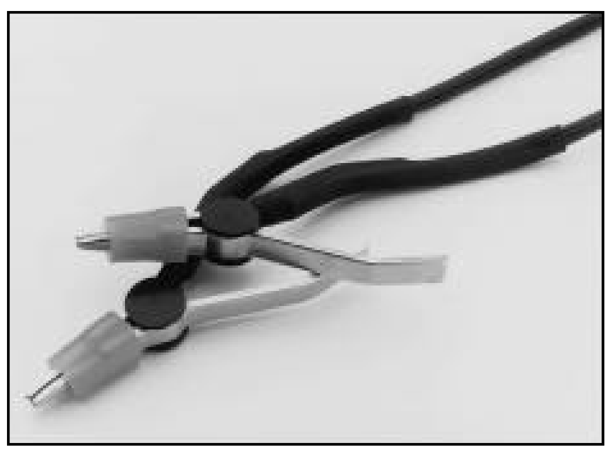
Measurements can be performed at a number of discrete frequencies: 100 Hz, 120 Hz, 1 kHz, 10 kHz or 100 kHz (SR720 only). The drive voltage can be selected to preset values (0.1, 0.25 and 1.0 V), or set from 0.0 to 1.0 V in 50 mV increments. Both series or parallel equivalent circuit models of a component can be measured. Capacitor measurements use either the internal 2.0 Vdc bias or an external DC source of up to 40 V. External bias is applied through the back panel banana jacks, and is fused at 250 mA.

Either continuous or triggered measurements can be selected. Measurements are triggered from the front panel, through the optional handler interface, or through either computer interface. Measurements can be taken at rates of 2, 10, or 20 measurements per second for test frequencies of 1 kHz or higher. Consecutive readings can be averaged between 2 and 10 times for increased accuracy.

Simple to Operate

The power and flexibility of the SR715/720 does not come at the expense of ease of use. A convenient AUTO measurement mode automates the selection of setup parameters and quickly determines the appropriate device model for whatever component is being measured. Auto Range automatically selects the appropriate units and display resolution for the DUT. And up to 9 complete instrument setups can be stored in non-volatile memory for quick recall at a later time.

Surface Mount Tweezers



Kelvin Clips

Convenient Calibration

The SR715/SR720 makes it simple to compensate for lead impedance and stray fixture and cable capacitance. The null calibration procedure, which takes just over a minute to perform, automatically corrects both open and short circuit parameters at all frequencies and all ranges. Null calibration should be performed after the fixture configuration has been changed, e.g. when the tweezers are added or removed. However, since the null calibration data is stored with instrument settings, it's easy to store calibrations for a number of different fixture configurations and quickly recall them later.

Binning

The SR715/720 has built-in features to aid in component sorting. This is especially useful for production testing, incoming inspection, device matching or when you need to test multiple devices of similar value. The binning feature simplifies parts sorting by eliminating the need to read the major and minor parameters and then deciding what to do with each part. Binning configurations can be entered from the keyboard or over any of the computer interfaces. The SR715/7220 allows you to sort components into as many as ten different bins. The meters support three types of binning schemes: pass/fail, overlapping and sequential. Pass/Fail has only two bins; good parts and everything else. Overlapping (or nested) bins have one nominal value and are sorted into progressively larger bins (i.e., ±1%, ±2%, ±3%). Sequential bins can have different nominal values, each separated by a percentage or a nominal value and asymmetrical limits. Binning parameters are also easily sorted to non-volatile RAM for quick setup in production environments.

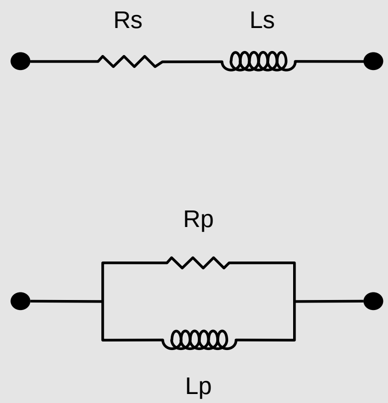
Rear Panel

Two rear panel input connections are provided for the external bias voltage. The bias supply should be floating and well filtered. The applied voltage must be 40 Vdc or less, and current limited to 250 mA. This input is independently fused to protect the meter.

An optional handler interface provides control lines to a component handler for sorting. The interface has an input trigger line and output lines indicating bin-data available, busy and 10 separate bins. The connector is a male DB25, and uses open collector logic. A standard RS-232 interface allows complete control of all instrument functions by a remote computer. A GPIB interface is included with the handler option.

SR715/SR720 Rear Panel

Ideal Device Models



$$Z_s = R_s + j\omega L_s = R_s(1+jQ) = \omega L_s\left(\frac{1}{Q} + j\right)$$

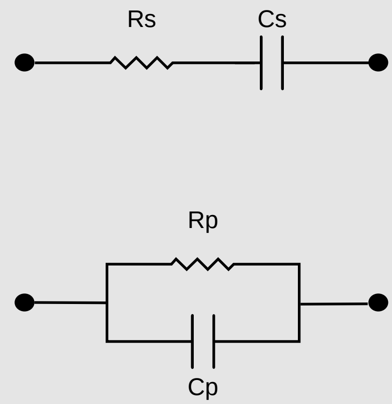
$$Z_p = \frac{j\omega L_p R_p}{R_p + j\omega L_p} = \frac{R_p(1+jQ)}{1+Q^2}$$

$$Q = \frac{R_p}{\omega L_p} = \frac{\omega L_s}{R_s} \qquad Q = \frac{1}{D}$$

$$L_s = \frac{Q^2}{1+Q^2}L_p \qquad R_s = \frac{\omega L_s}{Q} \qquad R_p = Q\omega L_p$$

$$R_p = \frac{1}{G_p} \qquad B_p = -\frac{1}{\omega L_p}$$

$$Y = G_p - \frac{j}{\omega L_p}$$



$$Z_s = R_s - j\frac{1}{\omega C_s} = \frac{1}{\omega C_s}(D-j) = R_s\left(1 - \frac{j}{D}\right)$$

$$Z_p = \frac{R_p}{1 + j\omega R_p C_p} = \frac{D^2 R_p + \frac{1}{j\omega C_p}}{1+D^2}$$

$$D = \omega R_s C_s = \frac{1}{\omega R_p C_p} \qquad D = \frac{1}{Q}$$

$$C_s = \left(1+D^2\right)C_p \qquad R_s = \frac{D^2}{1+D^2}R_p$$

$$R_p = \frac{1}{G_p} \qquad B_p = \omega C_p$$

$$Y = G_p + j\omega C_p$$

All non-ideal passive devices (resistors, inductors and capacitors) can be modeled as a real component (resistor) either in series or in parallel with a reactive component (capacitor or inductor). The impedance of these components changes as a function of frequency. The series and parallel models are mathematically equivalent and can be transformed back and forth with the shown equations. The SR715/720 can switch between either parallel or series equivalent circuits.

Usually one model is a better representation of the device under operating conditions. The most accurate model depends on the device and the operating frequency. Certain devices are tested under conditions defined by the manufacturer or industry standard. For example, electrolytic capacitors are often measured in series at 120 Hz in the C+R mode, so the ESR (equivalent series resistance) can be measured.

The equivalent series resistance in capacitors includes things like dielectric absorption in addition to the ohmic losses due to leads. It is often listed on data sheets for electrolytic capacitors used in switching power supplies. At high frequencies, the ESR is the limiting factor in the performance of the capacitor.

The quality factor, Q, is the ratio of the imaginary impedance to the real impedance. For inductors, a high Q indicates a more reactively pure component. A low Q indicates a substantial series resistor. Q varies with frequency. With resistors, often all that is stated is that the resistor has low inductance.

The dissipation factor, D, is equal to 1/Q and is the ratio of the real impedance to the imaginary impedance. A low D indicates a nearly pure capacitor. D is commonly used when describing capacitors of all types.

Overview

The SR715/720 LCR Meters are multifrequency impedance measuring instruments, capable of measuring resistance, capacitance or inductance over a range of more than 13 orders of magnitude. The SR720 has a basic accuracy of 0.05% and has 5 test frequencies. The SR715 has a basic accuracy of 0.2% and 4 test frequencies. The built-in Kelvin fixture includes adapters for axial as well as radial components. Options include GPIB/Handler Interface, SMD tweezers, Kelvin clips, and a 4 wire breakout box. The flexibility and accuracy of the SR715/720 make them widely accepted in high accuracy measurements for a variety of applications.

Measurement Modes

Measurement Modes	Auto, R+Q, L+Q, C+D, C+R	
Equivalent Circuit	Series or Parallel	
Parameters Displayed	Value, Deviation, % Deviation or Bin Number. Deviation and % Deviation are calculated relative to a stored value.	
Averaging	2 - 10 Measurements	
Measurement Range		
R+Q:	R	0.0001Ω-2000MΩ
	Q	0.00001 - 50
L+Q:	L	0.0001 µH - 99999 H
	Q	0.00001 - 50
C+D:	C	0.0001 pF - 99999 µF
	D	0.00001 -10
C+R:	C	0.0001 pF - 99999 µF
	R	0.00001Ω - 99999 kΩ

Test conditions

Test Frequency	Fixed frequencies at 100 Hz, 120 Hz, 1 kHz, 10 kHz (SR715 and SR720) 100 kHz (SR720 only).
Frequency accuracy	±100ppm.
Drive Voltage	Preset Levels: 0.10, 0.25, and 1.0 Vrms. Vernier: 0.1 to 1.0 Vrms with 50 mV resolution.
Drive levels accuracy	±2%.
Measurement Rate	Slow, Medium, Fast: 2, 10, or 20 measurements per second at test frequencies of 1 kHz and above and about 0.6, 2.4, or 6 measurements per second at 100 Hz and 120 Hz.
Ranging	Auto or Manual
Triggering	Continuous, Manual, or Remote over RS-232, GPIB or Handler Interface
Bias Voltage	Internal: 2.0 Vdc ±2% External: 0 to +40 Vdc, fused @ 250mA

Accuracy

Basic Accuracy	SR715: 0.20% SR720: 0.05%

See Accuracy Graphs on facing page.

Features

Fixture	4-Wire Kelvin fixture for radial leaded parts with adapters for axial leaded parts.
Protection	Protected up to 1 Joule of stored energy (for charged capacitors). Fused at 0.25 A output current for biased measurement.
Zeroing	Open and Short Circuit Compensation.
Compensation Limits	Short: R < 20 Ω, Z < 50 Ω Open: Z > 10 kΩ
Binning	Up to 8 Pass Bins, QDR and General Fail Bins, defined from the front panel or over the computer interfaces. Binning setup may be stored in non-volatile memory.
Store and Recall	Stores 9 Complete Instrument Setups. Recall 0 recalls Default Setup.
RS-232 Interface	All instrument functions can be controlled or read over the interface.
GPIB and Handler	Standard IEEE-488 Interface Optional, Handler Interface is a 25 pin connector, positive logic for binning and control.

General

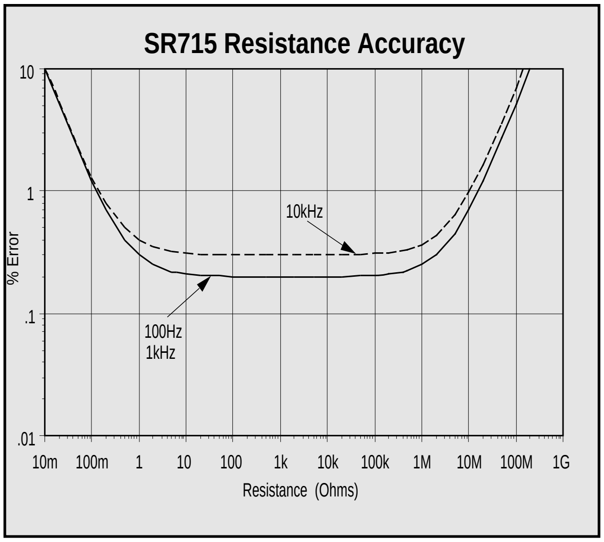
Operating Conditions	0 - 50 °C, < 85% relative humidity.
Power	20 Watts, 100/120/220/240 VAC, 50 or 60 Hz.
Dimensions (W x H x L)	13.5" W x 14" L x 4" H
Weight	10 lbs
Warranty	One year parts and labor on any defects in material or workmanship.

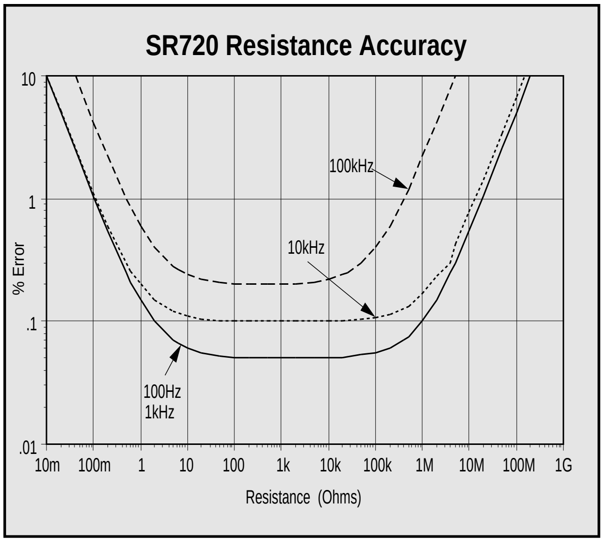
Ordering Information

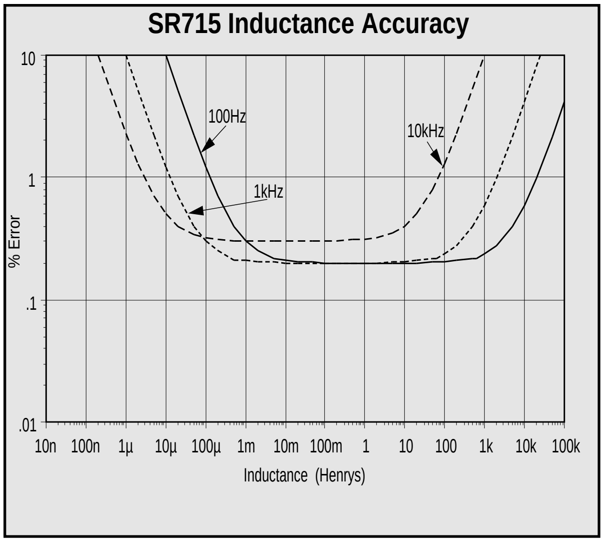
SR715	**LCR Meter, 0.2% Accuracy**	**$1295**
SR720	**LCR Meter, 0.05% Accuracy**	**$1995**
Option 01	**Parts Handler and GPIB Interface**	**$495**
SR726	**Kelvin Clips**	**$300**
SR727	**Surface Mount Tweezers**	**$300**
O727RT	**2 pair Replacement Tips**	**$45**
SR728	**BNC Adapter**	**$20**

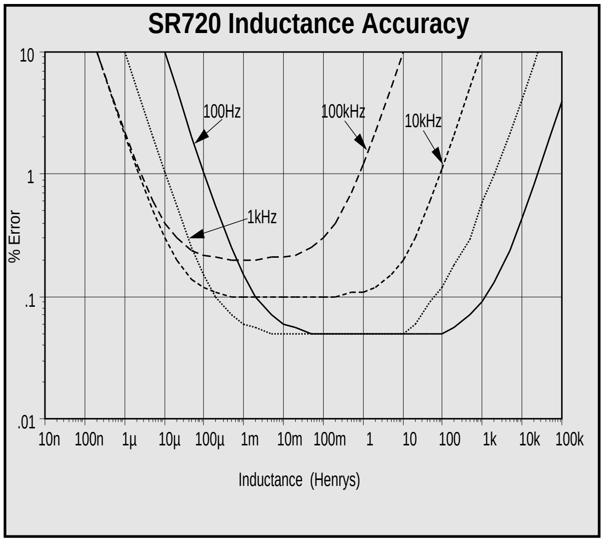
Note

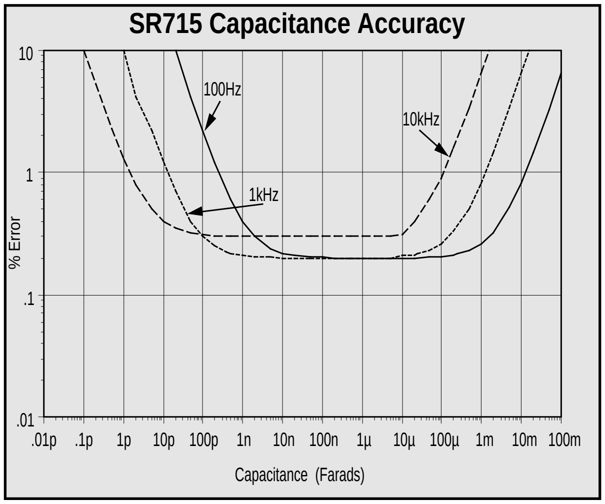
The SR715 is not available for sale in Pacific Rim/Far Eastern countries.

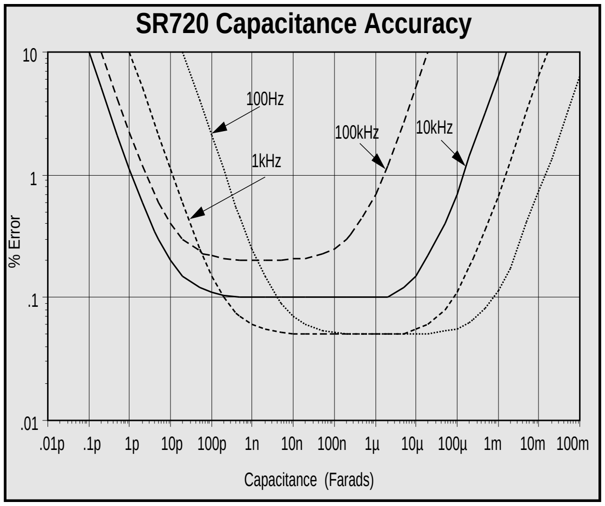

SR715 Resistance Accuracy
10kHz
100Hz
1kHz
% Error
10
1
.1
.01
10m 100m 1 10 100 1k 10k 100k 1M 10M 100M 1G
Resistance (Ohms)


SR720 Resistance Accuracy
100kHz
10kHz
100Hz
1kHz
% Error
10
1
.1
.01
10m 100m 1 10 100 1k 10k 100k 1M 10M 100M 1G
Resistance (Ohms)


SR715 Inductance Accuracy
100Hz
10kHz
1kHz
% Error
10
1
.1
.01
10n 100n 1µ 10µ 100µ 1m 10m 100m 1 10 100 1k 10k 100k
Inductance (Henrys)


SR720 Inductance Accuracy
100Hz
100kHz
10kHz
1kHz
% Error
10
1
.1
.01
10n 100n 1µ 10µ 100µ 1m 10m 100m 1 10 100 1k 10k 100k
Inductance (Henrys)


SR715 Capacitance Accuracy
100Hz
10kHz
1kHz
% Error
10
1
.1
.01
.01p .1p 1p 10p 100p 1n 10n 100n 1µ 10µ 100µ 1m 10m 100m
Capacitance (Farads)


SR720 Capacitance Accuracy
100Hz
100kHz
10kHz
1kHz
% Error
10
1
.1
.01
.01p .1p 1p 10p 100p 1n 10n 100n 1µ 10µ 100µ 1m 10m 100m
Capacitance (Farads)

Thermocouple Monitor

Model SR630 — 16 Channel Scanning Thermocouple Monitor

- *16 Channels*
- *B, E, J, K, R, S and T Type Thermocouples*
- *0.1 °C Resolution*
- *Displays °C, °K, °F, and Volts DC*
- *2,000 Point Non-Volatile Memory*
- *Four Analog Outputs: V = ±mX + b*
- *May be used as a 1:15 Analog Multiplexer*
- *IEEE-488, RS-232, and Printer Interfaces*

SR630 Overview

The SR630 is a 16 channel thermocouple monitor designed to read, scan, print, and log temperatures or voltages. You can use any one of seven standard thermocouple types to read temperatures from -200 °C to +1700 °C. For remote monitoring applications, the SR630 can time-stamp and store up to 2000 readings in non-volatile memory for later analysis.

Accessible Data

Temperature readings from the SR630 can be viewed on the front panel, or queried via the instrument's standard RS-232 or GPIB interfaces. In addition, the standard Centronics printer port provides convenient hardcopy output in either tabular or stripchart format.

Flexible Configuration

Temperature measurements can be made up to 12 times per second, with values reported in either °Fahrenheit, °Celsius, °Kelvin, or Volts. Programmable audible-alarm limits for each channel alert you to out-of range temperature conditions. A back panel relay output is available to provide shut-down capability when the limit conditions are exceeded.

Voltage Monitor

The SR630 can also be configured as a 16 channel DC voltmeter, with full scale ranges from 30 mV (1 µV resolution) to 100 V. As a voltmeter, the unit has 0.05% accuracy and 1 µV input offset drift. Each channel can be set independently to monitor either temperature or voltage, giving the SR630 even more flexibility in your application. Four rear-panel outputs are also available to provide a voltage proportional to the temperature of the first four input channels. The voltage outputs can be used to drive recorders, or to control external instrumentation.

SR630 Features

Inputs

Sixteen screw-terminal inputs are mounted on a rear-panel isothermal block for cold junction compensation. The isolated differential inputs have a 250 V breakdown level, allowing the SR630 to tackle difficult applications such as temperature profiling of electrically live equipment. Each of the 16 channels may be independently set to display in units of millivolts, volts, °K, °C, or °F. Similarly, thermocouple type, nominal temperature, temperature limit and alarms may be uniquely set for each channel, providing complete flexibility in the configuration of the instrument. Access to any channel parameter is provided through the front panel or via the computer interfaces. The SR630 can store up to nine entire instrument configurations, including thermocouple type and temperature limits for all 16 channels.

Outputs

There are four analog outputs on the rear panel of the instrument. These outputs have a full scale range of ±10.0 Vdc, with 5 mV resolution. The outputs can be set directly through the computer interfaces, or they can output a voltage proportional to the reading on channels one through four. When used as proportional outputs, both the slope (m) and offset (b) of the equation, Vout = ±mX + b, are changed by setting the nominal temperature and chart span for the corresponding channel. X is either temperature or voltage, depending on the setting for each channel.

You may use the analog output to drive chart recorders, or they can provide a feedback signal for proportional temperature control systems. They may also be used as general purpose analog control signals, since each level can be set via the computer interfaces to a fixed value.

Alarms

Each channel can be programmed individually with an upper and lower temperature limit. An audio alarm and a relay closure indicate when any channel exceeds its preset temperature or voltage limits. The front-panel LED display will indicate which channel has exceeded its limit.

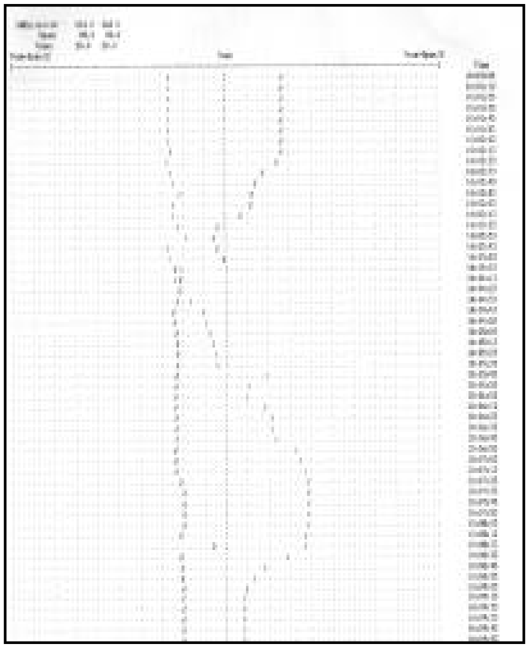
Datalogging Capability

An internal, battery backed-up clock/calendar is used to time-stamp temperature readings. The SR630 can be set to scan, time-stamp, and record up to 16 channels at intervals from 10 to 9999 seconds. Data is either sent immediately to the printer, or stored in the internal 2,000 point buffer for output at a later time (via GPIB or RS-232).

Built-in Printer Interface

A standard Centronics printer interface makes it simple to get hardcopy output of temperature or voltage scans. Two hardcopy formats are provided. The SR630 can print a continuous stripchart showing up to 16 different temperatures (see the example below). In addition, data

SR630 Rear Panel



Stripchart Formatted Printer Output

can be printed in a tabular format which logs the time, date, and temperature or voltage for each channel.

Analog Multiplexer

The SR630 can also function as a 1:15 analog multiplexer. Any of the first 15 input channels can be switched out to channel 16 and passed on to other instruments. This feature is useful in ATE systems and many other monitoring applications.

Thermocouples Available

Although the SR630 will read types B, E, J, K, R, S, and T type thermocouples, for many applications type K, or Chromel/Alumel, will serve well. K type thermocouples offer a wide temperature range (-200°C to +1250°C), a low standard error, and good corrosion resistance. SRS can supply welded K-type thermocouples with either Teflon or high-temperature glass insulators.

TIME	DATE	Ch1 C	Ch2 C	Ch3 C	Ch4 C	Ch5 C	Ch6 C	Ch7 C	Chl
10:31:05	25/06/92	20.5	37.6	29.4	20.4	27.4	27.3	27.4	27.
10:31:15	25/06/92	20.5	37.5	29.3	20.3	27.5	27.5	27.5	27.
10:31:25	25/06/92	20.0	37.1	29.3	19.8	27.7	27.8	27.8	27.
10:31:35	25/06/92	19.4	36.2	28.6	18.5	28.1	27.9	28.3	28.
10:31:45	25/06/92	20.1	36.1	27.8	17.6	30.6	30.5	30.6	30.
10:31:55	25/06/92	20.5	36.5	27.6	16.6	31.2	31.1	31.3	31.
10:32:05	25/06/92	20.0	35.0	26.0	14.6	32.1	31.7	32.2	32
10:32:15	25/06/92	15.1	29.5	20.9	10.3	33.1	32.5	33.0	33.
10:32:25	25/06/92	10.4	26.1	18.0	8.2	34.9	34.4	34.8	34
10:32:35	25/06/92	11.5	26.9	19.6	9.9	38.1	38.0	38.5	38.
10:32:45	25/06/92	12.9	28.5	20.1	10.9	38.8	38.8	38.8	38
10:32:55	25/06/92	13.5	28.9	20.3	11.5	38.5	38.5	38.5	38.
10:33:05	25/06/92	14.3	29.6	20.9	12.3	38.6	38.6	38.7	38
10:33:15	25/06/92	15.4	30.7	22.5	13.4	38.7	38.7	38.7	38
10:33:25	25/06/92	16.3	31.6	23.5	14.6	38.6	38.6	38.6	38
10:33:35	25/06/92	17.0	32.5	24.6	15.6	38.5	38.4	38.4	38
10:33:45	25/06/92	17.8	33.3	25.8	16.4	38.2	38.3	38.2	38
10:33:55	25/06/92	18.3	33.9	26.3	17.2	38.0	38.0	38.0	38

Tabular Output

Overview

The SR630 is a 16 channel thermocouple monitor designed to read, scan, print and log temperatures or voltages. It can use any one of seven standard thermocouple types to remotely read temperatures from -200 °C to plus 1700 °C. The SR630 can be used as a remote datalogger, time stamping the readings and storing 2000 points for analysis at a later time. Readings are directly output via a Centronics printer port, and RS-232 and IEEE-488 interfaces are included.

Thermocouple

Channels	16
Thermocouple Types	B, E, J, K, R, S, T
Display Units	Degrees C, F and K
Display Resolution	0.1 Degree C
Temperature Displays	Actual, Nominal, or Offset
Open Check Current	250 µA
Accuracy	0.5 Degrees C for J, K, E, and T 1.0 Degrees C for R, S, and B

Errors are for the SR630 only. Standard errors for thermocouple wire are 2 to 5 times the error due to the SR630. See section on thermocouple reference data for additional information.

Voltmeter

Channels	16
Input Type	Independent, floating, and differential
Input Resistance	10 MΩ between + and -, >1 GΩ to ground
Input Capacitance	.001 µF
Input Bias Current	<100 pA
Input Protection	250 Vrms
Full Scale Display	±9.999, ±99.99, or ±999.9 mVdc ±9.999 or ±99.99 Vdc
Range Select	Automatic
Resolution	±1 of least significant displayed digit
Offset	±2 of least significant displayed digit
Gain Accuracy	0.05%
Conversion Rate	10/sec for 50 Hz line, 12/sec for 60 Hz
Line Rejection	>100:1
Common Mode	±200V

Scanning and Data Logging

Scan Time	10 to 9999 seconds to read all selected channels
Alarm	Temperature or voltage limit for each channel
Scan Enable	Channel may be scanned or skipped
Proportional Outputs	For Channels 1, 2, 3 and 4.
Printer Output	Voltages, Temperatures, Time and Date as a list or in a graphical format.
Data Memory	Last 2000 measurements in battery backed-up memory.

General

Relay output	Switching 10 A
Store and Recall	Nine locations for instrument set-up
Interfaces	RS-232, GPIB, and Centronics Printer (Standard) All instrument functions may be set and read via RS-232 or GPIB.
Power	10 W, 100/120/220/240 Vac, 50/60 Hz.
Rack Mount	Optional
Dimensions	8.5" x 3.5" x 13" (W x H x D)
Weight	9 lb.
Warranty	One year parts and labor on any defects in material or workmanship

Ordering Information

SR630	**Thermocouple Monitor**	**$ 1495**
0630RMS	**Single Rack Mount**	**$ 85**
0630RMD	**Double Rack Mount**	**$ 85**
0630KT1	**K Type Thermocouples, 5', Teflon Insulated, Qty. 5**	**$ 50**
0630KT2	**K Type Thermocouples,10', Teflon Insulated, Qty. 5**	**$ 75**
0630KF1	**K Type Thermocouples,5', Fiberglass Insulated, Qty. 5**	**$ 50**
0630KF2	**K Type Thermocouples,10', Fiberglass Insulated, Qty. 5**	**$ 75**

The Thermocouple Effect

It has been known for a long time (Seebeck, 1822) that a voltage exists across the junction of dissimilar metals. Figure 1 shows a thermocouple junction formed by joining two metallic alloys, A and B. The voltage across the thermocouple junction depends on the type of metals used and the temperature of the junction. The mechanism responsible for this voltage is quite complicated, however, there are certain phenomenological results which make the effect useful for measuring temperature.

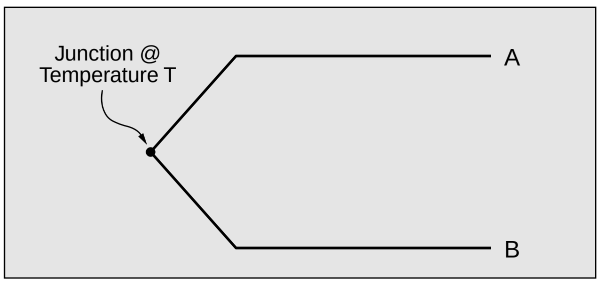


Fig. 1 Thermocoupble Junction

The first of these results is that the voltage is approximately linear with temperature. The change in junction voltage as a function of junction temperature is given by the equation :

$$\Delta V = a \times \Delta T$$

where 'a' is the Seebeck coefficient. The magnitude of this coefficient depends on the metals used to form the junction: typical values range from 0 to 100 µV/°C.

Nonlinearities

Unfortunately, the magnitude of the coefficient depends on temperature. It is generally smaller at low temperatures, and may change by more than a factor of two over the useful operating range of a thermocouple. Despite this non-linearity, the induced voltage is (usually) a monotonically increasing function of temperature, and the voltages generated by certain pairs of dissimilar metals have been accurately tabulated. These tabulated values are referenced to the voltage seen across a junction at 0 °C.

Additional Junctions

A problem arises when measuring the voltage across a dissimilar metal junction–two additional thermocouple junctions form where the wires connect to the voltmeter (Fig. 2). If the wire leads which connect to the voltmeter are made of alloy "C", then there exist thermal emf's at the A-C and B-C junctions. There are two approaches to solve this problem: use a reference junction at a known temperature, or make corrections for the thermocouples formed by the connection to the voltmeter.

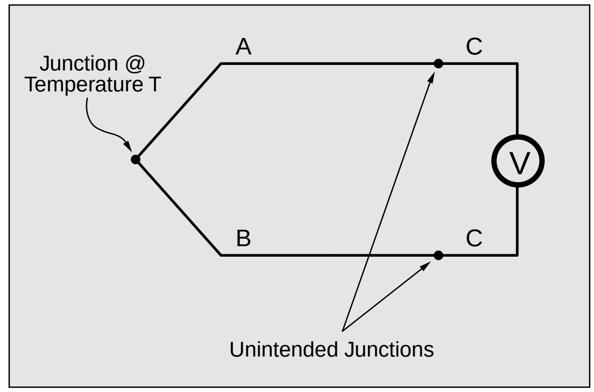


Fig. 2 Additional Junctions

Figure 3 shows the use of a "reference" or "compensating" junction. With this arrangement, there are still two additional thermocouple junctions formed where the compensated thermocouple is connected to the voltmeter. However, the junctions are identical (they are both junctions between alloys A and C). If the junctions are at the same temperature, then the voltages across each junction will be equal and opposite, and will not affect the measurement. Typically, the reference junction is held at 0 °C (by an ice bath, for example) so that the voltmeter readings may be used to look up the temperature.

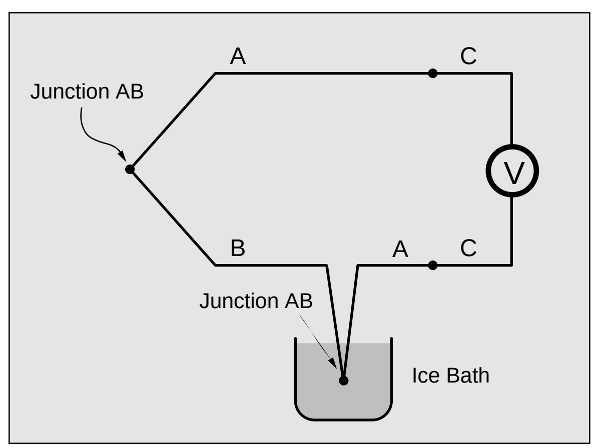


Fig. 3 Reference Junction Compensation

Compensation Without Reference Junctions

The second approach to the problem relies on the fact that the voltage across the junction A-C plus the voltage across the junction C-B is the same as the voltage across a junction of A-B. As long as all the junctions are at the same temperature, the presence of an intermediate metal (C) has no effect. This allows us to correct for the voltage seen by the voltmeter in Figure 2 by measuring the temperature at the A-C and B-C junctions and subtracting the voltage which we would expect for an A-B junction (at the measured temperature). In the SR630 the temperature of the A-C and C-B junctions are measured with a low cost, high resolution semiconductor detector, and the subtracted voltage is the tabulated voltage of the A-B thermocouple at the measured temperature of the A-C and C-B junctions. The advantage of this method is that any type thermocouple may be used without having to change compensation junctions or maintain ice baths.

Characteristics of Thermocouple Types

Any two dissimilar metals may be used to make a thermocouple. Of the infinite number of thermocouple combinations which can be made, the world has standardized seven types which exhibit a range of desirable features. These thermocouple types are known by a single letter designation: J, K, T, E, R, S or B. While the composition of these thermocouples are international standards, the color codes of the wires are not. For example, in the USA, the negative lead is always red, while the rest of the world uses red to designate the positive lead. Often, the standard thermocouple types are referred to by their trade names. For example, K type is sometimes called Chromel-Alumel, which is the trade names of the Ni-Cr and Ni-Al wire alloys.

It is important for a good thermocouple to have a large, stable Seebeck coefficient, wide temperature range, corrosion resistance, etc. Generally, each wire of the thermocouple is an alloy. Variations in the alloy composition and the condition of the junction between the wires are sources of error in temperature measurements. The standard error of thermocouple wire varies from ±0.8 °C to ±4.4 °C, depending on the type of thermocouple used.

Voltage vs. temperature measurements have been tabulated by NIST for each of the seven standard thermocouple types. These tables are stored in the read-only memory of the SR630. The instrument operates by converting a voltage measurement to a temperature, with the internal microprocessor interpolating to achieve 0.1 °C resolution.

The K type thermocouple is recommended for most general purpose applications. It offers a wide temperature range, low standard error, and has good corrosion resistance. The K type thermocouples provided by SRS have a standard error of ±1.1 °C, half the standard error designated for this type.

Type	B	E	J	K	R	S	T
Positive Material	Pt/Rh(30%)	Ni/Cr	Fe	Ni/Cr	Pt/Rh(13%)	Pt/Rh(10%)	Cu
Negative Material	Pt/Rh(6%)	Cu/Ni	Cu/Ni	Ni/Al	Pt	Pt	Cu/Ni
Positive Color(USA)	Grey	Purple	White	Yellow	Black	Black	Blue
Negative Color(USA)	Red	Red	Red	Red	Red	Red	Red
Lowest Temperature	50C	-200C	0C	-200C	0C	0C	-200C
Highest Temperature	1700C	900C	750C	1250C	1450C	1450C	350C
Minimum Std Error	±4.4C	±1.7C	±2.2C	±2.2C	±1.4C	±1.4C	±0.8C

Figure 4: Thermocouple Reference Data

High Voltage DC Power Supplies

Models PS310 (1.25 kV), PS325 (2.5 kV), and PS350 (5.0 kV)

- ***Up to 5 kV with 1 V Resolution***
- ***25 Watts Output Power***
- ***0.001% Regulation, 0.05% Accuracy***
- ***Low Output Ripple***
- ***Dual Polarity***
- ***Programmable Limits and Trips***
- ***Store and Recall of Settings***
- ***Optional GPIB Interface***

PS300 Series Overview

The PS300 series of high voltage DC power supplies consists of three efficient, microprocessor-controlled, switching power supplies capable of delivering 25 W of output power at voltages up to 5 kV.

Model	Output Voltage	Maximum Current
PS310	12 to ±1.25 kV	20 mA
PS325	25 to ±2.50 kV	10 mA
PS350	50 to ±5.00 kV	5 mA

High Accuracy and Low Ripple

All models have 0.001% regulation and 0.05% accuracy. Output ripple is less than 0.0015% of full scale, and the output voltage can be adjusted with 1 V resolution over the entire operating range.

Voltage and Current Displays

Convenient front-panel LED displays indicate both the voltage and current delivered to the load at all times. Adjustable voltage and current limits prevent you from operating the instrument above safe levels, while a current trip feature shuts off the high voltage in the event the current limit is exceeded.

Flexible Operation

Voltages can be programmed from the front panel, the optional GPIB interface, or by a rear-panel analog control signal. Rear-panel monitors provide analog signals proportional to the current and voltage output levels. Either positive or negative voltages can be selected with a rear-panel switch. For quick, convenient setup, up to nine complete instrument configurations can be stored in non-volatile memory for later recall.

PS300 Series Features

Voltage Output

All PS300 series supplies have rear-panel SHV (Kings type 1704-1 or equivalent) connectors. Optional cables allow connection with standard high voltage connectors (SHV or MHV). A three-position high voltage enable switch on the front panel prevents the high voltage from being turned on under computer control unless the switch as been manually armed. A highly visible red LED always indicates when the high voltage is on.

Limits and Trips

An adjustable voltage limit prevents the high voltage from being inadvertently set above a safe level. An independent current limit lowers the voltage setting until the current drawn by the load does not exceed the limit setting. Additionally, a current trip may be set which turns off the high voltage when the current limit is exceeded. After a trip, the unit can be configured to either attempt to turn the high voltage back on, or to leave it off until a manual reset.

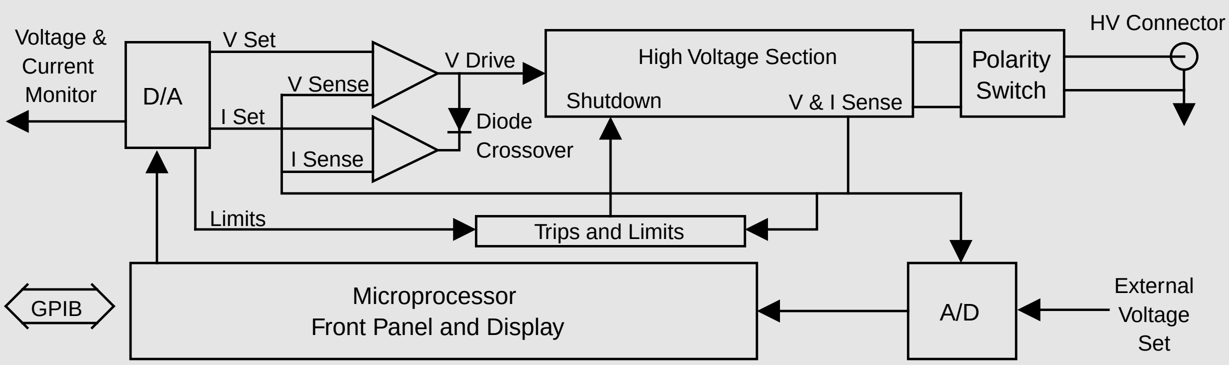
Adjustable Polarity

The output polarity is set with a rear-panel switch. Changing polarity must be done while the unit is off, and polarity cannot be changed via the computer interface. Output polarity is always displayed on the front panel with the voltage level.

Voltage and Current Monitors

Two rear-panel BNC connectors provide voltage and current monitoring capabilities. These connectors provide a 0 V to 10 V output corresponding to 0% to 100% of full scale output. These outputs are capable of driving 10 mA and have a 1 Ω output impedance.

The voltage monitor output can also be configured as a voltage control input. In this mode, a 0 V to 10 V signal applied at the input will cause the output high voltage to vary between 0% and 100% of full scale.The output is updated at a rate of 16 Hz. Additionally, the bandwidth of the voltage control input is limited by the overall slew rate specification of the instrument. (0% to 100% of full scale in less than 0.3 s under full load.)



All PS300 Series power supplies operate in the same manner, differing only in maximum voltage, maximum current and output ripple.
A high voltage (HV) section converts low drive voltage into high voltage (all high voltage components in the HV section are shielded). The output voltage (V) and current (I) are sensed and fed back to high gain compensation circuitry where they are compared to the programmed values. The compensation circuit controls the output voltage by setting the level of the drive voltage. A diode crossover allows control of both voltage and current.
Programmed values for the output, limits and trips are set by the microprocessor through a D/A converter.
Fast acting limit circuits check the sensed voltage and current. These work independently of the microprocessor to react quickly in protecting both the supply and load.
An A/D converter reads the sensed values for the front panel display or for reading over the GPIB interface.
The A/D also reads the External Voltage Set (when enabled), to ensure that limits are functional in the analog programming mode.

Stored Settings

All instrument settings are stored in non-volatile memory so that the instrument will remember its previous settings when powered up. For safety, the high voltage is always initially turned off. Up to nine complete instrument configurations, including limit and trip settings, can be stored in non-volatile memory for later recall. Settings can be recalled manually, or under computer control.

Optional GPIB Interface

An optional GPIB interface allows control of all instrument functions except output polarity. The required common commands of the IEEE-488.2 (1987) standard are supported. The power supply can be configured to generate service requests (SRQ) in the event of limit or trip conditions.

PS300 Series Rear Panel

Overview

The PS300 series are programmable precision high voltage dc power supplies for laboratory or test applications. They feature reversible polarity, excellent regulation and low output voltage ripple. The digital displays provide accurate readings of voltage and current. Also, digital entry of the current and voltage provides accurate resettability. Output voltage can be set from either the front panel, a remote control voltage, or over the optional GPIB interface. Voltage and current signals are also available for remote monitoring.

Voltage Output

Model	Output Voltage	Maximum Output Current
PS310	12 to ±1250 Volts	20 mA
PS325	25 to ±2500 Volts	10 mA
PS350	50 to ±5000 Volts	5 mA

Voltage Set Accuracy	0.01% + 0.05% of full scale
Voltage Display Accuracy	V Set Accuracy ±1 Volt, typical (±2 Volt, max)
Voltage Resolution	1 Volt (set and display)
Voltage Resettability	1 Volt
Voltage Limit Range	0 to 100% of full scale
Voltage Regulation	Line: 0.001% for ±10% line voltage change Load: 0.005% for 100% load change, typical.
Note:	Regulation specifications apply for >.5% (full load) to >1% (no load) of full scale Voltage. Below these values the unit may not regulate correctly.
Output Ripple	< 0.0015% of full scale, Vrms, typical < 0.002% of full scale, Vrms, maximum
Current Limit Range	0 to 105% of full scale
Current Set Accuracy	0.01% + 0.05% of full scale
Current Resolution	
PS310	10 µA
PS325	10 µA
PS350	1 µA
Current Display Accuracy	
PS310	±10 µA, typical (±20 µA,max)
PS325	±10 µA, typical (±20 µA,max)
PS350	±1 µA, typical (±2 µA, max)
Stability	0.01% per hour, < 0.03% per 8 hours
Temperature Drift	50 ppm / °C, 0° to 50° C, typical
Protection	Arc and short circuit protected; programmable voltage limits, current limits and current trip
Recovery Time	12 ms for 40% step change in load current, typical
Discharge Time	< 6 sec (to < 1% of full scale voltage with no load, typical)

Monitor Outputs

Output Scale	0 to +10 Volts for 0 to full scale output regardless of polarity
Current Rating	10 mA, maximum
Output Impedance	< 1Ω
Accuracy	0.07% of full scale
Update Rate	8 Hz

External Voltage Set

Input Scale	0 to +10 Volts for 0 to full scale output regardless of polarity
Input Impedance	1 MΩ
Accuracy	0.07% of full scale
Update Rate	16 Hz
Output Slew Rate	< 0.3 sec for 0 to full scale under full load

Mechanical Specifications

High Voltage Connector	SHV male (Kings Type 1704-1 or equiv.)
Mating Connector	SHV female (Kings Type 1705-14 or equivalent, not included)
Dimensions	16.0" x 8.1" x 3.5" (L x W x H) The PS300 series are 1/2 rack width (19 inch standard rack). Optional rack mounting kits are available for single or double rack mounts. The single rack mount provides mounting for one supply. The double rack mount provides side by side mounting for 2 supplies.
Weight	8 lbs
Input power	50 watts, 100, 120, 220, 240 VAC ±10%, 50 or 60 Hz
Warranty	One year parts and labor on any defects in materials or workmanship

Ordering Information

PS310	**1.25 KV HV Power Supply**	**$ 1250**
PS325	**2.5 KV HV Power Supply**	**$ 1250**
PS350	**5.0 KV HV Power Supply**	**$ 1250**
Option 01	**GPIB Interface**	**$ 495**
Option 02S	**Single Rack Mount**	**$ 85**
Option 02D	**Dual Rack Mount**	**$ 85**
Option 03A	**SHV to SHV cable, 10'**	**$ 50**
Option 03B	**SHV to MHV cable, 10'**	**$ 50**

Dual Channel Programmable Filters

Models — SR640 Lowpass, SR645 Highpass, SR650 Highpass/Lowpass

Two Independent Filter Channels
- ***115 dB / Octave Rolloff***
- ***1 Hz to 100 kHz Cutoff Frequency***
- ***0.1 dB Passband Ripple***
- ***80 dB Stopband Attenuation***
- ***4 nV/√Hz Input Noise***
- ***Up to 80 dB of Gain (x10,000)***
- ***GPIB and RS-232 Interfaces***

SR640 Series Overview

0The SR640 series consists of three dual channel filters. Each channel has a low-noise preamplifier, a precision highpass or lowpass filter section, and an output amplifier. Applications include anti-aliasing, audio analysis, and general frequency-domain signal conditioning.

Model	Configuration
SR640	Dual Lowpass
SR645	Dual Highpass
SR650	Highpass/Lowpass (Can be cascaded for Bandpass)

Sharp Filter Cutoff

Each filter has an 8-pole, 6-zero elliptic frequency response that provides extremely sharp rolloff. The filters roll off at 115 dB per octave, with only 0.1 dB of passband ripple and 80 dB of stopband attenuation. Cutoff frequencies may be set between 1 Hz and 100 kHz with three digits of resolution.

Low Noise Amplifiers

Each channel contains a low-noise differential preamplifier with an adjustable gain of 0-60 dB. Preamplifier noise is less than 4 nV/√Hz (at 1 kHz), while common mode rejection is greater than 90 dB at 1 kHz. An additional amplifier stage provides up to 20 dB of postfilter gain. The filter section can be bypassed, allowing the SR640 series filters to operate as flexible, low-noise amplifiers with a bandwidth of approximately 750 kHz.

Completely Programmable

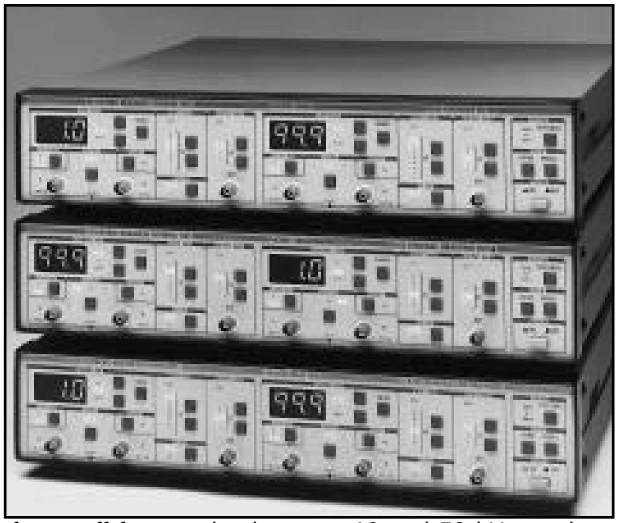
All functions of the SR640 series filters are completely programmable via the standard GPIB and RS-232 interfaces. Frequency selection, filter bypassing, and amplifier gains can be quickly programmed using a simple ASCII command set. An advanced, opto-isolated design eliminates direct electrical connections between the microprocessor circuitry and the analog filter sections, reducing filter noise.

Inputs and Preamplifiers

Each filter channel has a floating, fully differential input with 1 MΩ input impedance and 4 nV/√Hz of input noise (at 1 kHz). The analog ground for each filter channel is available at a rear-panel BNC connector to provide complete flexibility in grounding. The filter grounds may be tied to the instrument's chassis ground, or to any other point. Both AC and DC input coupling can be selected, and an invert key allows the phase of the signal to be shifted by 180° with a single keypress. Up to 60 dB of preamplifier gain can be specified in 10 dB increments. A simple offset adjust screw lets you null DC components in your signal.

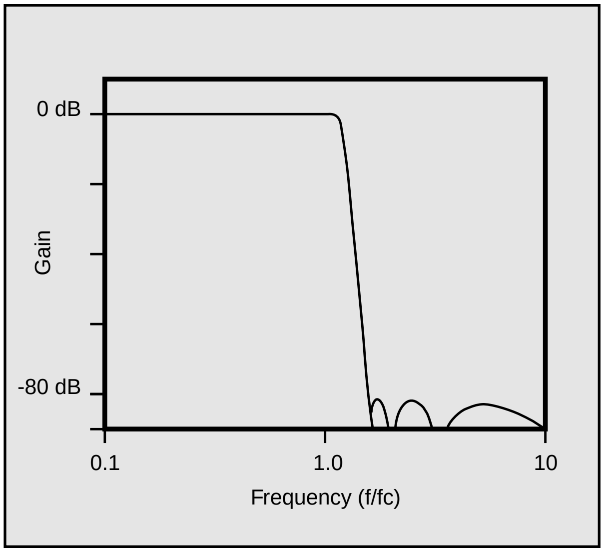
Filter Characteristics

Filter characteristics for the SR640 series filters are shown below. The exact filter transfer function is given in the specifications section. (The graph shows the lowpass response, the highpass response is found by inverting the curve around f=1.) The large amount of excess phase, typical of elliptical filters, makes the SR640 series filters optimum in frequency domain applications. Note that the 0.1 dB passband ripple specification for the SR640 series filters is a measured specification–not just the theoretical passband ripple of the transfer function. Other filter characteristics include an attenuation slope of 115 dB/octave and a stopband attenuation of >80 dB. Worst case phase match between channels is better than ±0.75° from DC to the cutoff frequency. The phase match improves to ±0.5° for cutoff frequencies between 10 and 50 kHz, and to ±0.25° for cutoff frequencies below 10 kHz.

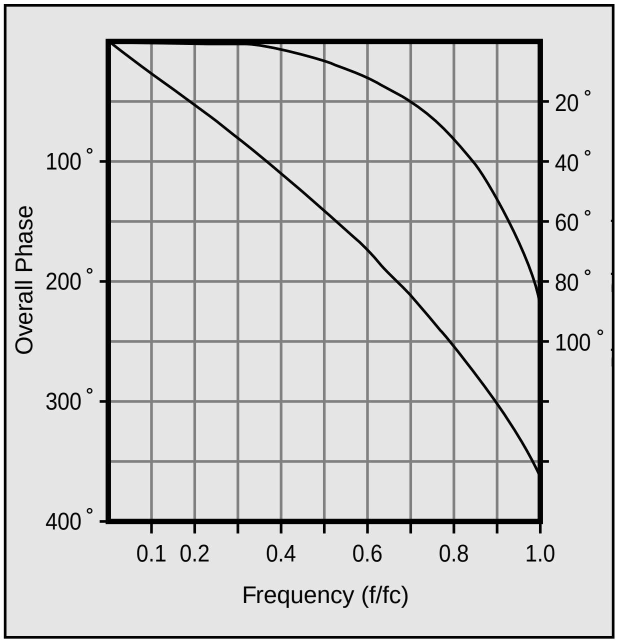


Postfilter Gain and Bypassing

After the filter, another amplifier section provides up to 20 dB additional gain. The standard output drives loads of >300 Ω at 10 Vpk-pk, while the optional high power output drives 10 Vpk-pk into a 50 Ω load. Each filter may be independently bypassed from the front panel or from the computer interface. When bypassed, both the



SR640 Amplitude Response



SR640 Phase Response

81

prefilter and postfilter gains remain active, and the resulting amplifier has a bandwidth of 450 kHz.

Programmable

A key feature of the SR640 series is its programmability. Both GPIB and RS-232 interfaces are standard, and all instrument settings can be queried and changed via the interfaces. Cutoff frequencies can be changed in under 100 ms, making the SR640 series ideal for applications involving frequency-agile filters. Careful attention has been paid to minimizing noise resulting from the presence of the microprocessor and GPIB and RS-232 interfaces. Both the interface electronics and the instrument's microprocessor control circuitry are optically isolated from the filter sections in order to reduce digital-analog crosstalk. The CMOS microprocessor "sleeps" while the interface and front panel are inactive in order to reduce the amount of digital noise in the box.

Stored Settings

Up to nine complete instrument configurations, including amplifier gains and cutoff frequencies for both channels, can be stored in non-volatile memory for later recall. Settings can be recalled from the front panel, or

SR640 Rear Panel

through the computer interfaces.

Overview

Each SR640 series programmable filter consists of two independent filter channels, a low-noise preamplifier and a variable gain output amplifier. The filters are an 8-pole, 6-zero elliptical design that have fast rolloff characteristics useful for anti-aliasing and general frequency-domain signal conditioning. RS-232 and GPIB interfaces allow remote instrument control by an external computer.

Filter

Frequency Range	1Hz to 100 kHz with 3-digit resolution
Type	8-pole, 6-zero elliptic
Rolloff	115 dB/octave
Passband Ripple	< 0.1 dB pk-pk
Stopband Attenuation	> 80 dB

Transfer Function (Biquad Cascade)

Stage 1:	
ωp	.634752
Q	.543939
ωz	na
Stage 2:	
ωp	.805995
Q	.950701
ωz	1.675935
Stage 3:	
ωp	.985073
Q	2.095314
ωz	1.935978
Stage 4:	
ωp	1.076768
Q	7.375143
ωz	2.883197

Input

Impedance	1 MΩ // 15 pF
Configuration	single ended (A or B) or differential (A-B)
Common Mode Reject.	> 85 dB at 1000 Hz for Input Gains>20dB
Coupling	AC or DC
Input Noise	6 nV/√Hz at 1 kHz with 60 dB input gain
Gain	0,10,20,30,40,50,60 dB ±0.2 dB
Maximum Input Signal	10 Volts pk-pk

Output

Impedance	< 1Ω
Full Scale Output Signal	10 Volts pk-pk into >300Ω
Option 01	10 V pk-pk into 50Ω
DC Offset	Adjustable to 0 Volts dc Gain 0,10, 20 dB ± 0.2 dB
Harmonic Distortion (<1Vpp Ouput)	< -80 dB below full scale at 100 Hz < -50 dB below full scale at 1 kHz
Spurious Components	< -80 dB below full scale with input source < 50Ω
Channel Crosstalk	< -110 dB below full scale with input source < 50Ω
Phase Match Between Channels, DC to f_c:	
1 Hz < fc < 10 kHz	±0.25°
10 kHz < fc < 50 kHz	±0.5°
50 kHz < fc < 100 kHz	±0.75°

General

Interfaces	GPIB and RS-232 standard. All instrument functions can be controlled and read through the interface.
Stored Settings	9 complete 2 channel instrument configurations may be stored in non-volatile memory.
Power	45 W, 100/120/220/240 VAC, 50/60 Hz
Dimensions	15.7"W X 3.0"H X 14.0"L
Weight	12 lbs
Rackmount Hardware	Included
Warranty	One year parts and labor on any defects

Ordering Information

SR640	**Dual Channel Lowpass Filter**	**$ 2990**
SR645	**Dual Channel Highpass FIlter**	**$ 2990**
SR650	**Lowpass/Highpass Filter**	**$ 2990**
Option 01	**High Level Output**	**$ 150**

Digital Lock-In Amplifier

Model SR850 — .001 Hz to 102 kHz DSP Lock-in Amplifier

- ***1 mHz to 102 kHz Frequency Range***
- ***100 dB Dynamic Reserve Without Prefilters***
- ***Digital Demodulator - No Nonlinearities***
- ***Synthesized Reference Oscillator***
- ***Ultra-low Drift***
- ***65,536 Point On Screen Display***
- ***16 Bit D/A outputs and A/D inputs***
- ***Built in Disk Drive, GPIB, and RS232***

SR850 Overview

The SR850 is the world's first completely digital lock-in amplifier. Based on an innovative DSP (Digital Signal Processing) architecture, the SR850 boasts a number of significant performance advantages over traditional lock-in amplifiers. Higher dynamic reserve (up to 100 dB), lower drift, lower distortion, and dramatically higher phase resolution are just a few of the improvements incorporated in the SR850. In addition, the CRT display and 65,536 point on-board memory make it possible to display and process data in a variety of formats unavailable with conventional lock-ins.

Digital Precision

At the input of the SR850 is a precision 18-bit A/D converter which digitizes the input signal at 256 kHz. The A/D converter, together with a high-speed DSP chip, replace the analog demodulator (mixer), low pass filters, and DC amplifiers found in conventional lock-ins. Instead of using analog components, the SR850 is implemented by a series of exact, precise, mathematical calculations which eliminate the drift, offset, nonlinearity, and aging inherent in analog parts. The same DSP chip digitally synthesizes the reference oscillator, providing a source with less than -70 dBc distortion, 100 µHz frequency resolution, and 2 mV of amplitude resolution.

Digital Flexibility

The SR850, with its 7" CRT display, offers a much larger choice of output options than are available with conventional lock-ins. Data can be displayed in a numeric format, as a bargraph, a polar plot, or a stripchart. With 65,536 points of on-board memory, and data acquisition rates ranging from 0.0625 Hz to 512 Hz, you are able to see exactly how your data changes in time–not just what the current output value is. And after the data has been acquired, the SR850 offers a variety of data reduction options such as Savitsky-Golay smoothing, curve-fitting, and statistical analysis. A built-in 3.5" disk drive, along with standard RS-232 and GPIB, makes it easy to transfer data to your computer .

SR850 Features

Input Channel

The SR850 has a fully differential input with 6 nV/√Hz of rms input noise at 1 kHz. Input impedance is 10 MΩ. Minimum full-scale input voltage is 2 nV, without any additional preamplifiers. In addition, the input can also be configured as a current input with selectable current gains of 10^6 and 10^8 Volts/Amp. A line filter (50 Hz or 60 Hz), and a 2x line filter (100 Hz or 120 Hz) are provided to eliminate line related interference. However, unlike conventional lock-in amplifiers, no tracking bandpass filter is needed at the input of the SR850. This filter is used by conventional lock-ins to increase dynamic reserve. Unfortunately, bandpass filters also introduce noise, amplitude and phase error, and drift. The DSP based design of the SR850 has such inherently large dynamic reserve that no tracking bandpass filter is required.

Reference Channel

The reference source for the SR850 can be an externally applied sine or square wave, or its own digitally synthesized reference source. Because the internal reference source is synthesized from the same digital signal that is used to multiply the input, there is virtually no reference phase noise when using the internal reference. The internal reference can operate at a fixed frequency, or can be swept linearly or logarithmically over the entire operating rage of 1 mHz to 102 kHz. Harmonic detection can be performed at any integer harmonic of the reference frequency, not just the first few harmonics.

The DSP approach also offers considerable advantages when working with an external reference. The time to acquire an external reference is only 2 cycles + 5 mS, (or 40 ms, whichever is greater) – about 10 times faster than conventional lock-ins.

Reference phase control is another area where the DSP based design excels. Because the SR850 uses a digital phase-shifting technique rather than analog phase-shifters, the reference phase can be adjusted with 1 millidegree resolution. In addition, the X and Y outputs are orthogonal to within 1 millidegree.

Outputs and Time Constants

Because the output time constants on the SR850 are implemented digitally, many flexible filtering options are available. Lowpass filter rolloffs of 6, 12, 18, and 24 dB/octave are available, with time constants ranging from 10 µs to 30 ks. Below 200 Hz, the SR850 can perform synchronous filtering. Synchronous filters notch out multiples of the reference frequency, an especially useful feature at low frequencies where the proximity of 2f components often necessitate the use of long time constants. The SR850 makes working at low frequencies far less tedious than with other lock-ins.

High Dynamic Reserve

The SR850 has the highest dynamic reserve of any lock-in available. The term “dynamic reserve”, despite being commonly used when comparing lock-ins, is often used imprecisely. At Stanford Research Systems, dynamic reserve is defined as follows:

The dynamic reserve of a lock-in amplifier at a given full-scale input voltage is the ratio (in dB) of the Largest Interfering Signal to the full-scale input voltage. The Largest Interfering Signal is defined as the amplitude of the largest signal at any frequency that can be applied to the input before the lock-in cannot measure a signal with its specified accuracy.

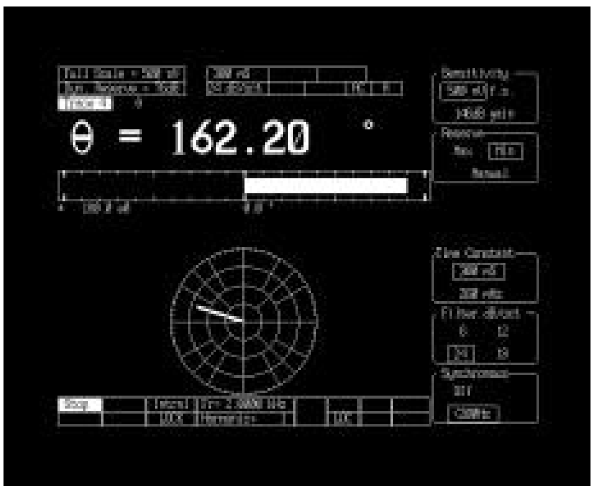
The SR850 has up to 100 dB of dynamic reserve. You might ask why only “up to” 100 dB of dynamic reserve. The answer, of course, is that no interfering signal can ever exceed the input range of the lock-in. For the SR850, this is 1Vrms. So at the 1Vrms full-scale input range, the dynamic reserve is 0 dB. But for full-scale input ranges of 10 µV or less, a full 100 dB of dynamic reserve is available.

It is important to keep in mind each manufacturer’s definition of dynamic reserve when comparing specifications. Some manufacturers define the Largest Interfering Signal as the largest interfering signal that can be applied without overloading the instrument. However, long before the instrument becomes overloaded its accuracy may be severely degraded, making this number an overly optimistic estimate of the useful dynamic reserve. Some manufacturers quote dynamic reserve for interfering signals that are outside the frequency range of the instrument and are therefore easier to reject. These points must be taken into account before simply comparing two dynamic reserve specifications.

In conventional lock-in amplifiers dynamic reserve is increased at the expense of stability. Because of the digital nature of the filtering and gain process in the SR850, the ultra-high dynamic reserve is obtained without any sacrifice in stability or accuracy. In addition, the SR850's high dynamic reserve is obtained without the use of analog bandpass filters, eliminating the noise and error that such filters introduce.

Traces and Displays

Data is acquired by the SR850 into arrays called "traces." Each trace can be defined as A x B / C, where A, B, and C can be chosen from: The in-phase component of the signal (X), the quadrature component (Y), the magnitude of the signal ($\sqrt{(X^2+Y^2)}$), the phase ($\tan^{-1}$Y/X), the reference frequency, the noise of the X and Y signals, or any of the four rear-panel auxiliary inputs. Common operations, such as ratioing, can all be performed in real time by defining an appropriate trace. Voltages proportional to the trace values are available on front-panel outputs, or the trace values can be displayed on the CRT. The SR850 can store up to 65,536 trace values internally.

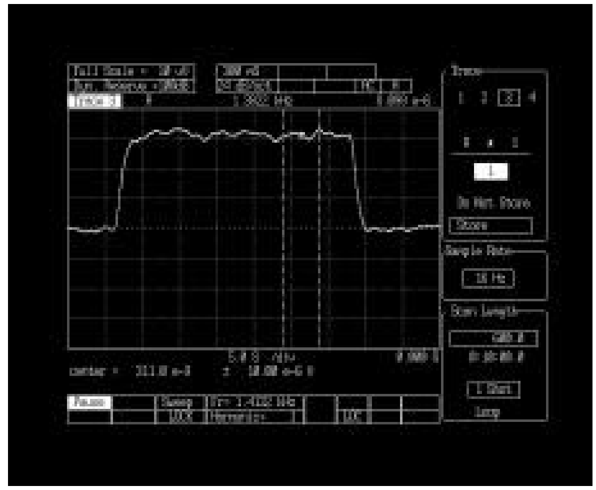


Polar and Bargraph Displays

The 7" CRT gives the SR850 more flexibility in displaying your data than any other lock-in available. Trace values can be displayed as a bargraph with an associated large numerical display, or as a stripchart showing the trace values as a function of time. Additionally, you can display polar plots showing the phasor formed by the in-phase and quadrature components of the signal. All displays can be easily scaled from the front panel or the computer interface, and an autoscale feature is available to quickly optimize the display. The screen can be configured as a single large display or as two vertically split displays.

Convenient Auto Measurements

Common measurement functions have been incorporated into single-key "auto" functions. The gain, phase, dynamic reserve, and display scaling can all be quickly optimized with a single keypress. For many common measurements, the instrument can be completely configured simply by using the autofunctions.



Stripchart Display

Auxiliary A/Ds and D/As

Four rear-panel A/D inputs provide millivolt resolution measurements of external signals. The measured values can be incorporated into one the SR850's trace definitions (making it easy to ratio with an external signal for instance), or can simply be displayed on the front panel or read via the computer interface. Four D/A outputs can provide either fixed output voltages, or a voltage level which scans synchronously with the SR850's frequency scans. Both the A/D inputs and the D/A outputs have a ±10 V range.

Advanced Analysis Features

The SR850's performance doesn't stop once data has been acquired–a full set of data processing features is also included. Multiple-range Savitsky-Golay smoothing can be applied to any of the trace arrays, and statistical information (mean, variance, sum) can be calculated for a selected trace region. A curve fitting routine calculates best fits to lines, exponentials, and gaussians for any portion of your data. And a trace "calculator" lets you perform a variety of simple arithmetic and trigonometric operations on trace data.

Interfaces and Hardcopies

The SR850 comes standard with RS-232 and GPIB interfaces. All instrument functions can be queried and controlled via the interfaces. For convenient debugging, characters received and sent via the interfaces can be viewed on the front panel. Several hardcopy options are available on the SR850. Screens can be dumped to a dot-matrix or LaserJet compatible printer through the standard Centronics printer interface. Displays can also be plotted on any HP-GL compatible plotter via GPIB or RS-232.

About DSP Lock-ins

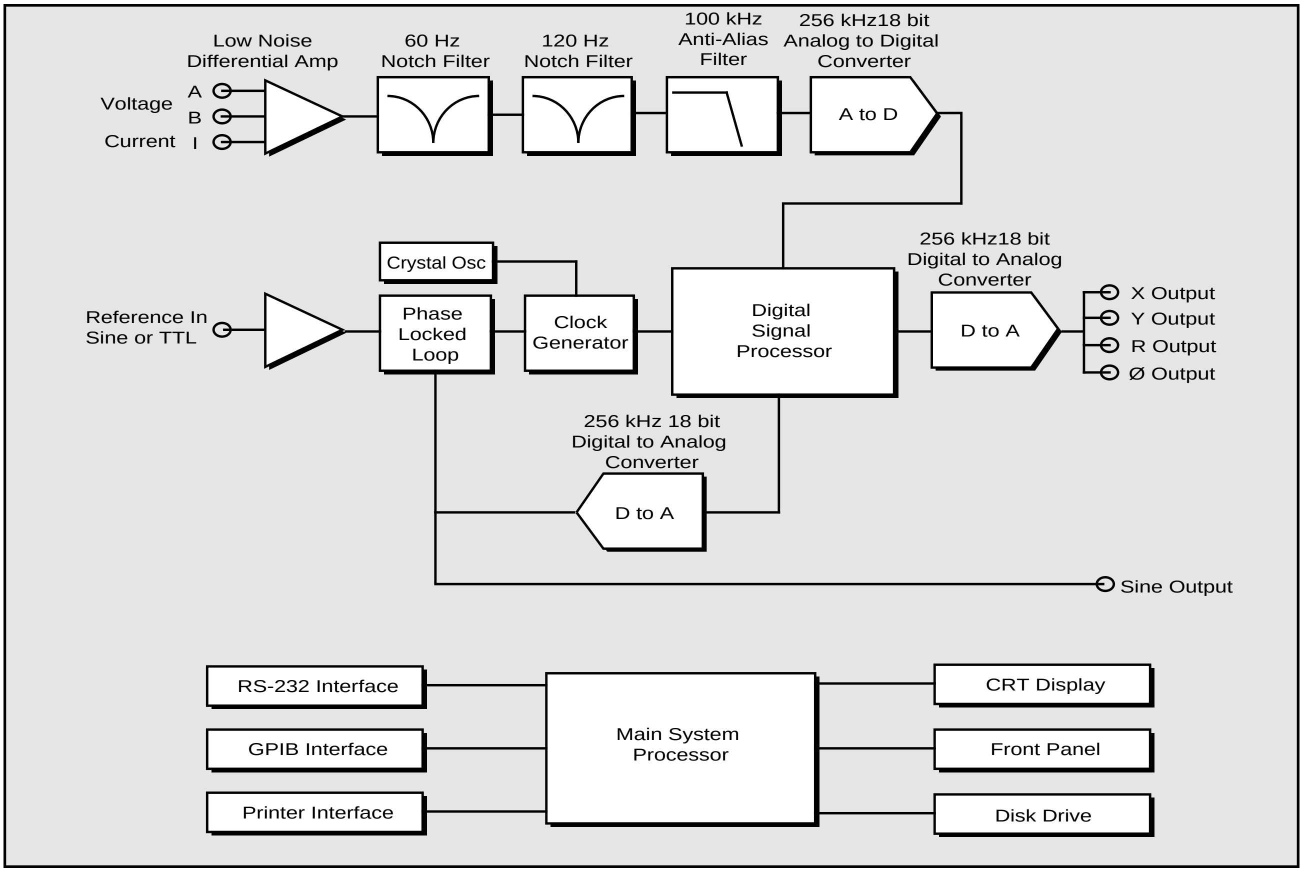
DSP lock-in amplifiers differ considerably from their analog counterparts as a glance at the block diagram below will quickly reveal. Although the front-end of both types of lock-ins contain a low-noise AC amplifier, and 60 Hz and 120 Hz line filters are provided in both cases, the similarity ends there. In the SR850 the signal is filtered with a 100kHz 9th-order elliptical anti-aliasing filter. This filtering is crucial to ensure that the signal can be digitized by the 256 kHz 18-bit A/D converter with no aliasing.

The A/D converter passes the digitized signal to the DSP chip. The DSP chip synthesizes a 24-bit digital reference sine wave at the reference frequency. In the internal-reference case, the reference is synthesized from a high-accuracy crystal oscillator. In the external-reference case, a phase-locked loop locked to the external reference serves as a source for the reference signal. The reference is multiplied by the signal in the DSP chip, which is capable of performing 16 million 24-bit x 24-bit multiplies and additions each second. After multiplication, up to four stages of digital lowpass filtering are applied to generate time constants from 10 µs to 30 ks, with filter rolloffs of 6, 12, 18, and 24 db/octave.

The resulting X and Y (in-phase and quadrature) signals are used to digitally calculate the values of R and θ. The results are sent to the main system processor for display on the CRT, and passed through an 18-bit digital to analog converter to generate the front-panel outputs.

The same digital reference that was used to multiply the signal is converted by another 18-bit D/A converter and is used as the SR850's internal oscillator. Thus, the internal oscillator output is actually the same signal as the reference, and there is virtually no reference phase noise when using it. Phase shifting the reference is also simple in this model, since only digital calculations are involved.

It is this digital architecture which makes possible the performance advantages found in the SR850. Comparing the diagram below with a block diagram of a conventional lock-in shows that many of the most troublesome components, noisy input prefilters, nonlinear demodulators, inaccurate analog filters, and drift-prone high gain DC amplifiers, have all been removed, along with their performance penalties. The resulting instrument comes closest to implementing the theoretical model of lock-in amplification.



Overview

The SR850 DSP lock-in amplifier is a high performance, full featured lock-in amplifier with a frequency range of 1 mHz to 102 kHz. Based on a novel digital signal processing based architecture, the SR850 offers performance specifications far beyond those achievable with conventional lock-in amplifiers. Up to100 dB of dynamic reserve can be achieved without the use of analog input prefilters. A 7" CRT display and a 65536 point data buffer makes possible a wide variety of storage and display options, including stripcharts, polar plots, and bargraphs. GPIB, RS-232, printer and plotter interfaces and a 3.5" disk drive are all standard.

Signal Channel

Voltage inputs	Single-ended or differential
Sensitivity	2 nV to 1 V
Current input	10^6 or 10^8 Volts/Amp
Input Impedance:	
Voltage Input	10 MΩ + 25 pf, AC or DC coupled
Current Input	1 kΩ to virtual ground
Gain accuracy	± 1%
Noise	6 nV/√Hz at 1 kHz (typical) 0.13 pA/√Hz at 1 kHz (10^6 V/A) 0.013 pA/√Hz at 100 Hz (10^8 V/A)
Line filters	60 [50] Hz and 120 [100] Hz notch (Q=5)
CMRR	90 dB at 100 Hz (DC coupled)
Dynamic reserve	>100 dB (without prefilters)

Reference Channel

Frequency range	0.001 Hz to 102 kHz
Reference input	TTL or sine (400 mVp-p minimum)
Input impedance	1 MΩ, 25 pf
Phase resolution	0.001°
Absolute phase error	< 1°
Relative phase error	< 0.001°
Orthogonality	90° ± 0.001°
Phase noise:	
Internal Reference	Synthesized, < 0.0001° rms at 1 kHz.
External reference	0.005° rms at 1 kHz, 100 ms, 12 dB/oct.
Phase drift	< 0.01°/°C below 10 kHz, < 0.1°/°C 10 kHz - 100 kHz.
Harmonic detection	2F, 3F, ... nF to 102 kHz.
Acquisition time	2 cycles + 5 ms or 40 ms (whichever is greater)

Demodulator

Stability:	
Digital outputs,display	no drift.
Analog outputs	< 5 ppm/°C for all dynamic reserves .
Harmonic rejection	-90 dB
Offset / Expand	± 100% offset. Expand up to 256x.
Time constants	10 µs to 30 ks (6, 12, 18, 24 dB/oct rolloff). Synchronous filtering available below 200 Hz.

Internal Oscillator

Range	1 mHz to 102 kHz
Accuracy	25 ppm + 30µHz
Resolution	0.01% or 0.1 mHz, whichever is greater.
Distortion	-80dBc (f<10kHz) -70dBc (f>10kHz) @ 1 Vrms amplitude.
Amplitude	0.004 to 5 Vrms into 10 kΩ (2mV resolution).
Output Impedance	50Ω
Amplitude accuracy	1%
Amplitude stability	50 ppm/°C
Outputs	Sine and TTL. (Both can be phase-locked to an external reference)
Sweeps	Linear and Log

Inputs and Outputs

Interfaces	IEEE-488, RS-232 and Centronics printer interfaces standard. All instrument functions can be controlled and read through the interfaces.
X, Y outputs	Sine and cosine components (± 10V). Updated at 256 ksamples/sec.
CH1 output	± 10V output of X, R, or Trace 1- 4 (each trace defined as AxB/C or AxB/C^2 where A, B, C are selected from X, Y, R, θ, X noise, Y noise, R noise, Aux 1-4 or frequency).
CH2 output	± 10V output of Y, θ or Trace 1- 4
Aux. A/D inputs	4 BNC inputs, 1 mV res., ± 10 V.
Aux. D/A outputs	4 BNC outputs, 1 mV resolution, ± 10 V, (fixed or swept amplitude).
Sine Out	Internal oscillator analog output.
TTL Out	Internal oscillator TTL output.
Trigger In	TTL signal either starts internal oscillator sweeps or triggers instrument data taking (rates to 512 Hz).
Remote pre-amp	Provides power to the optional SR550, SR552, and SR554 preamplifiers.

Displays

Screen format	Single or dual display
Displayed quantities	Each display shows one trace. Traces are defined as AxB/C or AxB/C^2 where A, B, C are selected from X, Y, R, θ, X noise, Y noise, R noise, Aux 1 - 4 or frequency
Display types	Large numeric readout with bar graph, polar plot or strip chart
Data buffer	64k data points can be stored and displayed as strip charts. The buffer can be configured as a single trace with 64k points, 2 traces with 32k points each, or 4 traces with up to 16k points each.
Sample Rate	0.0625 Hz to 512 Hz, or external up to 512 Hz

Analysis Functions

Smoothing	5, 9, 17, 21 or 25 point Savitsky-Golay.
Curve fitting	Linear, exponential or Gaussian
Calculator	Arithmetic, trigonometric and logarithmic calculations on trace region.
Statistics	Mean and standard deviation of trace region.

General

Hardcopy	Screen dumps to dot matrix or LaserJet printers. Plots to HP-GL compatible plotters (RS-232 or GPIB).
Disk drive	3.5 inch MS-DOS compatible format, 1.44 Mbyte capacity. Storage of data and instrument setups (binary or ASCII). Screens can be saved to disk as PCX files.
Rackmount	Included
Power	60 Watts, 100/120/220/240 VAC, 50/60 Hz.
Dimensions	17"W x 6.25"H x 19.5"L
Weight	40 lbs.
Warranty	One year parts and labor on all defects in materials and workmanship.

Application Information

Application note #3, “About Lock-in Amplifiers,” discusses the fundamentals of lock-in amplification and defines many of the terms used when discussing them.

Ordering Information

SR850	**DSP Lock-in Amplifier**	**$ 7500**
SR550	**FET Input Preamplifier**	**$ 495**
SR552	**Bipolar Input Preamplifier**	**$ 495**
SR554	**Transformer Prefamplifier**	**$ 995**
O760H	**Carrying Handle**	**$ 100**

SR850 Rear Panel

Digital Lock-in Amplifiers

Model SR810— DSP Lock-in Amplifier
Model SR830— DSP Lock-in Amplifier

- ***1 mHz to 102 kHz frequency range***
- ***100 dB dynamic reserve without pre-filtering (< 5 ppm stability)***
- ***Auto-gain, phase and reserve***
- ***Time constants from 10μs to 30 ks***
- ***6, 12, 18, 24 dB/oct rolloff***
- ***Harmonic detection (2F, 3F, ..., nF)***
- ***Standard GPIB and RS-232 Interfaces***

SR810/SR830 Overview

The SR810 and SR830 are the latest additions to the SRS family of DSP Lock-in amplifiers. Using the same revolutionary DSP technology pioneered by SRS in the SR850 lock-in, the SR810 and SR830 deliver unmatched digital performance at a price below that of most conventional lock-in amplifiers. While sharing most features and specifications, the SR810 and SR830 differ in a few important respects. In particular, the SR830 has two data displays, allowing the simultaneous display of X and Y, or R and Θ, while the SR810 has only a single data display (although the SR810 allows access to all parameters, including X,Y, R and Θ over the computer interfaces.)

Digital Performance

Like their predecessor, the SR850, the SR810 and SR830 offer performance unmatched by any analog lock-in. Both offer 100 dB of true dynamic reserve without the need for input prefiltering. Both offer time constants from 10 μs to 30 ks with 6, 12, 18 and 24 dB/octave filter rolloff. And both include a precision synthesized reference source with 25 ppm frequency accuracy and <-80 dBc distortion. Of course, RS-232 and GPIB interfaces capable of reading and setting all instrument functions are standard.

SR810/SR830 Features

Input Channel

The SR810 and SR830 both have a fully differential input with only 6 nV/√Hz of input noise at 1 kHz and an input impedance of 10 MΩ. The minimum full scale input voltage is 2 nV, and the lock-ins both include a current input amplifier with a switchable gain of 10^6 or 10^8 Volts/Amp. Line (50 Hz or 60 Hz) and 2x line filters are included to reduce line related interference, and the SR810's and SR830's digital architecture eliminate the need for input bandpass filters. Digital design makes it possible to obtain 100 dB of dynamic reserve without the noise and phase error introduced by input bandpass filters.

Digital Demodulator

The heart of the SR810 and SR830, like any lock-in amplifier, is the demodulator, which multiplies the input signal by the reference signal and filters the result. The demodulator of the SR810 and SR830 is a DSP processor capable of performing 16 million 24 bit multiplications and additions every second. In the case of the SR810 and SR830 the input is digitized at 256 kHz by a precision 18-bit A/D convertor. It is this digital demodulation technique which is at the root of so many of the SR810 and SR830's performance advantages. The digital technique eliminates the drift, distortion, and aging that are inherent in all analog demodulation designs and allow the SR810 and SR830 to achieve up to 100 dB of dynamic reserve. Dynamic reserve, a key figure of merit for lock-in amplifiers, is a specification often mentioned by lock-in manufacturers without offering a precise definition. At Stanford Research Systems, dynamic reserve is defined as follows:

The dynamic reserve of a lock-in amplifier at a given full-scale input voltage is the ratio (in dB) of the Largest Interfering Signal to the full-scale input voltage. The Largest Interfering Signal is defined as the amplitude of the largest signal at any frequency that can be applied to the input before the lock-in cannot measure a signal with its specified accuracy.

It is always important to keep in mind that many manufacturers do not use this conservative definition of dynamic reserve. Figures may be quoted for interfering signals that are outside the frequency range of the instrument and therefore easier to reject. Some manufacturers define the largest interfering signal as that necessary to overload the instrument, at which point the lock-ins output accuracy may be so degraded as to make the reading useless, The SR810 and SR830 both have up to 100 dB of **true** available dynamic reserve.

Digital Filtering

The digital signal processor also handles the task of output filtering, allowing time constants from 10 µs to 30,000 s to be selected, with a choice of 6, 12, 18 and even 24 dB/oct filter rolloff. For low frequency measurements (below 200 Hz), synchronous filters can be engaged to notch out multiples of the reference frequency. Since the harmonics of the reference have been eliminated (notably 2F), effective output filtering can be achieved with much shorter time constants. This is particularly useful at low frequencies, where the proximity of 2f components to the signal frequency often forces users of conventional analog lock-ins to employ very long time constants and consequently extend measurement times.

Digital Phase Shifting

The clumsy analog phase shifting circuits found in conventional lock-ins have been replaced in the SR810 and SR830 with a precise numerical calculation performed by the DSP processor. This allows phase to be measured with 0.01° resolution and the X and Y outputs to be orthogonal to 0.001°. This represents a significant improvement over analog instruments.

Frequency Synthesizer

The built-in direct digital synthesis (DDS) source generates a very low distortion (-80 dBc) reference signal. Single frequency sinewaves can be generated from 1 mHz to 102 kHz with 4 1/2 digits of frequency resolution. Both frequency and amplitude can be set from the front panel or from a computer. When using an external reference, the synthesized source is phase locked to the reference signal.

Easy Operation

Unlike some lock-in amplifiers, the SR810 and SR830 are simple to use. All instrument functions are set from the front panel keypad, and a spin knob is provided to quickly adjust parameters. The SR830 has two data displays which can be quickly configured to show X, Y, R, Θ, Xnoise, Ynoise or either of the two aux inputs. The SR810 has a single data display. Both the SR810 and SR830 have an additional LED display which can be set to display the reference frequency, phase, or amplitude.

Up to nine different instrument configurations can be stored in non-volatile RAM for fast and easy instrument setup. Standard RS-232 and GPIB (IEEE-488) interfaces provide communication with computers. All functions can be controlled and read through the interfaces.

Auto Gain, Phase and Offset

Auto-functions allow parameters needing frequent adjustment to automatically be set by the instrument. Gain, phase, offset and dynamic reserve are each quickly optimized with a single key press. The offset and expand features are useful when examining small fluctuations in a measurement. The input signal is quickly nulled with the auto-offset function, and resolution is increased by expanding around the relative value by up to 100 times. Harmonic detection is no longer limited to the 2F component. Digital design allows any harmonic (2F, 3F, ... nF) up to 102 kHz to be measured without changing the reference frequency.

Analog Inputs and Outputs

Both instruments have a user-defined output for measuring X, R, X-noise, Aux1, Aux 2 or the ratio of the input signal to an external voltage. The SR830 has a second user-defined output that measures Y, Θ, Y-noise, Aux 3, Aux 4 or ratio. The SR810 and SR830 both have X and Y analog outputs (rear panel) that are updated at 256 kHz. Four auxiliary 16-bit inputs are provided for general purpose use and can be read from the front panel or the computer interface. Four programmable 16 bit outputs provide voltages from -10.5 V to +10.5 V and can be set via the front panel or computer interfaces.

Internal Memory

The SR810 has an 8,000 point memory buffer for recording the time history of a measurement at rates up to 512 samples/sec. The SR830 has two 16,000 point buffers to simultaneously record two measurements, like R and Θ. Data is transferred from the buffers using the computer interfaces. A trigger input is also provided to externally synchronize data recording.

Absolute Value

The SR810 and SR830 DSP Lock-In Amplifiers from Stanford Research Systems offer outstanding performance, features, and value. Specification by specification, feature by feature, no other lock-ins can compare.

SR810/SR830 Rear Panel

Signal Channel

Voltage inputs	Single-ended or differential
Sensitivity	2 nV to 1 V
Current input	10^6 or 10^8 Volts/Amp
Impedance	Voltage: 10 MΩ + 25 pf, AC or DC coupled Current: 1 kΩ to virtual ground
Gain accuracy	± 1%
Noise	6 nV/√Hz at 1 kHz (typical) 0.13 pA/√Hz at 1 kHz (10^6 V/A) 0.013 pA/√Hz at 100 Hz (10^8 V/A)
Line filters	60 [50] Hz and 120 [100] Hz notch (Q=4)
CMRR	90 dB at 100 Hz
Dynamic reserve	> 100 dB without prefilters (< 5 ppm/°C)

Reference Channel

Frequency range	0.001 Hz to 102 kHz
Reference input	TTL or sine (400 mVp-p minimum)
Input impedance	1 MΩ, 25 pf
Phase resolution	0.01° front panel, 0.008° through computer interfaces.
Absolute phase error	< 1°
Relative phase error	< 0.001°
Orthogonality	90° ± 0.001°
Phase noise:	
Internal Ref.	Synthesized, < 0.0001° rms at 1 kHz.
External Ref.	0.005° rms at 1 kHz, 100 ms, 12 dB/oct.
Phase drift	< 0.01°/°C below 10 kHz, < 0.1°/°C below 100 kHz.
Harmonic detection	2F, 3F, ... nF to 102 kHz (n<19,999).
Acquisition time	2 cycles + 5 ms or 40 ms (largest applies)

Demodulator

Stability:	Digital outputs and display: no drift. Analog outputs: < 5 ppm/°C for all dynamic reserve settings.
Harmonic rejection	-90 dB
Time constants	10 µs to 30 ks (6, 12, 18, 24 dB/oct rolloff). Synchronous filters available below 200 Hz.

Internal Oscillator

Range	1 mHz to 102 kHz
Frequency accuracy	25 ppm + 30 mHz
Frequency resolution	4 1/2 digits or 0.1 mHz, whichever greater
Distortion	- 80 dBc (f<10kHz), -70 dBc (f>10kHz) @ 1 Vrms amplitude.
Amplitude	0.004 to 5 Vrms into 10 kΩ (2 mV resolution). 50Ω output impedance. 50 mA maximum current into 50 Ω.
Amplitude accuracy	1%
Amplitude stability	50 ppm/°C
Outputs	Sine, TTL.

Displays

Channel 1	4 1/2 digit LED display with 40 segment LED bar graph. X, R, X-noise, Aux 1 or Aux 2. The display can also be any of these quantities divided by Aux 1 or Aux 2.
Channel 2 (SR830)	4 1/2 digit LED display with 40 segment LED bar graph. Y, Q, Y-noise, Aux 3 or Aux 4. The display can also be any of these quantities divided by Aux 3 or Aux 4.
Offset	X, Y, R can be offset ±105% of full scale.
Expand	X, Y, R can be expanded by 10 or 100.
Reference	4 1/2 digit LED display.

Inputs and Outputs

CH1 output	± 10V output of X, R, X noise, Aux 1 or Aux 2. Updated at 512 Hz.
CH2 output (SR830)	± 10V output of Y, Q, Y noise, Aux 3 or Aux 4. Updated at 512 Hz.
X, Y outputs	In phase and quadrature components (± 10V). Updated at 256 kHz. (Rear panel)
Aux. A/D inputs	4 BNC inputs, 16 bit, ± 10 V, 1 mV resolution, sampled at 512 Hz.
Aux. D/A outputs	4 BNC outputs, 16 bit, ± 10 V, 1 mV resolution.
Sine Out	Internal oscillator analog output.
TTL Out	Internal oscillator TTL output.
Data Buffer	8,000 points. (SR810) 2 x 16,000 points (SR830)
Data Storage Rate	Up to 512 Hz
Trigger In	TTL signal synchronizes stored data recording.
Remote pre-amp	Provides power to the optional SR550, SR552, and SR554 preamplifiers.

General

Interfaces	GPIB and RS-232 interfaces standard. All instrument functions can be controlled and read through GPIB or RS-232 interfaces.
Power	40 Watts, 100/120/220/240 VAC, 50/60 Hz.
Dimensions	17"W x 5.25"H x 19.5"L
Weight	23 lbs.

Ordering Information

Introductory Prices:

SR810	**Single Phase DSP Lock-in**	**$3650**
SR830	**Dual Phase DSP Lock-in**	**$3950**
SR550	**FET Input Preamplifier**	**$ 495**
SR552	**Bipolar Input Preamplifier**	**$ 495**
SR554	**Transformer Preamplifier**	**$ 995**

RF Lock-in Amplifier

Model SR844 — 25 kHz to 200 MHz Lock-in Amplifier

- ***25 kHz to 200 MHz Frequency Range***
- ***Uses DSP Lock-in Technology***
- ***Time Constants from 100μs to 30ks.***
- ***6, 12, 18, 24 dB/octave roll-offs.***
- ***Auto-gain, Auto-sensitivity, Auto-offset***
- ***Auto-ranging, Auto-threshold External Reference***
- ***Internal or External Reference Source***
- ***X, Y, R, and θ Measurements***
- ***Manual Scan Capability***
- ***Readings in Volts or dBm***
- ***Settling Time Indicator***
- ***Save and Recall 9 Instrument Setups***
- ***16-bit Auxiliary Inputs (2) and Outputs (2)***
- ***GPIB and RS-232 Interfaces Standard***

SR844 Overview

The SR844 is the world's first self-contained 200 MHz lock-in amplifier. Unlike simple RF down-converters, which require an additional lock-in amplifier to operate, the SR844 is a true RF lock-in amplifier which uses the latest DSP technology from Stanford Research Systems. The SR844 seamlessly covers the frequency range from 25 kHz to 200 MHz—no range switching is ever necessary.

Digital Technology

The SR844's wide frequency range doesn't mean you'll have to sacrifice the performance and features you've come to expect from SRS lock-ins. Because it uses the same advanced DSP technology found in the SR850, SR810, and SR830, the SR844 gives you the same low drift, time constants from 100μs to 30ks, filter rolloffs of 6, 12, 18, and 24 dB/oct filter and a host of auto-setup features.

And of course, like all SRS lock-ins, the SR844 offers unmatched value. Even with its unsurpassed performance and features it's still priced at thousands less than other RF lock-in solutions.

SR844 Features

Signal Input

The SR844 offers a user-selectable input impedance of 50Ω, or 1 MΩ, for high impedance sources or in situations where it is desired to use a standard 10X scope probe to measure signals. Up to 60 dB of RF attenuation, or 20 dB of RF gain can be selected in 20 dB increments Full scale input sensitivity is adjustable from 100 nVrms (-127 dBm) to 1 Vrms (+13 dBm) in a 1/3/10 sequence. Separate selection of RF attenuation, gain, and input sensitivity allow the user to optimize gain allocation for maximum dynamic range or lowest possible noise.

Reference Channel

The SR844 includes a flexible reference channel that allows internal and external reference operation from 25 kHz to 200 MHz. The external reference automatically locks to frequencies over the entire operating range, eliminating the need for the user to select a specific frequency range. The external reference input includes an auto-threshold feature which automatically detects the maximum and minimum of the applied external reference and sets the threshold midway between the two, allowing the unit to operate with reference signals that are offset from 0V, or have a low duty-cycle. Sine-waves, square waves, and pulse trains (with pulse widths > 2ns), can all be used as external references.

The SR844 also includes an internally synthesized reference source. The internal source can be adjusted over the entire frequency range with 3-digit resolution. The reference signal is always available at the front-panel reference out BNC as a ±1V square wave.

Time Constants and Settling Time

Time constants from 100µs to 30 ks can be selected in 1/3/10 steps. Filter rolloffs of 6, 12, 18, and 24 dB/oct can be selected to maximize noise rejection. The SR844 includes a unique “settling time” indicator which counts time in units of the selected time constant, allowing you to normalize measurement times over a wide range of time constants and filter rolloffs.

Displays

Three bright, easy-to-read, 4 1/2 digit LED displays can be user-configured to display any measurement result. The Channel 1 display can show X, R(Volts or dBm), X-noise (normalized to a 1 Hz bandwidth), or the value of the Aux-1 input. The Channel 2 display can show Y, phase, Y-noise (normalized to a 1 Hz bandwidth), or the value of the Aux-2 input. The X, Y, R, and phase measurements can all be offset and expanded (x10 or x100). The reference display shows the reference frequency, phase, internal chopping frequency, or auxiliary output voltages.

Ratio Mode

The SR844’s ratio mode lets you divide the demodulated signal (see the “About RF Lock-ins” Technote) by the Aux-1 or Aux-2 input. Ratioing is performed before the time-constant filter is applied, so that fast signal fluctuations can be normalized out. If the ratio mode is on, the X and Y signals are both ratioed so that quantities computed from X and Y, such as R, X-noise, and Y-noise, are also ratioed.

Auto Features

The SR844 included four built-in “auto” features: autogain, auto-sensitivity, auto-offset and autophase. Auto gain optimizes the RF gain and attenuation. Auto-sensitivity causes the SR844 to automatically adjust the sensitivity and dynamic reserve based on the signal magnitude and any overload conditions. Auto-offset and phase quickly null the displayed magnitude or phase of the input signal allowing you to use the x10 or x100 expand to examine small changes around a nominal value.

Scan Capability

A unique “scan” feature is built into the SR844 which allows the instrument to scan up to 11 frequency points while in the internal reference mode. The instrument calculates the frequency points based on a geometric interpolation of a user entered start frequency, stop frequency, and number of points (up to 11). The unit will then automatically switch to the next precalculated frequency with a single keypress. In addition, the SR844 permits a baseline, or Cal, measurement to be taken at each scan frequency. The baseline measurements will be automatically subtracted from subsequent signal scans.

Inputs and Outputs

Two front panel BNC’s output a ±10V signal proportional to the quantity selected for the Channel 1 and Channel 2 displays. When the displayed quantity is X or Y, the update rate for the output is ≥ 48 kHz. If the displayed quantity is R or phase, the update rate is ≥ 10 kHz, while the noise measurement output update rate is 512 Hz. Two ±10V 16 bit auxiliary inputs, with input bandwidths of 1 kHz, as well as two ±10V auxiliary outputs with 1 mV resolution, are available on the rear panel.

GPIB and RS232

Of course, as with all SRS scientific instruments, GPIB and RS-232 interfaces are standard. All instrument features can be queried and set via the computer interfaces. 16 kPoints/channel of internal data storage is included to record the time history of a measurement at rates up to 512 samples/sec. Both binary and ASCII data transfer commands are included to provide maximum convenience and speed.

Outstanding Value

The SR844 represents a true breakthrough in lock-in amplifier design. No other lock-in comes close to matching its performance over its 200 MHz frequency range.

SR844 Rear Panel

About RF Lock-ins

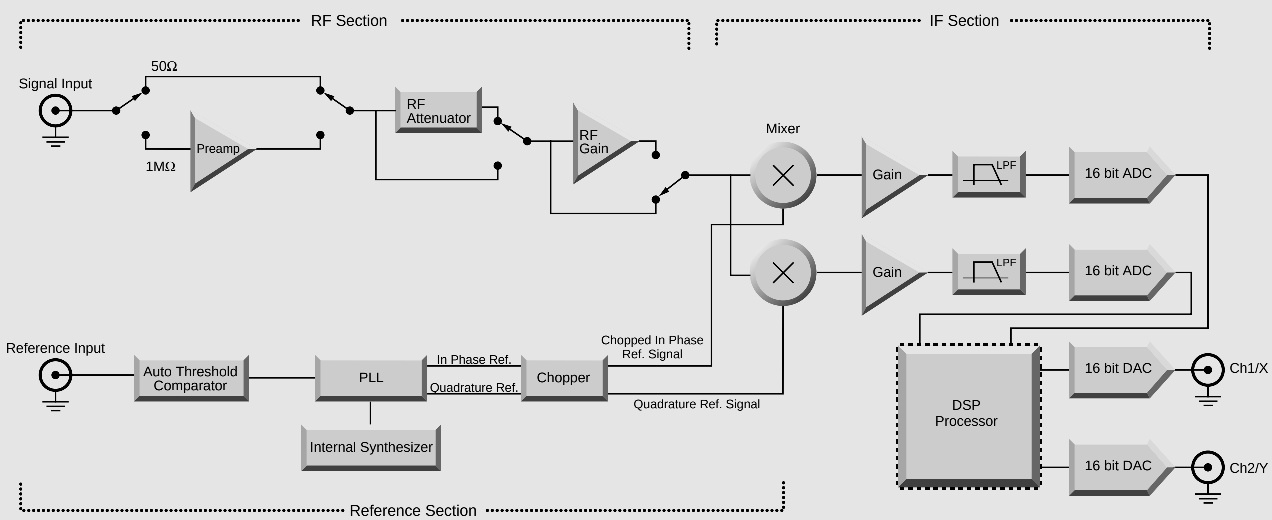
The SR844 RF lock-in amplifier utilizes a combination of analog and digital techniques to obtain maximum performance over a wide frequency range. Since it is not feasible to use pure digital techniques at the SR844's maximum 200 MHz operating frequency, analog down-conversion is used to transform the signal frequency to a range suitable for DSP processing.

A block diagram of the SR844 lock-in amplifier is shown below. The RF input signal passes through adjustable RF attenuators and gain stages depending on the selected input sensitivity. The signal is then mixed with two reference signals which differ in phase by 90°.

The reference signals are generated from either the external reference input, or the built-in frequency synthesizer using a phased-lock loop circuit (PLL) If the reference frequency were exactly at the signal frequency, the output of the mixers would be at DC. Since it is difficult to build drift and offset free amplifier chains, the SR844 chops the reference signals at a chopping frequency f_c which is chosen to be fast relative to the fastest time constants, yet slow relative to the input signal frequency. The IF amplifier and filter chain can now be AC coupled, eliminating DC offset and drift problems.

Once the in-phase and quadrature IF signals have been amplified and low-pass filtered, the signals are digitized by two precision 16-bit analog-to-digital converters. The digital IF signals are ratioed (if ratio mode is selected) and digitally low-pass filtered (with 6/12/18/24 dB/oct filter slopes) allowing the original signal amplitude and phase to be recovered

Note that the SR844 uses a square wave mixer, not a sine wave mixer like other SRS lock-in amplifiers. This is because precision sine generation is impractical with current technology over the SR844's operating range. The effect of using a square wave mixer is that the lock-in will respond not only at the reference frequency, but at all the Fourier components of the square wave reference. Since a square wave consists of odd harmonics with amplitudes 1/3, 1/5, 1/7, etc., the SR844 will respond at odd multiples of the reference frequency as well as at the reference frequency as itself. This limitation usually does not present a problem, as long as it is understood.



SR844 Block Diagram

Signal Channel

Voltage input	single-ended BNC
Input impedance	50 Ω or 1 MΩ + 30 pF
Damage Threshold	±5 V (DC+AC)
Bandwidth	25 kHz to 200 MHz
Sensitivity:	
< 1 MHz	100 nVrms to 1 Vrms full scale (-127 dBm to +13 dBm full scale)
<50 MHz	1 mVrms to 1 Vrms full scale
<200 MHz	10 mVrms to 1 Vrms full scale
Gain accuracy:	
< 50 MHz	± 0.25 dB
< 200 MHz	± 0.50 dB
Gain Stability	0.2% /°C
Coherent Pickup	Low Noise Reserve, Sens.<30mV
f < 10 MHz	<100 nV (typical)
f < 100 MHz	< 1.0 µV (typical)
f < 200 MHz	< 2.5 µV (typical)
Input noise: 50 Ω Input	
100 kHz < f < 100 MHz	2 nV/√Hz (typ.), 4 nV/√Hz (max.)
25 kHz < f < 200 MHz	<5nV/√Hz(typ.), <8 nV/√Hz (max.)
Input noise: 1 MΩ Input	
25 kHz < f < 200 MHz	5 nV/√Hz (typ.), <8 nV/√Hz (max.)
Dynamic reserve	up to 80 dB

Reference

External Reference Input	25 kHz to 200 MHz
Impedance	50Ω or 10 kΩ + 40 pF
Level	0.7 Vpp pulse or 0 dBm sine
Pulse Width	> 2ns at any frequency
Threshold Setting	Automatic, midpoint of waveform
Acquisition Time	<10s (auto-ranging,any frequency) < 1s (within same octave)
Internal Reference	25 kHz to 200 MHz
Frequency Resolution	3 digits
Frequency Accuracy	± 0.1 in the 3rd digit
Harmonic Detection	2F (signal from 50 kHz to 200 MHz)
Reference Outputs	Phase locked to Int. or Ext. reference
Front Panel Ref Out	25 kHz to 200 MHz square wave 1.0 Vpp nominal into 50Ω
Rear Panel TTL Out	25 kHz to 1.5 MHz, 0 to +5 V nominal, <3 V into 50 Ω
Phase Resolution	0.02°
Absolute Phase Error:	
< 50 MHz	< 2.5°
< 100 MHz	< 5.0°
< 200 MHz	< 10.0°
Rel. Phase Error, Orthog.	< 2.5°
Phase Noise (external)	0.005° rms at 100 MHz, 100 ms TC
Phase Drift	
< 10 MHz	< 0.1°/°C
< 100 MHz	< 0.25°/°C
< 200 MHz	< 0.5°/°C

Demodulator

Zero Stability	Digital displays have no zero drift. Analog outputs have < 5 ppm/°C drift for all dynamic reserve settings.
Filtering	
Time Constants	100 ms to 30 ks with 6, 12, 18 or 24 dB/octave roll-off
No Filter	10-20 µs update rate (X and Y outputs)
Harmonic Rejection	
Odd Harmonics	-9.5 dBc @3xRef, -14 dBc @ 5xRef, etc. (20log 1/n where n = 3, 5, 7, 9...)
Other Harmonics and sub-harmonics	< -40 dBc
Spurious Responses	-10 dBc @Ref ±2 x IF -23 dBc @Ref ±4 x IF < -30 dBc otherwise. (2 kHz < IF < 12 kHz)

Displays

Channel 1:	
Type	4.5 digit LED with 40 seg. bar graph
Quantities	X,R[V], R[dBm], X_{noise}, AuxIn 1
Channel 2:	
Type	4.5 digit LED with 40 seg. bar graph
Quantities	Y, q[deg], Y_{noise}[V], Y_{noise}[dBm], AuxIn 2
Expand	x10 or x100 for Ch1, Ch2 displays and outputs
Ratio	X and Y ratioed with respect to AuxIn 1 or AuxIn 2 before filtering and computation of R. The ratio input is normalized to 1 V and has a dynamic range greater than 100.
Reference:	
Type	4.5 digit LED
Quantities	Ref Freq, Phase, Offsets, AuxOut, IF Freq, Elapsed Time

Channel 1 and Channel 2 Outputs

Voltage Range	±10 V full scale proportional to X, Y or CH1, CH2 displayed quantity
Update Rate:	
X,Y	≤ 48 kHz
R, q, Aux Inputs	≤ 10 kHz
X noise, Ynoise	512 Hz

Auxiliary Inputs and Outputs

Inputs	2
Type	Differential, 1 MΩ
Range	±10 V
Resolution	1/3 mV
Bandwidth	3 kHz
Outputs	2
Range	±10 V
Resolution	1 mV
Data Buffers	Two 16,000 point buffers. Data is recorded at rates up to 512 Hz and is read using the computer interfaces.

General

Interfaces are can	IEEE-488 and RS-232 interfaces standard. All instrument functions be controlled and read through the interfaces.
Power	70 Watts, 100/120/220/240 VAC, 50/60 Hz.
Dimensions	17” W x 5.25” H x 19.5” D
Weight	23 lbs.
Warranty	One year parts and labor on materials and workmanship

Ordering Information

SR844	**RF Lock-in Amplifier (Rackmount Included)**	**$ 7950**

Lock-In Amplifiers

SR510 and SR530 — 0.5 Hz to 100 kHz Analog Lock-In Amplifiers

- ***Low Noise Current and Voltage Inputs***
- ***Sensitivity to 10 nV or 100 fA Full Scale***
- ***0.5 Hz to 100 kHz Reference Frequency***
- ***Dynamic Reserve up to 80dB***
- ***Tracking Bandpass and Line Notch Filters***
- ***Internal Reference Oscillator***
- ***Four A/D inputs, Two D/A outputs***
- ***GPIB and RS-232 Interfaces***

SR510/SR530 Overview

The SR510 and SR530 are analog lock-in amplifiers which can measure AC signals as small as nanovolts in the presence of much larger noise levels. Both the single phase SR510 and the dual phase SR530 have low noise voltage and current inputs, high dynamic reserve, two stages of time constants, and an internal oscillator. In addition, both lock-ins come equipped with a variety of features designed to make them simple to use and interface.

Sine Wave Conversion

The core of the SR510/SR530 is a precision analog sine wave multiplier. Lock-ins use a multiplier (demodulator) to translate the signal from the reference frequency down to DC, where it can be filtered and amplified. Older lock-ins use square wave multipliers, which introduce spurious harmonic responses. The SR510/SR530 use clean sine-wave multipliers which are inherently free of unwanted harmonics.

Packed With Features

Both the SR510 and SR530 come with standard RS-232 and GPIB computer interfaces. All instrument settings are stored in non-volatile RAM so that the lock-ins remember their setups when turned off. Four rear-panel A/D inputs and two D/A outputs let you read and provide voltages to the rest of your experiment. And a free PC-compatible software program lets you quickly get the lock-in running on a computer. For a combination of performance, features, and price, no analog lock-ins can compare to the SR510 and SR530.

SR510/SR530 Features

Signal Input

The SR510/SR530 have fully differential inputs with 7 nV/√Hz of input noise at 1 kHz and a 100 MΩ input impedance. The input can be configured as a voltage input, or as a current input with 10^6 Volts/Amp gain and an input impedance of 1 kΩ to virtual ground. Full scale sensitivities from 500 mV all the way down to 10 nV are available.

Three input prefilters can be selected. The first is a line notch filter providing 50 dB of rejection at the line frequency. The second filter similarly provides 50 dB of rejection at the second harmonic of the line frequency. The third filter is a bandpass filter which automatically tracks the reference frequency. These three filters can eliminate much of the noise in the signal before it is amplified.

Reference Input

The reference input can be set to lock to sine waves or to either edge of a pulsed reference. The reference frequency range is 0.5 Hz to 100 kHz, and detection at both the fundamental and second harmonic of the reference is allowed. A convenient, built-in frequency meter constantly measures and displays the reference frequency with four-digit resolution. The reference can be phase shifted with 0.025° resolution from the front panel, or shifted in 90° increments for easy measurement of quadrature signals. On the SR530 an autophase feature lets you quickly determine the phase of the signal relative to the reference with a single keypress.

6 and 12 dB/Octave Rolloff

Two stages of filtering after the phase sensitive detector provides variable filter rolloffs and time constants. Time constants can be chosen as long as 100 seconds for maximum noise reduction, or as short as 20 µs for use in real-time servo loops. By combining the two filter stages, rolloffs of 6 or12 dB/octave can be selected.

High Dynamic Reserve

Many manufacturers claim that their lock-ins have “high dynamic reserve” without specifying precisely what that means. At Stanford Research Systems, dynamic reserve is defined as follows:

The dynamic reserve of a lock-in amplifier at a given full-scale input voltage is the ratio (in dB) of the Largest Interfering Signal to the full-scale input voltage. The Largest Interfering Signal is defined as the amplitude of the largest signal at any frequency that can be applied to the input before the lock-in cannot measure a signal with its specified accuracy.

The SR510 and SR530 have a dynamic reserve of between 20 dB and 60 dB depending on the sensitivity scale. The dynamic reserve is automatically adjusted as the sensitivity changes, or for maximum flexibility, may be manually chosen. Selecting the bandpass filter adds an additional 20 dB of dynamic reserve making the maximum dynamic reserve for these lock-ins 80 dB. In general, however, in order to maximize output stability you should use the minimum dynamic reserve possible. (See the following Tech Note.)

It is always important to keep in mind each manufacturer’s definition of dynamic reserve when comparing specifications. Some manufacturers define the Largest Interfering Signal as the largest interfering signal that can be applied without overloading the instrument. However, long before the instrument becomes overloaded its accuracy may be severely degraded, making this number an overly optimistic estimate of the useful dynamic reserve. Some manufacturers quote dynamic reserve for interfering signals that are outside the frequency range of the instrument and are therefore easier to reject. These points must be taken into account before simply comparing two dynamic reserve specifications.

Offset and Expand

The SR510/SR530's offset and expand features make it easy to look at small changes in a large signal. Output offsets of 0% to 100% of full scale can be selected manually, or by using “auto-offset,” which automatically selects an offset equal to the signal value. Once the signal is offset, a x10 expand is available to provide increased resolution when looking at small changes from a nominal value.

Analog and Digital Displays

Precision, analog meters and four digit digital displays are standard on both lock-ins. On the SR510, you can select displays of the signal amplitude, the signal offset, or the measured noise. On the SR530, the first pair of displays show the signal components in rectangular form (X and Y), polar form (R and θ), the offset, noise, or the value of the rear-panel D/A outputs. The second digital display on both lock-ins can be configured to display either the reference phase or the reference frequency. Unlike other lock-ins, the signal and reference displays on the SR510 and the SR530 aren't multiplexed–you aren't forced to choose between displaying the signal parameters or the reference parameters.

Noise Measurement

The SR510/SR530's noise measurement feature lets you directly measure the noise in your signal at the reference frequency. Noise is defined as the RMS deviation of the signal from its mean. The SR510/SR530 will report the value of the noise in both a 1 Hz and 10 Hz bandwidth around the reference frequency.

Internal Oscillator

An internal voltage-controlled oscillator provides both a adjustable amplitude sine wave output, and a synchronous fixed amplitude reference output. The sine wave output amplitude is switch selectable between 0.01, 0.1, and 1 Vrms and can drive up to 20 mA. The oscillator frequency is contolled by a rear-panel voltage input, and can be adjusted between 1 Hz and 100 kHz. Typically, the sine wave output is used to excite some aspect of an experiment, while the reference output provides a frequency reference to the lock-in.

A/Ds and D/As

Four rear-panel A/Ds and two D/As provide an unprecedented level of flexibility in interfacing the SR510/SR530 with external signals. These input/output ports measure and supply analog voltages with a range of ±10.24 Vdc and a resolution of 2.5 mV. The A/Ds digitize signals at a rate of 1 kHz. The D/A output is ideal for controlling the frequency of the SR510/530's internal voltage controlled oscillator. A built-in ratio feature allows the SR510/SR530 to calculate the ratio of its output to a signal at one of the external A/D ports. This feature is important in servo applications to maintain a constant loop gain, or in experiments that normalize a signal to an intensity level.

Preamplifiers Available

Although the SR510 and SR530 are completely self contained and require no external preamplifiers for operation, sometimes an external remote preamplifier can be useful. Remote preamplifiers provide gain where it's most important, right at the detector, before the signal to noise ratio is permanently degraded by cable noise and pickup. The SR550 FET input preamplifier, the SR552 Bipolar input preamplifier, and the SR554 Transformer preamplifier are ideally suited for use with the SR510/SR530 lock-ins. These preamplifiers are especially useful when measuring extremely low-level signals. The SR550, SR552, and SR554 are described later in this section.

GPIB and RS-232

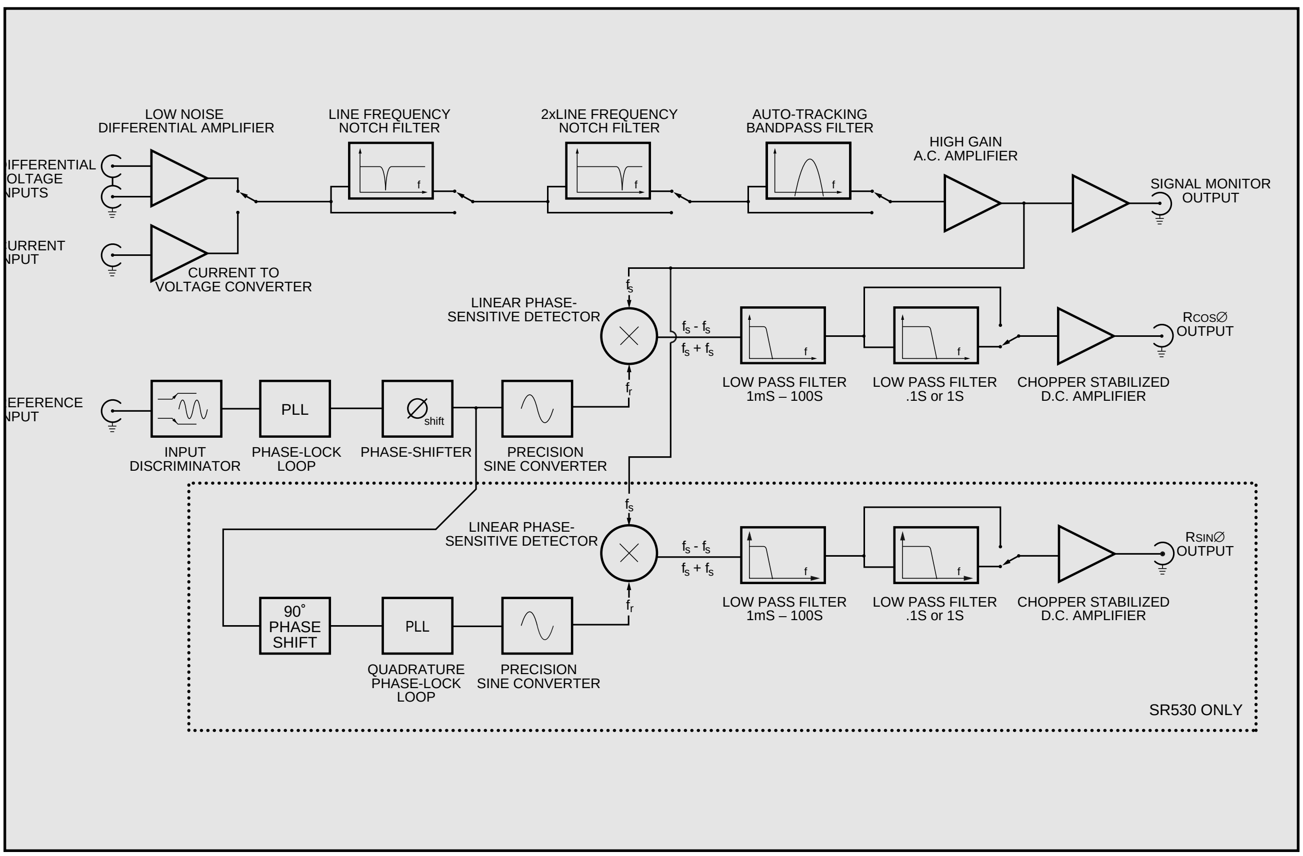
Both RS-232 and GPIB interfaces are standard on the SR510 and SR530. All features of the instruments (except the voltage controlled oscillator) can be queried and set via the computer interfaces.

Analog Lock-in Amplifiers

A block diagram of the SR510/SR530 analog lock-in amplifiers is shown below. The input signal is amplified by a low noise differential amplifier, selectively filtered to remove line frequency related interference and other unwanted signals, and amplified by a high-gain AC amplifier. The signal is then multiplied by a reference sine wave which is phase-locked to the reference input. The output of the multiplier contains frequency components near DC, (f_{signal} - $f_{reference}$) and 2f (f_{signal} + $f_{reference}$). In the SR530, a second multiplier multiplies the signal by a reference that has been phase shifted by 90°, allowing the lock-in to measure the in-phase and quadrature components of the signal simultaneously.

Two stages of lowpass filtering then provide the lock-in's "time constants." The purpose of the filtering is twofold. First, the filters remove the 2f components which are introduced by the multipliers. Secondly, the filters provide noise reduction by narrowing the lock-in's detection bandwidth. This is the essence of the lock-in technique. By only detecting signals in a narrow range of frequencies centered around the reference frequency, noise and interference at all other frequencies are rejected. The output of the filter stages is amplified by a chopper stabilized DC amplifier and becomes the lock-in's output.

The tradeoff between AC gain at the front-end of the lock-in and post-filter DC gain determines the dynamic reserve of the lock-in amplifier. If very little AC gain is used large interfering signals can be present without overloading the front-end. However, high DC gains must then be used which make the output more unstable. If the DC gain is lowered for more stability, higher AC gains must be used making the unit more susceptible to overloads. This tradeoff between dynamic reserve and stability is inherent to all analog lock-in amplifiers. The SR510 and SR530 allow you to manually select a dynamic reserve which is optimal for your experimental conditions.



Overview

THe SR510 and SR530 are single and dual phase analog lock-in amplifiers with a frequency range of 0.5 Hz to 100 kHz. Tracking bandpass filters combined with a sine-wave demodulator allow high dynamic reserves up to 80 dB. A low-noise front end, together with external low-noise preamplifiers produce input noise levels as low as 1.4 nV/√Hz. Two stages of output filtering provide time constants from 1 ms to 100 s and rolloffs of 6 and 12 dB/oct. RS-232 and GPIB interfaces are standard.

Signal Channel

Inputs	Voltage:Single-ended or True Differential Current:10^6 Volts/Amp
Impedance	Voltage:100 MΩ + 25 pF, ac coupled Current:1 kΩ to virtual ground
Full Scale Voltage	100 nV (10 nV on expand) to 500 mV
Sensitivity	Current:100 fA to 0.5 µA
Maximum Voltage	100 Vdc, 10 Vac damage threshold
Inputs	2 Vac peak-to-peak saturation Current:10 mA damage threshold 1 µA ac peak-to-peak saturation
Noise	Voltage:7 nV/√Hz at 1 kHz (typical) Current:0.13 pA/√Hz at 1 kHz (typical)
Common Mode	Range:1 Volt peak Rejection: 100 dB dc to 1kHz Above 1kHz the CMRR degrades by 6 dB/Octave
Gain Accuracy	1% (2 Hz to 100 kHz)
Gain Stability	200 ppm/°C
Signal Filters	60 Hz notch, -50 dB (Q=10, adjustable from 45 to 65 Hz) 120 Hz notch, -50 dB (Q=10, adjustable from 100 to 130 Hz)) Tracking bandpass set to within 1% of ref freq (Q=5)

Dynamic Reserve (DR):

Select	DR	Stability	Sensitivity Ranges
LOW	20 dB	5 ppm/°C	1 µV to 500 mV
NORM	40 dB	50 ppm/°C	100 nV to 50 mV
HIGH	60 dB	500 ppm/°C	100 nV to 5 mV

Bandpass filter adds 20 dB to dynamic reserve
Line Notch filters increase dynamic reserve to 100 dB

Reference Channel

Frequency	0.5 Hz to 100 kHz
Input Impedance	1 MΩ, ac coupled
Trigger:	
SINE	100 mV minimum, 1Vrms nominal
PULSE	±1 Volt, 1 µsec minimum width
Mode	Fundamental (f) or 2nd Harmonic (2f)
Acquisition Time	25 Sec at 1 Hz 6 Sec at 10 Hz 2 Sec at 10 kHz
Slew Rate	1 decade per 10 s at 1 kHz
Phase Control	90° shifts Fine shifts in 0.025° steps
Phase Noise	0.01° rms at 1 kHz, 100 msec, 12 dB TC
Phase Drift	0.1°/°C
Phase Error	Less than 1° above 10 Hz
Orthogonality*	90°±1°

Demodulator

Stability	5 ppm/°C on LOW dynamic reserve 50 ppm/°C on NORM dynamic reserve 500 ppm/°C on HIGH dynamic reserve
Time Constants:	
Pre	1msec to 100 sec (6 dB/Octave)
Post	1sec, 0.1 sec, none (6 dB/Octave)
Offset	Up to 1X full scale (10X on expand)
Harmonic Rejection	-55 dB (bandpass filter in)

Outputs & Interfaces

Channel 1 Outputs	X (RcosØ), X Offset, X Noise, R(magnitude)*, R Offset*, X5 (ext. D/A)*
Channel 2 Outputs*	Y (RsinØ), Y offset, Ø (phase), Y noise, X6 (ext. D/A)
Output Meter	2% Precision mirrored analog meter
Output LCD	Four digit auto-ranging LCD display shows same values as the analog meters
Output BNC	±10 V output corresponds to full scale input <1Ω output impedance
X Output*	X (RcosØ), ±10 V full scale, <1Ω output impedance
Y Output*	Y (RsinØ), ±10 V full scale, <1Ω output impedance
Reference LCD	Four digit LCD display for reference phase shift or frequency
RS-232	Interface controls all functions. Baud rates from 300 to 19.2 K
GPIB	Interface controls all functions. (IEEE-488 Std)
A/D	4 BNC inputs with 13 bit resolution (±10.24 V)
D/A	2 BNC outputs with 13 bit resolution (±10.24 V)
Ratio	Ratio output equals 10X signal output divided by the Denominator input.
Internal Oscillator:	
Range	1 Hz to 100 kHz, 1% accuracy
Stability	150 ppm/°C
Distortion	2% THD
Amplitude	1% accuracy, 500 ppm/°C stability 10 mV, 100 mV, 1Vrms

Front Panel Display Select

X, Y*	Signal Amplitude at the selected phase (AcosØ and AsinØ)
R*, Ø*	Magnitude and phase of signal
X OFST, Y OFST*	Display the offset which is being added to the signal output
X NOISE, Y NOISE*	Compute and display the noise on the signal

*SR530 Only

Analog Meters	Displays Signal, Offset, or Noise as a fraction of full scale
Output LCDs	Displays Signal, Offset, or Noise in absolute units
Output BNCs	Output follows Analog Meters, ± 10 V for ± full scale
Expand	Multiplies the Analog Meter and Output voltage by a factor X1 or X10.
REL	Set the Offset to null the output: subsequent readings are relative readings.
Offset	Enables or Disables Offset, and allows any offset (up to full scale) to be entered.
Time Constants	Pre-filter has time constants from 1 mS to 100 S (6 dB/Octave) Post-filter has time constants of 0, 0.1 or 1.0 S (6 dB/Octave)
ENBW	Equivalent Noise Bandwidth. Specifies the bandwidth when making noise measurements. (1 Hz or 10 Hz ENBW)
Reference Input	1 MΩ Input, 0.5 Hz to 100 kHz, 100 mV minimum
Reference Trigger	Trigger on rising edge, zero crossing, or falling edge
f/2f Mode	PLL can lock to either X1 or X2 of the reference input frequency
Phase Controls	Adjust phase in smoothly accelerating 0.025° steps, or by 90° steps. Press both 90° buttons to zero the phase.
Reference LCD	Display reference phase setting or reference frequency
Power Switch	Instrument settings from the last use are recalled on power-up

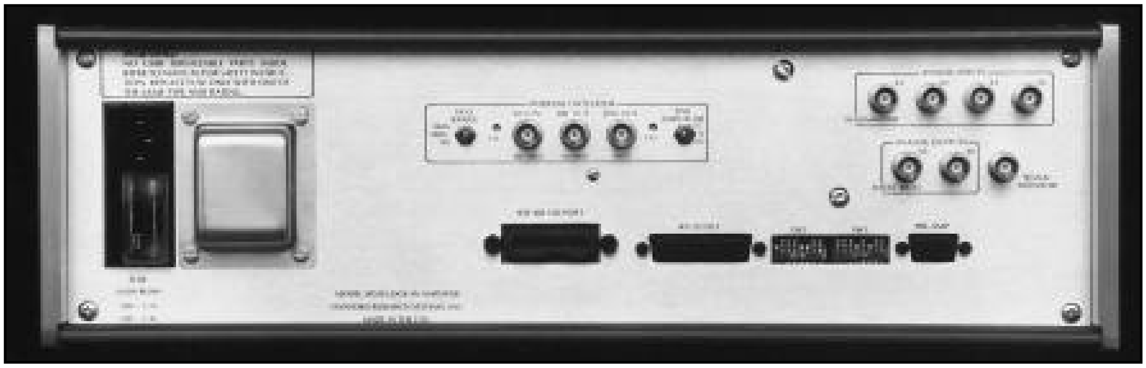
General

Power	100, 120, 220, 240 Vac (50/60 Hz); 35 W
Mechanical (SR510)	17" x 17" x 3.5" (Rack Mount Included) 12lbs.
Mechanical (SR530)	17" x 17" x 5.25" (Rack Mount Included) 16 lbs.

Ordering Information

SR510	**Single Phase Lock-in Amp.**	**$ 2495**
SR530	**Dual Phase Lock-in Amp.**	**$ 2995**
SR550	**FET Input Preamplifier**	**$ 495**
SR552	**Bipolar Input Preamplifier**	**$ 495**
SR554	**Transformer Preamplifier**	**$995**

SR510 Rear Panel



SR530 Rear Panel

Low Noise Preamplifier

Model SR554 — Low-noise Transformer Preamplifier

- ***0.1nV/√Hz Input Noise***
- ***> 40 dB Isolation from DC to 500 MHz***
- ***0.1 Hz to 40 kHz Bandwidth***
- ***Gain of 100 or 500***
- ***Transformer Coupled Input***
- ***Ideal for Low Impedance Sources***
- ***Single Ended or Differential Operation***
- ***Can be Powered From Any SRS Lock-in***

SR554 Overview

The SR554 is a low-noise transformer coupled preamplifier optimized for source impedances between 0.05Ω and 1 kΩ. With an input noise of only 0.1 nV/√Hz, the SR554 can be used in a wide range of low noise applications. It is the ideal preamplifier for low-temperature synchronous detection applications where isolation between the experimental sample and the lock-in amplifier is critical.

Choice of Two Gains

The SR554 can operate in one of two modes. In the "bypassed" mode the instrument is simply a passive transformer with a turns ratio of 100 providing an overall voltage gain of 100. In the "non-bypassed" mode an additional gain of 5 output amplifier is added, increasing the gain to 500 and providing a low impedance (<1Ω) output. The frequency response of the SR554 ranges from 0.1 Hz to over 40 kHz and depends on the source impedance and bypass setting as shown on the graph on the next page.

DC Powered

In the bypassed mode, the SR554 requires no external power. When using the output amplifier, ±20Vdc must be supplied. The preamplifier connector on any SRS lock-in can be used to power the SR554.

Input

Mode	Single Ended or Differential
Noise	0.1 nV/√Hz
Recommended Source Impedance	0.05Ω to 1kΩ
CMRR	>120 dB below 1 kHz
Common Mode Range	Primary can float to ±100Vdc

Amplifier

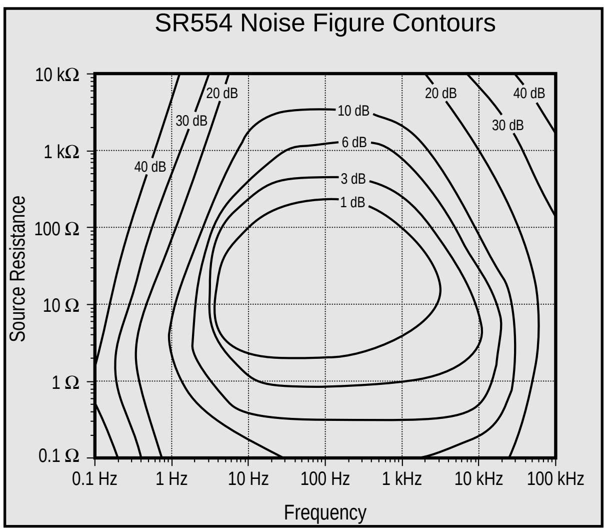
Gain:	
Transformer Only	100
Transformer and Buffer	500
Max Freq. Response	0.1 Hz to 40 kHz
Isolation	> 40 dB DC to 500 MHz

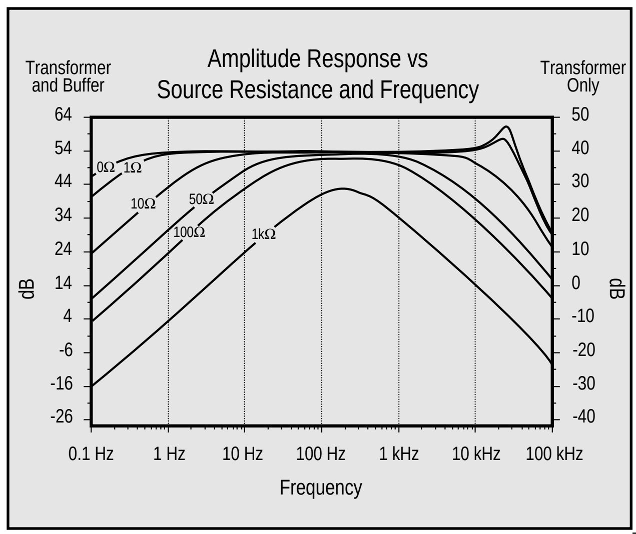
Output

Mode	Single Ended or Differential
Maximum Output:	
Transformer Only	100 Vpp
Transformer and Buffer	20 Vpp
Output Impdedance:	
Transformer Only	>5 kΩ
Transformer and Buffer	<1Ω

General

Size	3"H x 3.75"W x 7.5"D
Weight	4 lbs





Ordering Information

SR554	**Transformer Preamplifier**	**$ 995**

Lock-In Preamplifier

Model SR550 Low-noise FET Input Preamplifier

- ***Reduced Front-End Noise - 3.6 nV√Hz***
- ***FET input, 100 MΩ input impedance***
- ***Gains of 1, 2, 5, and 10***
- ***Single Ended or Differential Input***

Overview

The SR550 Remote Voltage Preamplifier is designed to work with any SRS lock-in amplifier. Preamplifiers are designed to provide gain close to the experimental detector, before the signal to noise ratio is permanently degraded by cable capacitance and pickup. The pre-amplifier minimizes noise and pickup in the connecting lines and reduces measurement time in noise limited experiments. When used with the SR510 or SR530, the preamplifier gain is automatically chosen to optimize noise performance without compromising dynamic reserve. When used with the SR810, SR830, or SR850, the preamplifier gain is automatically set to its maximum value. Power and control signals are brought from the lock-in by a nine pin cable (included). The SR550 may also be operated independently from the lock-ins by applying appropriate biasing (±20Vdc, +5Vdc).

SR550 Specifications

Input

Input Impedance	100 MΩ + 25 pF
Inputs	Single Ended or Differential (switch selectable)
Maximum Input	250 mV RMS for overload 100 Vdc, 10 Vac damage threshold
Noise	3.6 nV/√Hz at 1 kHz (typical) 4.0 nV/√Hz at 100 Hz 13 nV/√Hz at 10 Hz
Coupling	AC, 100 MΩ, .1 µF
Noise	See graph on next page

Common Mode

Range	1 V peak
Rejection	90 dB at 100 Hz

Gain

Settings	1,2,5,10 Automatically set by SR510 or SR530
Full Scale Sensitivity	10 nV to 200 mV
Gain Accuracy	2% (2 Hz to 100 kHz)
Gain Stability	100 ppm/°C

General

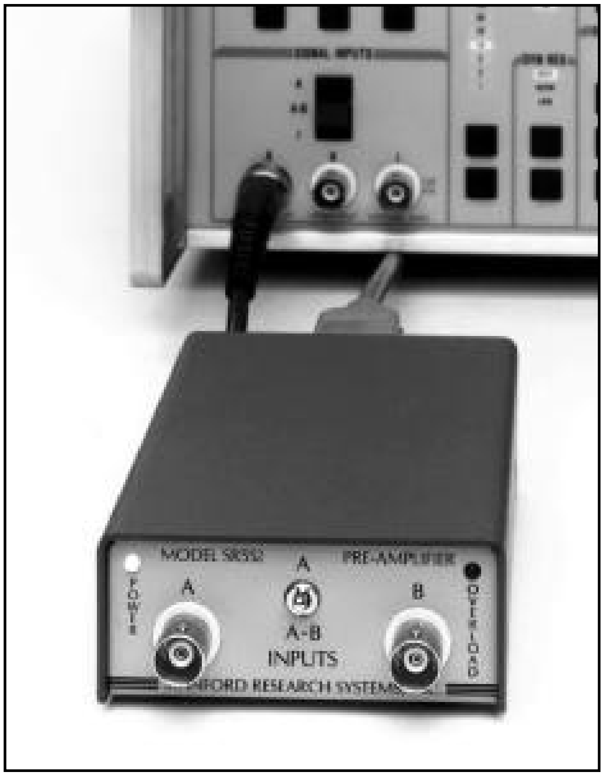
Outputs	(A) single ended (600Ω impedance) (B) shielded ground
Maximum Output	7 Volts peak to peak
Power	Supplied by SR510, SR530, SR810, SR830, or SR850 lock-in via connector cable.
Mechanical	1.3" X 3.0" X 5.1" , 1 lb.
Warranty	One year parts and labor on any defects in materials or workmanship

Ordering Information

SR550	**Low Noise Preamplifier**	**$ 495**

Lock-In Preamplifier

Model SR552 Low-noise Bipolar Input Preamplifier



- ***Reduced Front-End Noise — 1.4 nV/√Hz***
- ***Bipolar Input, 100 kΩ Input Impedance***
- ***Adjustable Gain — 10, 20, 50, 100***
- ***Single Ended or Differential Input***

Overview

The SR552 remote low-noise preamplifer is designed to work with any SRS lock-in amplifier, providing gain where it is needed most–right at the experiment. The preamplifier minimizes noise and pickup in the connecting lines and can reduce measurement time in noise limited experiments. The SR552 differs from the SR550 in its bipolar front-end design. The SR552 has a lower input noise and a correspondingly lower input impedence (100 kΩ vs 100 MΩ) than the SR550. When used with the SR510 or SR530, the preamplifier gain is automatically set by the lock-in. When used with the SR850, SR810, or SR830, the gain is always set to 100. Power and control signals are brought from the lock-in by a nine pin cable (included). The SR552 may also be operated independently by applying appropriate power supply voltages (±20Vdc, +5Vdc).

SR552 Specifications

Input

Input Impedance	100 kΩ + 25 pF
Inputs	Single ended or differential (switch selectable)
Maximum Inputs	70 mV rms for overload Damage threshold: 20 Vac, 50 Vdc
Noise	1.4 nV/√Hz at 1kHz (typical) 1.6 nV/√Hz at 100 Hz 2.5 nV/√Hz at 10 Hz
Coupling	AC, 100 kΩ and 100 µF
Noise	See graph on next page

Common Mode

Range	1 Volt peak
Rejection	110 dB at 100 Hz 100 dB at 1 kHz 80 dB at 10 kHz 60 dB at 100 kHz

Gain

Values	10,20,50,100 Automatically set by SR510 or SR530 lock-In depending on sensitivity and dynamic reserve.
Full Scale Input	10 nV to 200 mV
Gain Accuracy	1% (2 Hz to 100 kHz)
Gain Stability	200 ppm/°C

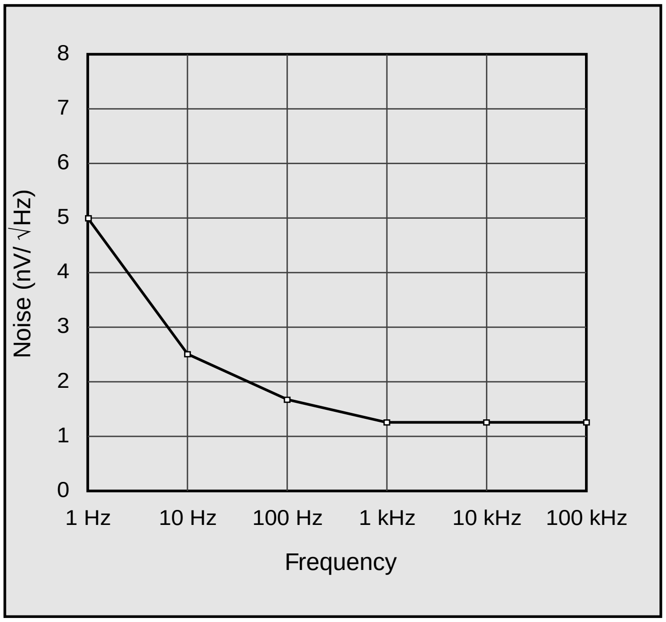
General

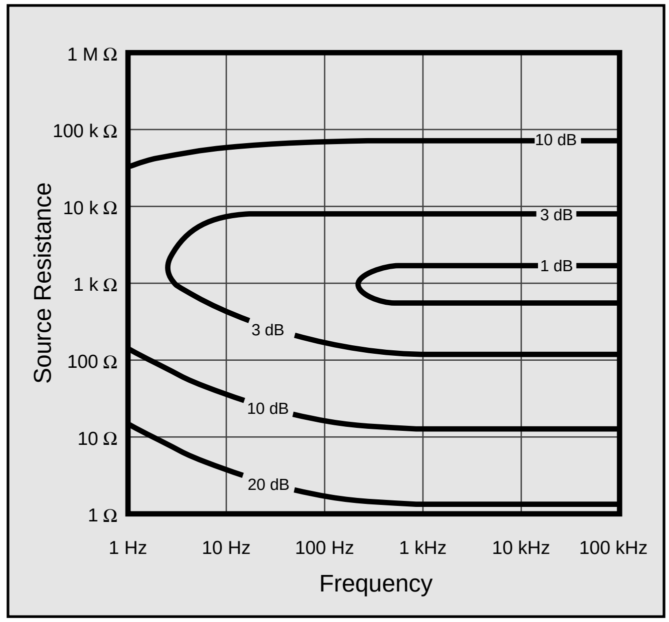
Outputs	(A) single ended (600 Ω impedance) (B) shielded ground
Maximum Output	10 Volts peak
Power	Supplied by SR510, SR530, SR810, SR830,or SR850 lock-In via control cable.
Warranty	One year parts and labor on any defects in material or workmanship

Ordering Information

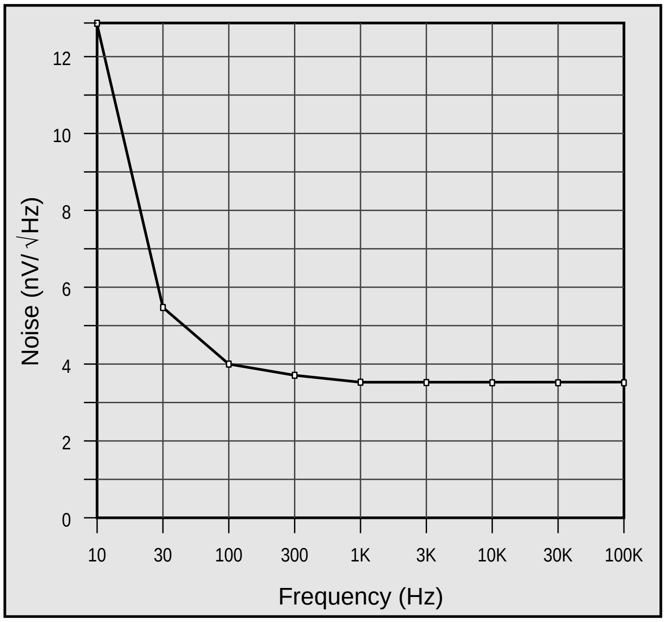
SR552	**Low Noise Preamplifier**	**$ 495**

SR552 Preamplifier Noise Specifications





SR550 Preamplifier Noise Specifications



Optical Choppers

Model SR540 — 4 Hz to 3.7 kHz Chopper

- ***4 Hz to 3.7 kHz Chopping Frequencies***
- ***Low Phase Jitter***
- ***Single and Dual Beam Experiments***
- ***Sum and Difference Reference Outputs***
- ***Bolt Clamp or Rod Mounting***

The SR540 chopper will handle all your optical chopping requirements–from simple experiments to dual beam and intermodulation measurements. The SR540 has a voltage control input, four-digit frequency display, ten-turn frequency control, and two reference outputs with selectable operating modes. Two anodized aluminum blades are provided: a 5/6 slot blade for frequencies up to 400 Hz and a 25/30 slot blade for frequencies up to 3.7 kHz. Reference outputs are provided for frequencies corresponding to each row of slots, as well as the sum and difference frequencies.

SR540 Specifications

Chop Frequency	4 Hz to 400 Hz with 5/6 slot blade 400 Hz to 3.7 kHz with 25/30 slot blade
Frequency Stability	250 ppm/°C (typical)
Frequency Drift	<2%, 100 Hz < f < 3700 Hz
Phase Jitter	0.2° rms from 50 Hz to 400 Hz 0.5° rms from 400 Hz to 3.7 kHz
Frequency Display	4-digit, 1 Hz resolution, 1 Hz accuracy
Frequency Control	10-turn pot with 3 ranges: 4 Hz to 40 Hz 40 Hz to 400 Hz 400 Hz to 3.7 kHz
Input Control Voltage	0 to 10 VDC for 0-100% of full scale. Control voltage overrides frequency dial.

Reference Modes:

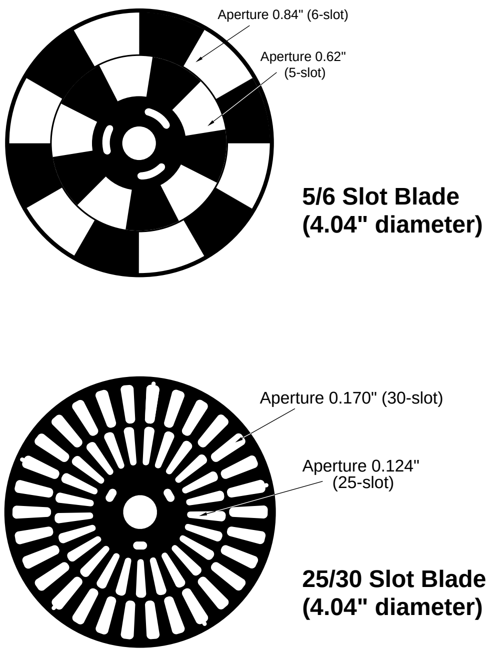
Switch	Left BNC	Right BNC
up	f_{inner}	f_{outer}
middle	$5xf_{outer}$	f_{outer}
down	$f_{inner}+f_{outer}$	$f_{outer}-f_{inner}$

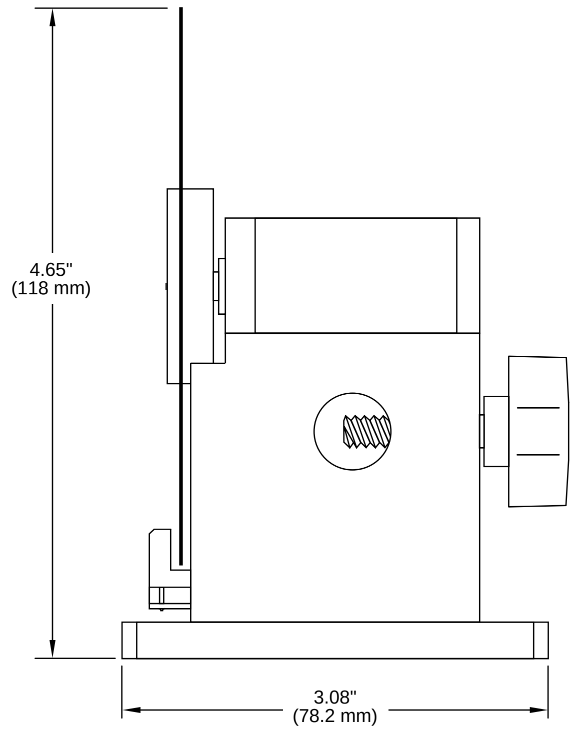
Dimensions:

Controller	7.7" x 5.1" x 1.8"
Chopper Head	2.8" x 2.1" x 1.0"
Blade Diameter	4.04" ± 0.002"
Cable Length	6 feet
Power	100/120/220/240 VAC 50/60 Hz, 12 Watts
Warranty	One year parts and labor on materials and workmanship. 90 days on motor.

Ordering Information

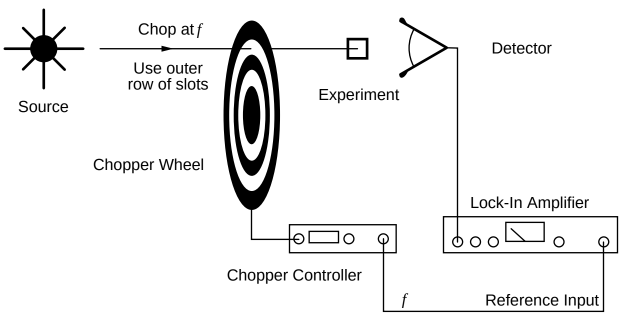
SR540	**Optical Chopper**	**$ 995**
0540RCH	**Replacement Chopper Head**	**$ 220**
0540BLD	**Replacement Chopper Blades (Specify Type)**	**$ 35**



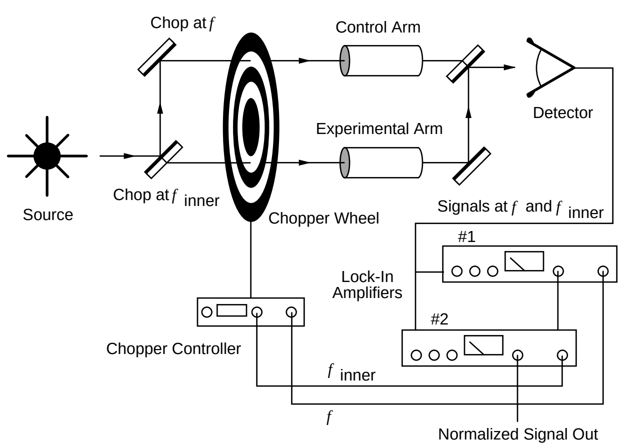


SR540 Applications

Single Beam Chopper Experiment
In this application, a single optical beam is chopped by the outer row of slots, and the reference output from the right BNC is used to lock the lock-in amplifier to the chop frequency. Note that the inner row of slots could be used, in which case the reference from the left BNC would be used. In either case, the REFERENCE MODE switch is in the "up" position.

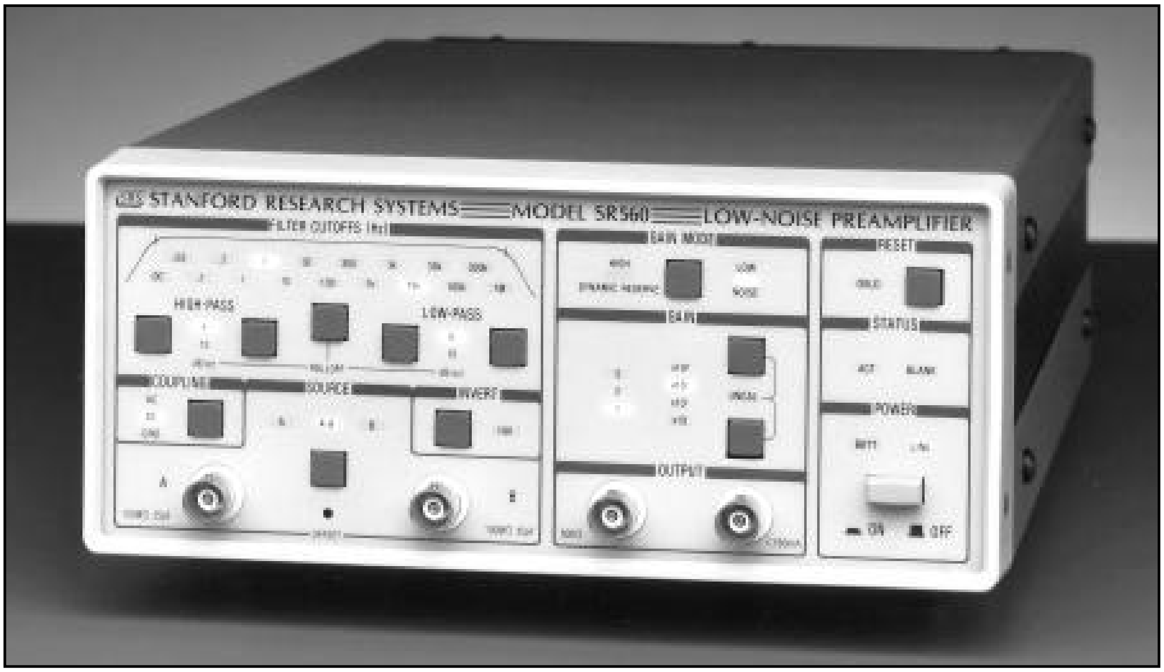


Dual Beam Chopper Experiment
In this arrangement, the output from a single source is split in two and chopped at two different frequencies by the two rows of chopper slots. One beam passes through the experiment, while the other is used as a reference beam. The beams are recombined and sent to the same detector. Two lock-ins are used to detect the signals at f_{inner}, corresponding to the experimental signal, and f_{outer}, corresponding to the reference beam. If the detected signal in the experimental arm is ratioed to the detected signal in the control arm, then effects due to changing source intensity and detector efficiency are removed.



Low Noise Preamplifier

Model SR560 — DC to 1 MHz Voltage Preamplifier



- ***4 nV/√Hz Input Noise***
- ***1 MHz Bandwidth***
- ***Variable Gain from 1 to 50,000***
- ***AC or DC Coupled***
- ***True Differential Or Single-ended Input***
- ***100 dB CMRR***
- ***Two Configurable Signal Filters***
- ***Selectable Gain Allocation***
- ***Line or Battery Operation***
- ***RS-232 Interface***

SR560 Overview

The SR560 is a high performance, low noise, general purpose preamplifier. With an input noise of only 4 nV/√Hz, a 1 MHz bandwidth, and a gain of up to 50,000, the SR560 is ideal for a wide variety of applications including low temperature measurements, optical detection, and audio engineering.

Signal Filters

Two configurable filters condition signals at frequencies from DC to 1 MHz. Choose flat, lowpass, bandpass, or highpass filtering to attenuate unwanted interference. Selectable gain allocation lets you further optimize performance for low noise or high dynamic reserve.

Intelligent Design

The microprocessor that runs the SR560 is “asleep” except during the brief interval it takes to change the instrument's setup. This ensures that no digital hash will contaminate your low-level analog signals. You can change settings from the front panel or via the standard RS-232 interface. The RS-232 interface allows a single computer to control up to four SR560s. The RS-232 interface is optoisolated to further isolate the analog circuitry from any source of digital noise.

Line or Battery Operation

For complete isolation from the power line, the SR560 may be operated from its internal batteries for up to 15 hours. The battery voltage is regulated to ensure that amplifier performance is never degraded as the batteries discharge. Internal recharging circuits always maintain a fully charged battery when the unit is plugged into the line and automatic discharge detection circuitry prevents battery damage when the unit is left on with the batteries.

SR560 Features

Inputs

The SR560 has a fully differential front end with 4 nV/√Hz of input noise and an input impedance of 100 MΩ // 25 pf. Complete noise specifications, including noise figure contours, are shown on the following pages. The SR560's inputs are fully floating, i.e. the input BNC shields are not connected to the instrument's chassis ground. Both the amplifier ground and the chassis ground are available at rear-panel connectors for complete flexibility in grounding the instrument. Input offset nulling is accomplished by a front panel potentiometer, accessible with a small screwdriver.

In addition to the signal inputs, a rear-panel TTL blanking input lets you quickly turn off and on the instrument's gain. This is useful, for instance, to prevent overloading the amplifier during an interval when the input signal is known to be large. The gain turns off 5 µs after the TTL level goes high, and back on again within 10 µs after the TTL signal goes low.

Gains are selectable from 1 to 50,000 in a 1-2-5 sequence. An adjustable gain feature lets you specify the gain as a percentage of any of the fixed gain settings with 1 % resolution. Gain can be selectively allocated before the filters to optimize noise performance, or after the filters, to reduce susceptibility to overloads. (See the graph of dynamic reserve on the next page.)

Outputs

Two insulated output BNCs are provided: a 600Ω output and a 50Ω output. Both outputs are capable of driving 10 Vpp into their respective loads. Two rear panel power supply outputs provide up to 200 mA of ±12 VDC referenced to the amplifier ground. The outputs provide clean DC power for use as a bias source.

Filters

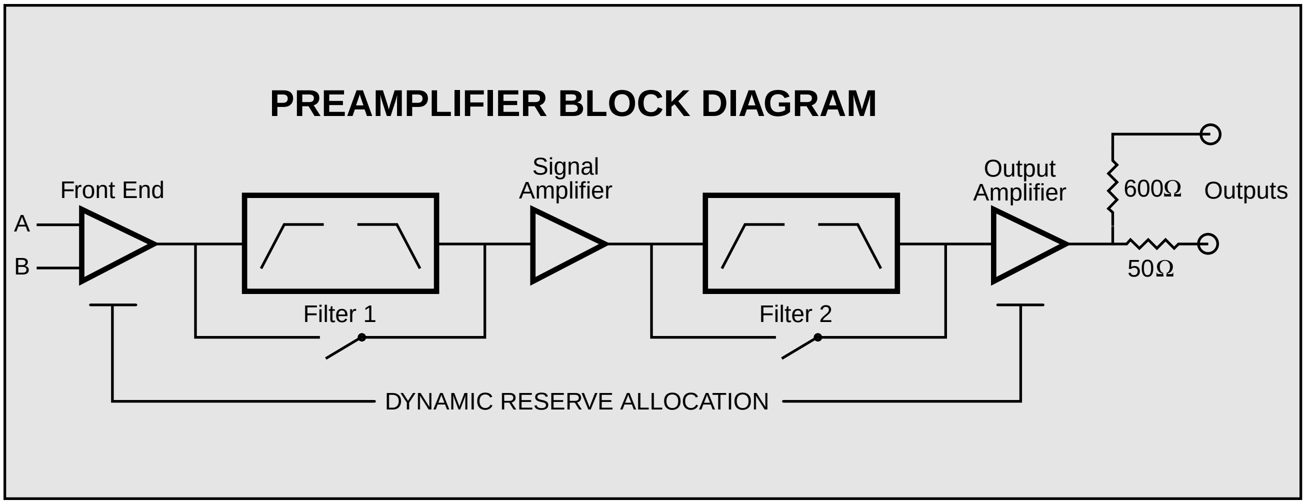
The SR560 contains two identical first order RC filters whose cutoff frequency and type can be configured from the front panel. Together, the filters can be configured as a 6 or 12 dB/oct lowpass or highpass filter, or as a 6 dB/oct bandpass filter. A filter reset button is included to shorten the overload recovery time of the instrument when long filter time constants are being used. Filter cutoff frequencies can be set in a 1-3-10 sequence from 0.03 Hz to 1 MHz.

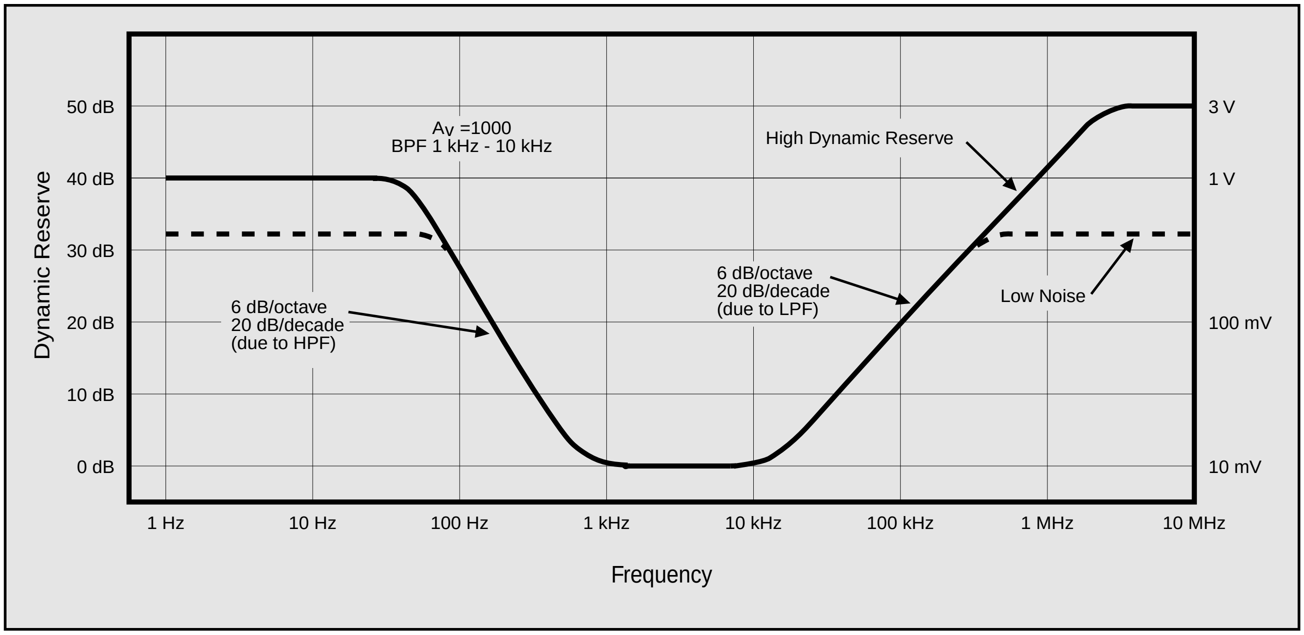
Battery Operated

Three internal, rechargable, lead-acid batteries provide up to 15 hours of battery powered operation. An internal battery charger automatically charges the batteries when the unit is connected to the line. The charger senses the battery state and adjusts the charging rate accordingly.

RS-232 Interface

The RS-232 interface allows listen-only communication with the SR560 at 9600 baud. Up to four SR560s can be controlled from a single computer, with each SR560 being assigned a unique address. A "Listen" command specifies which SR560 will respond to commands on the RS-232 line. All functions of the instrument (except power on) can be set via the RS-232 interface. The RS-232 interface electronics are optoisolated from the amplifier circuitry to provide maximum noise immunity.





Overview

The SR560 is a general purpose low noise preamplifier for amplifying and conditioning very small voltage signals. Flexibility, wide bandwidth, and high gain make this instrument perfect for audio, low temperature, and signal conditioning applications. The battery isolation removes any possible contamination from power lines, and the optoisolated design decreases contributed noise from any external pickup. An RS-232 interface is included for simple computer control.

Input

Inputs	AC or DC coupled, single-ended or differential
Input Impedance	100 MΩ + 25 pF
Maximum Input	3 V peak to peak
CMRR	100 dB from DC to 1 kHz. Decreases by 6dB/octave from 1 kHz to 1 MHz.
Noise	4 nV/√Hz at 1 kHz
Gain	1 to 50,000 ±1% in 1-2-5 sequence. Vernier gain in 0.5 % steps.
Frequency Response (gains up to 1000)	± 0.5 dB to 1 MHz (small signal) ± 0.2 dB to 300 kHz -3dB @ 1MHz @1Vpp output

Filters

Signal Filters	2 configurable (low pass or high pass) 6 dB/oct filters. -3 dB points are settable in a 1-3-10 sequence from 0.03 Hz to 1 MHz.
Gain Allocation	High Dynamic Reserve: Gain is increased after the signal filters to prevent overloading. Low Noise: Gain is increased before the filters to improve noise figure.

Output

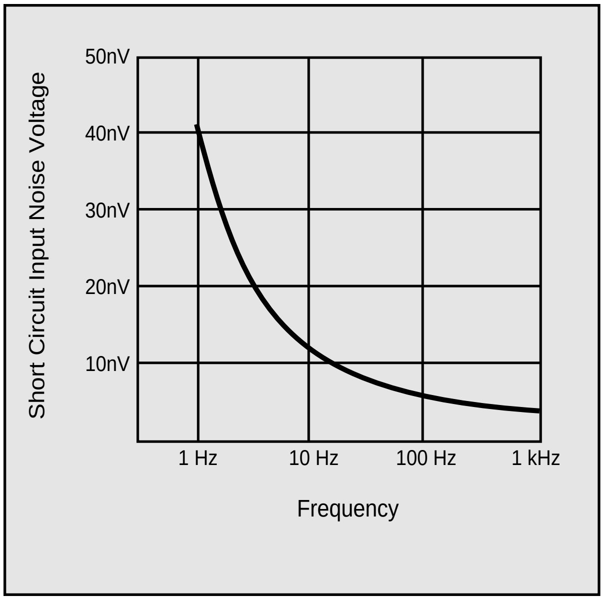
Maximum Output	10 V pp into 50Ω and 600Ω
Filter Reset	Long time constant filters may be reset with front panel button.
DC Drift	5 µV/°C referred to input (DC coupled)
Distortion	0.01% at 1 kHz
Rear Panel	±12 Vdc @200 mA referenced to amplifier ground

General

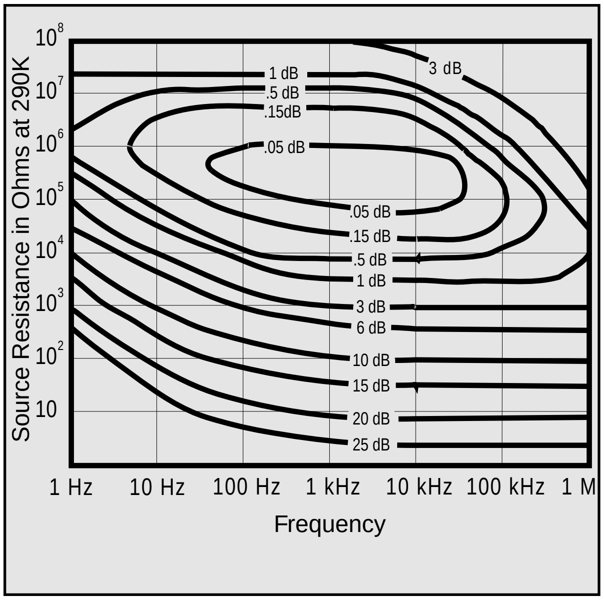
External Gating	TTL input sets gain to zero.
Interfaces	RS-232, 9600 BAUD, receive only.
Power	100,120,220, or 240 VAC from line. 6 W charged. 30 Watts while charging. Internal batteries provide 15 hours of operation between charges. Batteries are charged while connected to the line.
Dimensions	8.3" X 3.5" X 13.0"
Weight	15 lbs (batteries installed)
Warranty	One year parts and labor on any defects in materials or workmanship.

Ordering Information

SR560	**Low Noise Preamplifier**	**$ 1995**
O560RMS	**Single Rack Mount**	**$ 85**
O560RMD	**Dual Rack Mount**	**$ 85**
O560SB	**Spare Battery Set**	**$ 150**



SR560 Short Circuit Input Noise vs. Frequency



SR560 Noise Figure Contours

SRS Tech Note

About Noise Figure Contours

Noise figure contours are often provided with amplifier specifications but many users are unclear on what they signify. The noise figure of an amplifier is the ratio (usually expressed in dB) of the equivalent input noise of the amplifier at a given frequency to the thermal noise of a source with a given source impedance. Since the equivalent input noise of an amplifier is simply the output noise divided by the gain, this can be expressed as:

NF = 20 log(Output Noise/Gain)/(Source Thermal Noise)

What does this mean? Well, if the amplifier were noiseless, all the output noise would be due do the source thermal noise, and the noise figure would be 0 dB. To the extent that the amplifier adds some of its own noise, the Output Noise/Gain will be bigger than the source thermal noise and the noise figure will be non-zero. Thus, the noise figure is an indication of how much of the output noise the amplifier contributes.

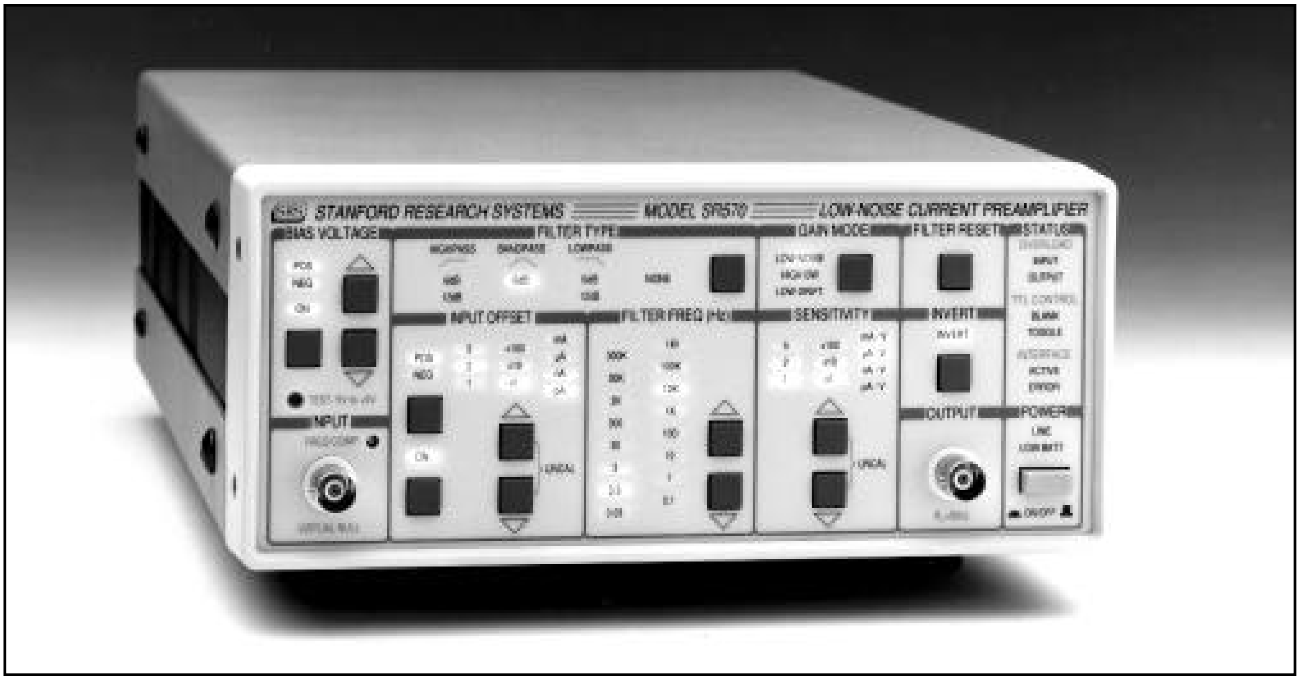
Since amplifier noise and thermal noise are functions of both frequency and source impedance, noise figures are often plotted as contours on a graph of source resistance versus frequency. At a given frequency, the noise figure is large at very low source resistance since the thermal noise of the source is small and any amplifier noise leads to a large noise figure. As the source resistance increases, the noise figure becomes better until the current noise of the amplifier passing through the high source resistance once again degrades the noise figure.

At a constant source resistance, the noise figure is poor at low frequencies because of the large 1/f noise of the amplifier. As the frequency is increased, 1/f noise decreases and the noise figure improves. Eventually, for all transistors, the current noise increases as a function of frequency, and the noise figure begins to increase again.

Does this mean that you should always increase the resistance of your source to reach the region of minimum noise figure? Certainly not. By doing that you would only be making your source noisier to improve the noise figure. Remember, the goal is to reduce total noise, not to minimize the noise figure.

Low Noise Current Amplifier

Model SR570 — Current Preamplifier



- ***5 fA/√Hz Input Noise***
- ***1 MHz Maximum Bandwidth***
- ***1 pA/V Maximum Gain***
- ***Adjustable Bias Voltage with Test Point***
- ***Variable Input Offset Current***
- ***Low Noise, High BW, and Low Drift Modes***
- ***Two Configurable Signal Filters***
- ***Rear Panel ±12 V Power Outputs***
- ***Line or Battery Operation***
- ***RS-232 Interface***

SR570 Overview

The SR570 is a low noise current preamplifier capable of current gains as large as 1 pA/V. High gain and bandwidth, low noise, and many convenient features make the SR570 ideal for a variety of photonic, low temperature, and other measurements.

Why a Current Amplifier?

Many people wonder why current amplifiers are necessary. Why not simply terminate a current source with a resistor and amplify the resulting voltage with a voltage preamplifier? The answer is twofold. To get a large voltage from a current, large resistors are necessary. In combination with cable capacitance and other stray capacitance, this can lead to unacceptable penalties in frequency response and phase accuracy. Current amplifiers have much better amplitude and phase accuracy in the presence of stray capacitance. Secondly, using resistive terminations forces the current source to operate into possibly large bias voltages–a situation which is unacceptable for many sources and detectors. Current amplifiers can sink current directly into a virtual null, or to a selected DC bias voltage.

Packed With Features

Two stages of configurable filters provide highpass, lowpass, or bandpass filtering to reject interference and noise. External blanking and toggle inputs can turn off or invert the gain with external TTL signals. Three rechargeable lead-acid batteries provide up to 15 hours of operation when line power is unavailable or undesirable.

Although microprocessor controlled, the SR570's processor "sleeps" except for brief intervals when the instrument's configuration is changed. Thus, low-level analog signals won't be contaminated by digital noise. An optoisolated RS-232 interface lets you control all the functions of the SR570 from a computer.

SR570 Features

Inputs

A block diagram of the SR570 is shown below. The input current can be offset to suppress any undesired DC background currents. Offset currents can be specified from ±1 pA to ±1 mA in roughly 0.1% increments. The voltage at the summing junctions of the amplifier is controlled by an adjustable ±5 V bias source. The default bias voltage is 0 V, making the SR570's input a virtual null. A front panel test point is provided to allow you to monitor the bias voltage.

The current to voltage conversion gain is controlled by the feedback resistor, R_f. Sensitivities from 1 pA/V to 1 mA/V can be selected in a 1-2-5 sequence, and a vernier adjustment option lets you select any sensitivity in between. Clearly, as the sensitivity increases and Rf gets larger, any capacitance present will cause the bandwidth of the instrument to suffer. The SR570 lets you choose between a “high bandwidth” mode, where Rf is chosen as small as possible, and the gain is made up later in the amplifier chain, and a “low noise” mode, where R_f is chosen as large as possible to optimize noise performance. Graphs of bandwidth vs. sensitivity are shown on the following pages.

Low Drift Mode

In addition to “low noise” and “high bandwidth” operation, a “low drift” mode is provided. This mode allocates gain just as the “low noise” mode, but replaces the ultra low noise op amp with one having a smaller input bias current, and hence smaller DC drift. Using the “low drift” mode reduces drift by up to a factor of 1000.

Toggle and Blanking

Two rear panel, optoisolated, TTL inputs modify the instrument’s gain characteristics. The blanking input lets you turn off the instrument’s gain with a TTL high level. This is useful, for instance, to avoid overloading the amplifier during periods of high input levels. The toggle input inverts the sign of the gain in response to a TTL signal, allowing you, for example, to perform synchronous detection with a chopped signal.

Filters

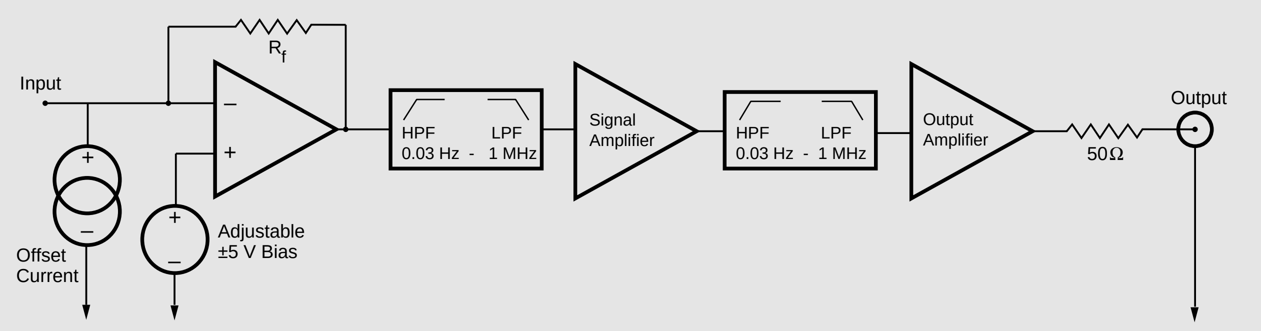
The SR570 contains two identical first order RC filters whose cutoff frequency and type can be configured from the front panel. Together, the filters can be configured as a 6 or 12 dB/oct lowpass or highpass filter, or as a 6 dB/oct bandpass filter. Cutoff frequencies are adjustable from 0.03 Hz to 1 MHz in a 1-3-10 sequence. A filter reset button is included to shorten the overload recovery time of the instrument when long filter time constants are used.

Battery Operated

Three internal, rechargeable, lead-acid batteries provide up to 15 hours of battery powered operation. An internal battery charger automatically charges the batteries when the unit is connected to the line. The charger senses the battery state and adjusts the charging rate accordingly. Rear panel LEDs indicate the charge state of the batteries.

RS-232 Interface

The RS-232 interface allows listen-only communication with the SR570 at 9600 baud. All functions of the instrument (except power on) can be set via the RS-232 interface. The RS-232 interface electronics are optoisolated from the amplifier circuitry to provide maximum noise immunity.



SR570 Block Diagram

Overview

The SR570 is a low noise current preamplifier for amplifying and conditioning small signals from current sources. The instrument is battery powered for optimum noise performance. Two configurable filters provide lowpass, highpass, and bandpass filtering. An RS-232 interface allows control by an external computer.

Input

Inputs	Virtual null or user set bias (-5 V to +5 V)
Input Offset	±1 pA to ±1 mA DC adjustable offset current
Maximum Input	±5 mA
Noise	See graphs on next page
Sensitivity	1 pA/V to 1 mA/V in 1-2-5 sequence Vernier sensitivity in 0.5 % steps.
Frequency Response	± 0.5 dB to 1 MHz Adjustable front panel frequency response compensation for source capacitance.
Grounding	Amplifier ground is fully floating. Amplifier and chassis ground are available at rear panel banana plug connectors. Input ground can float up to ±40V.

Filters

Signal Filters	2 configurable (low pass or high pass) 6 dB/oct filters. -3 dB points are settable in a 1-3-10 sequence from 0.03 Hz to 1 MHz.
Gain Allocation:	
Low Noise	Large feedback resistors are used for best noise performance
High Bandwidth	Smaller feedback resistors are used for better frequency response
Low Drift	Special low input bias current op amp is used for more accurate measurements at high sensitivity

Output

Gain Accuracy	±(0.5% of output + 10 mV) @25°C
Maximum Output	10 V pp into a high impedance
Filter Reset	Long time constant filters may be reset with front panel button.
Rear Panel	±12 VDC @200 mA, referenced to amplifier ground
DC Drift	see table below

General

External Blanking	TTL input sets gain to zero.
External Toggle	TTL input inverts gain polarity
Interfaces	RS-232, 9600 BAUD, receive only.
Power	100,120,220, or 240 VAC from line. 6 W charged. 30 Watts while charging. Internal batteries provide 15 hours of operation between charges. Batteries are charged while connected to the line.
Dimensions	8.3" X 3.5" X 13.0"
Weight	15 lbs (batteries installed)
Warranty	One year parts and labor on any defects in materials or workmanship.

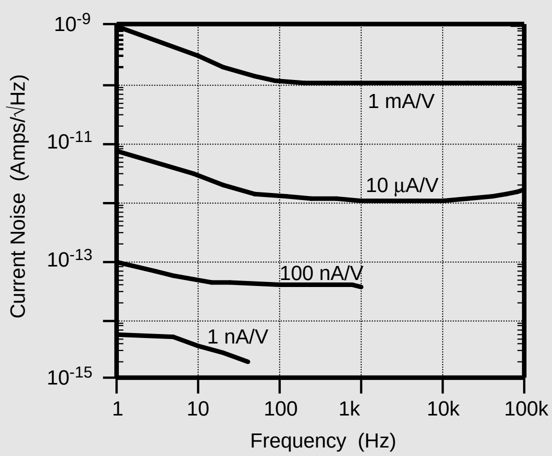
Ordering Information

SR570	**Current Preamplifier**	**$ 1995**
O570RMS	**Single Rack Mount**	**$ 85**
O570RMD	**Dual Rack Mount**	**$ 85**
O570SB	**Spare Battery Set**	**$ 150**

Sensitivity (A/V)	Bandwidth (3 dB) [1] High BW	Bandwidth (3 dB) [1] Low Noise	Noise/√Hz [2] Low Noise	Noise/√Hz [2] High BW	Temperature Coefficient Low Drift (11 ° - 28 °C) ±(%input + offset) /°C	DC Input Impedance
10^{-3}	1.0 MHz	1.0 MHz	150 pA	150 pA	0.01 % + 20 nA	1 Ω
10^{-4}	1.0 MHz	500 kHz	60 pA	100 pA	0.01 % + 2 nA	1 Ω
10^{-5}	800 kHz	200 kHz	2 pA	60 pA	0.001 % + 200 pA	100 Ω
10^{-6}	200 kHz	20 kHz	600 fA	2.0 pA	0.001 % + 20 pA	100 Ω
10^{-7}	20 kHz	2.0 kHz	100 fA	600 fA	0.001 % + 2 pA	10 kΩ
10^{-8}	2 kHz	200 Hz	60 fA	100 fA	0.003 % + 400 fA	10 kΩ
10^{-9}	200 Hz	15 Hz	10 fA	60 fA	0.022 % + 40 fA	1 MΩ
10^{-10}	100 Hz	10 Hz	5 fA	10 fA	0.024 % + 20 fA	1 MΩ
10^{-11}	20 Hz	10 Hz	5 fA	5 fA	0.038 % + 20 fA	1 MΩ
10^{-12}	10 Hz	10 Hz	5 fA	5 fA	0.040 % + 20 fA	1 MΩ

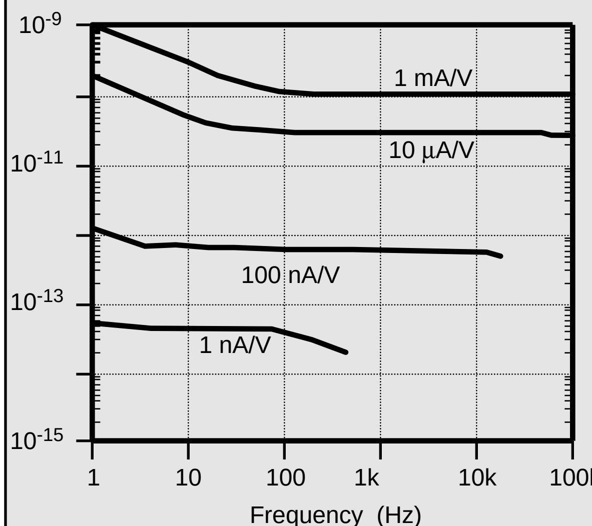
[1] Frequency Compensation adjusted for flat frequency response.
[2] Average noise in the freq. range below the 3 dB point but above the frequency where 1/f noise is significant.

Low Noise Mode



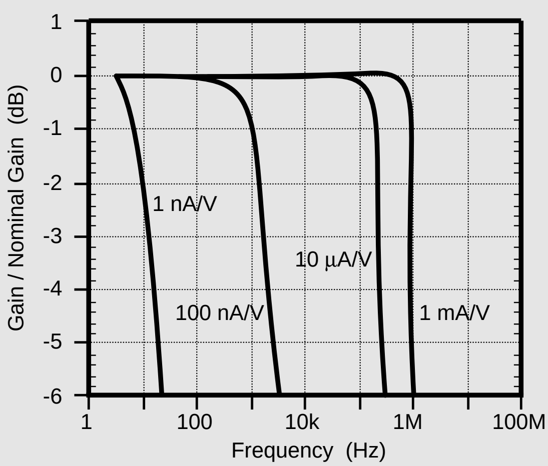
Current Noise vs. Frequency

High Bandwidth Mode



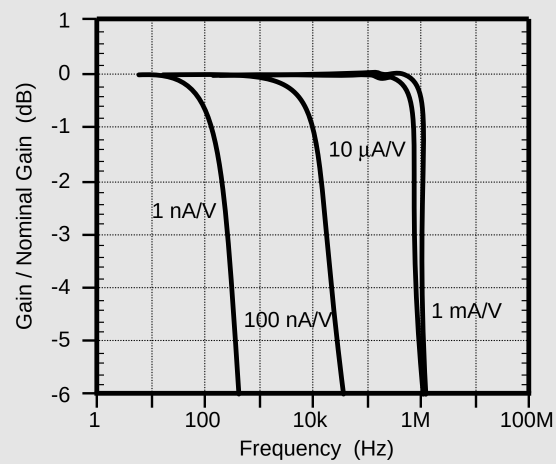
Current Noise vs. Frequency

Low Noise Mode



SR570 Bandwidth

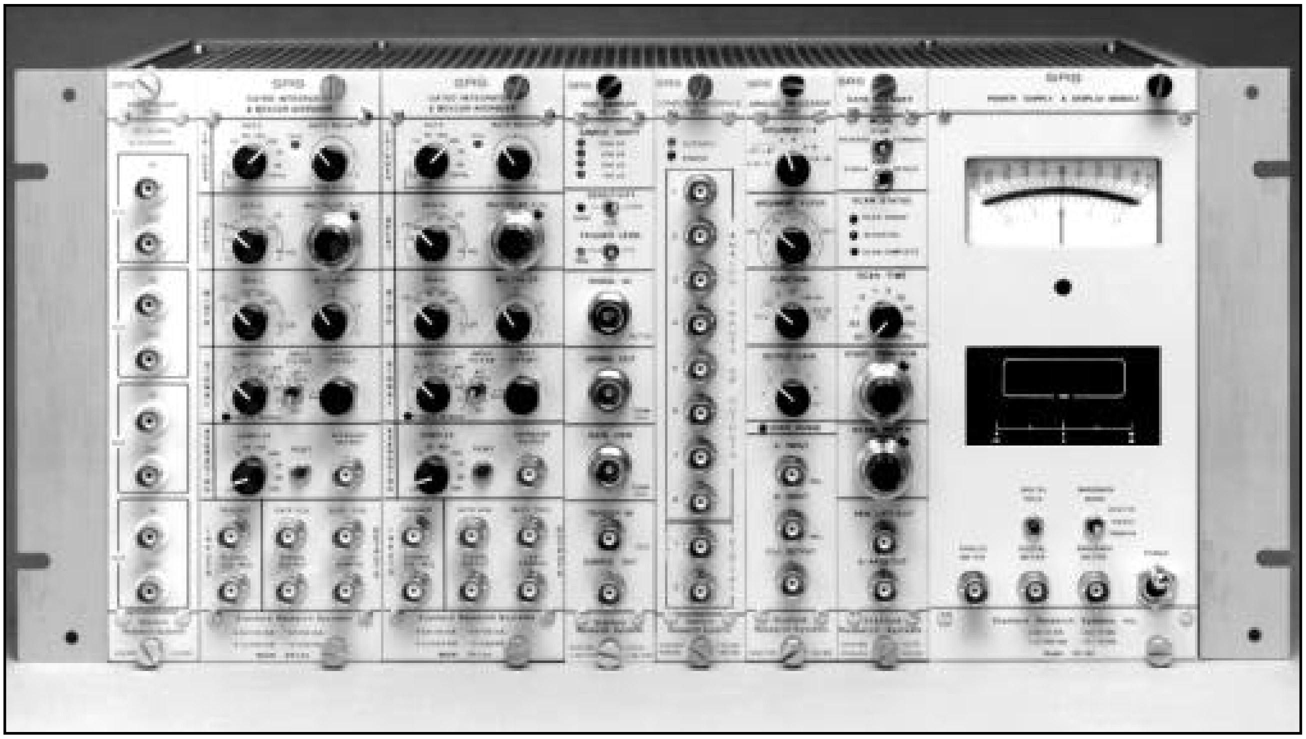
High Bandwidth Mode



SR570 Bandwidth

Gated Integrators and Boxcar Averagers

System SR250



System SR250 Overview

The SRS Boxcar Averager System SR250 is a modular instrumentation system designed to acquire and analyze fast analog signals. The system consists of a NIM compatible mainframe and modules which can be selected to tailor a system matched to your individual requirements. The system is flexible enough to handle gate widths from 100 ps to 150 µs, repetition rates from a fraction of a Hertz to 50 kHz, and output to computers or chart recorders. With its low-noise inputs, low-drift outputs, and unparalleled flexibility and modularity, the SR250 system continues to set an unmatched standard for gated integrators and boxcar averagers.

What is a Gated Integrator and Boxcar Averager?

Gated integrators and boxcar averagers are designed to recover fast, repetitive, analog signals with time scales of 100's of picoseconds to 100's of microseconds. In a typical application, a time "gate" is generated characterized by a set delay from an internal or external trigger and a certain width. A gated integrator amplifies and integrates the signal that is present during the time the gate is open, ignoring noise and interference that may be present at other times. Boxcar Averaging refers to the practice of averaging the output of the gated integrator over many shots of the experiment. Since any signal present during the gate will add linearly, while noise will add in a "random walk" fashion as the square root of the number of shots, averaging N shots will improve the signal-to-noise ratio by a factor of $\sqrt{N}$.

Stand Alone or Computer Controlled

System SR250 is designed to work without a computer, allowing you to quickly "tune-up" your experiment, or with a computer, automating complex measurement sequences. Modules like the SR235 Analog Processor perform signal processing functions such as ratioing and background subtraction without the need for additional software development, while the SR245 computer interface gives you complete access to boxcar measurements over GPIB or RS-232. No matter what your application, the System SR250 provides all the speed, accuracy, flexibility and convenience required for your fast analog data acquisition needs.

System SR250 Components

SR250 Gated Integrator Module

The SR250 Gated Integrator Module is the basic building block of System SR250. Each module provides a complete single channel gated integrator and boxcar averager with gate widths from 2 ns to 150 µs. You can average from 1 to 10,000 samples, and the "last sample" feature lets you bypass averaging and access the gated integrator output directly.

SR280 Mainframe and Display Module

The SR280 Mainframe and Display module provides a rack-mountable NIM bin which can house up to 9 single-width NIM modules. The three rightmost slots in the bin are occupied by a NIM compatible power supply and display which provides ±12 V and ±24V to the other NIM modules. The display module also contains an analog meter, digital meter, and bargraph meter useful for monitoring the output of the boxcar system.

SR275 Display Module

The SR275 dislay module is provided for users who already own a NIM bin and power supply. It contains the same three meters as the SR280, but has no power supply or NIM bin.

SR200 Gate Scanner

The SR200 Gate Scanner is designed to automate waveform recovery using System SR250 by providing the voltages needed to scan the SR250's gate delay. Scan times from 10 ms to 5 minutes can be selected. Outputs are provided to control chart recorders or oscilloscope displays.

SR235 Analog Processor

The SR235 Analog Processor performs a variety of convenient signal processing functions without the need for a computer. Its two inputs (A and B) can be combined to form an "argument" A, B, A-B, AB/10, 10A/B, or $\sqrt{(A^2+B^2)}$. The SR235 can then output a function F(x) of this argument corresponding to x, x^2, ln(x), dx/dt, or (dx/dt)/100.

SR240 Quad 300 MHz Preamplifier

The SR240 contains four DC-300 MHz amplifier channels, each with a gain of 5. The SR240 is ideal for amplifying low-level signals from photomultipliers and photodiodes before being measured by the SR250.

SR245 Computer Interface Module

The SR245 adds both analog and digital data acquisition capabilities to System SR250. Eight analog I/O channels can be configured as inputs or outputs, two front panel digital I/O bits are provided, and an internal 8 bit digital I/O connector is available. The SR245 can communicate with your computer over the RS-232 and GPIB interfaces.

SR255 Fast Sampler Module

The SR255 fast sampler allows you to do gated integration with gate widths as short as 100 ps. Four discrete gate widths are provided: 100 ps, 200 ps, 500 ps, and 1000 ps. Output is provided in both analog and digital form, and special correction circuitry is provided to eliminate nonlinearities present in the sampling bridge.

SR272 Data Acquisition Software

The SR272 is a Windows 95® compatible software package designed to acquire, display, and analyze data taken with the System SR250. The program works with the SR245 Computer Interface module over the GPIB or RS-232 interface. Features include on-screen cursors for easy data read-out, autoscaling, and command logging. Command strings can be sent to other instruments via RS-232 or GPIB, allowing synchronization and control of other instruments in your experiment.

Ordering Information

SR200	**Gate Scanner**	**$ 1000**
SR235	**Analog Processor**	**$ 1500**
SR240	**300 MHz Preamplifier**	**$ 1000**
SR245	**Computer Interface Module**	**$ 1500**
SR250	**Gated Integrator Module**	**$ 2990**
SR255	**Fast Sampler**	**$ 2990**
SR272	**Data Acquisition Software**	**$ 500**
SR275	**Display Module**	**$ 800**
SR280	**Mainframe and Display Module**	**$ 1600**

Gated Integrators and Boxcar Averagers

Model SR250 — Single Channel Gated Integrator

- ***Gate Widths from 2 ns to 15 µs (expandable to 150 µsec)***
- ***Internal Rate Generator***
- ***Active Baseline Subtraction***
- ***Shot by Shot Output***
- ***Gate Output Allows Precise Gate Timing***
- ***Average 1 to 10,000 Samples***
- ***DC to 20 kHz Repetition Rate***
- ***Jitter Less Than 20 ps + 0.01% of Delay***
- ***Low drift <.5mV/hr***
- ***Trigger to Gate Delay <25 ns***
- ***Inputs Protected to 100 VDC***
- ***Standard Double Width NIM Module***

SR250 Overview

The SR250 Gated Integrator is a versatile, high speed, low cost, NIM module designed to recover fast analog signals from noisy backgrounds.

The SR250 consists of a gate generator, a fast gated integrator, and exponential averaging circuitry. The gate generator, triggered internally or externally, provides an adjustable delay from a few nanoseconds to 100 milliseconds before it generates a continuously adjustable gate with a width between 2 ns to 15 µs. The gate delay can be set from the front panel, or can be automatically scanned by applying a rear panel control voltage. Scanning the gate allows the recovery of entire waveforms.

The fast gated integrator integrates the input signal during the gate. The output from the integrator is then normalized by the gate width to provide a voltage proportional to the average of the input signal during the sampling gate. This signal is further amplified and sampled by a low droop sample and hold amplifier, and output via a front panel BNC connector. The last sample output allows for a shot-by-shot analysis of the signal, and makes the instrument a particularly useful component in a computer data acquisition system.

SR250 Features

Trigger

The SR250 may be triggered internally or externally. The internal rate generator is continuously variable from 0.5 Hz to 20 kHz in 9 ranges. The external trigger pulse may be as short as 5 ns, allowing the unit to be triggered with fast pulses from photodiodes and photomultipliers. Single shot and line triggering can also be selected.

Signal Inputs

The sensitivity of the instrument (volts out / volts in) may be set from 1V/1V to 1V/5mV. If additional gain is required, the SR250 can be used with the SR240 preamplifier. The input is protected to 100 Volts and has a 1 MΩ input impedance. An input filter rejects unwanted signals before the input is sampled by the integrator. Unwanted DC input offsets are easily nulled with a ten turn potentiometer.

Gate Timing

The delay of the sample gate from the trigger is set by the delay multiplier and scale. The delay scale is multiplied by the setting on the 10-turn multiplier dial, allowing continuously adjustable delays from a few nanoseconds to 100 milliseconds. The delay multiplier may also be changed from the rear panel control voltage input, a useful feature in applications requiring a scanning gate. Zero to ten volts at this input overrides the front panel 0 to 10x delay multiplier. Insertion delay from trigger to gate is only 25 ns, and gate delay jitter is only 20 ps + 0.01% of the full scale delay.

The width of the sampling gate may be continuously adjusted from 2 ns to 15 µs over 8 width ranges. A simple modification of the unit allows gate widths of up to 150 µs. The front panel gate output provides a representation of the gate that can be overlayed with the signal on an oscilloscope to provide a precise display of the gate timing.

Signal Outputs

A moving exponential average of 1 to 10,000 samples can be selected from the front panel. This traditional averaging technique is useful for pulling small signals from noisy backgrounds. In the case of a random white noise background, the signal-to-noise ratio increases as the square root of the number of samples in the average. This allows a S/N improvement of up to a factor of 100 using this technique alone. If no averaging is desired, or if averaging is to be performed on an external computer, the last sample output provides a voltage proportional to the average value of the input signal during the last gate period.

Average Reset

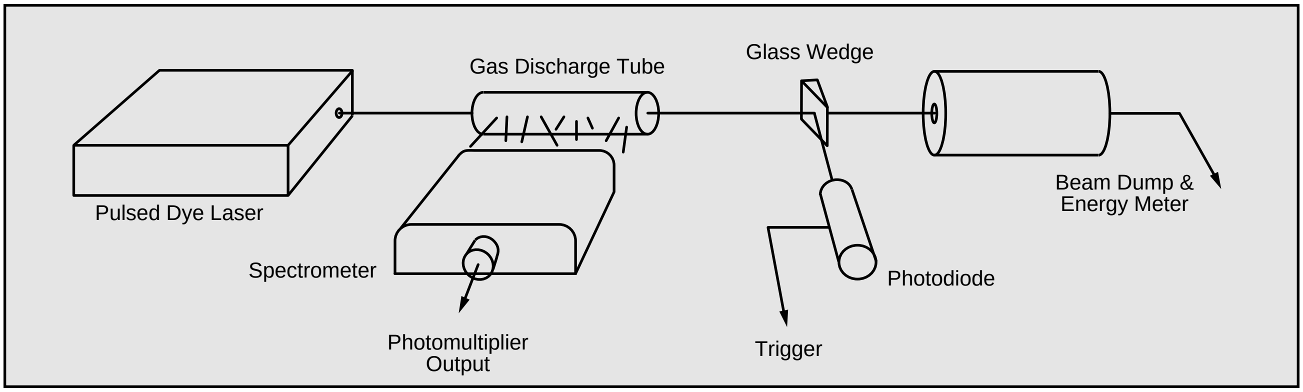
The reset button sets the average output to zero. The average may also be reset by a rear panel logic input. The average reset input will accept a TTL low level or a switch closure to ground to reset the moving average output. The input must be held low for at least 1 µs to reset the moving average.

Polarity Control and Active Baseline Subtraction

The polarity of the last sample and averaged outputs is controlled by rear panel toggle switches. Positive outputs can be selected for negative signals, and vice versa, allowing easy interfacing with unipolar analog to digital conversion systems. In addition to the traditional averaging modes, the SR250 possesses a unique Active Baseline Subtraction mode which allows you to actively cancel baseline drift. In the Active Baseline Subtraction mode the SR250 is triggered at twice the source repetition rate. On alternate triggers, when the signal is not present only the baseline is sampled and the SR250 inverts the polarity of the last sample output before it is added to the moving average. Thus, any baseline drift not associated with the source will be subtracted out.

Additional Outputs

The signal input is passed on to the signal output by a length of coaxial cable for termination and for gate timing. It is delayed exactly 3.5 ns from the input and can be terminated to optimize either signal gain or response time. The gate output provides a pulse synchronized with the internal gate signal. The gate output is timed so that it can be overlayed with the signal output for precise adjustment of gate timing. The busy output provides a TTL timing pulse which is high while the unit is integrating, and goes low when the SR250 is ready to accept another trigger. These outputs help simplify experimental setup and troubleshooting.



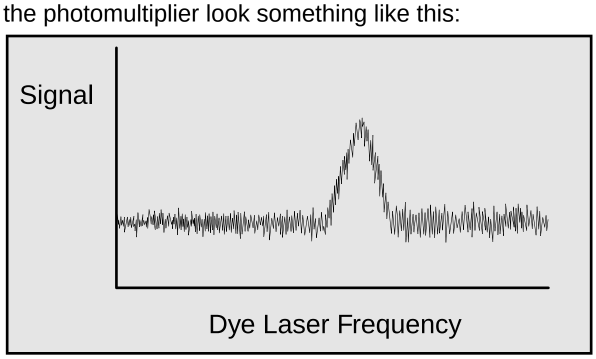
Sample Application

This section describes a specific application of the SR250 Gated Integrator working together with other System SR250 modules. Although the specific examples involve data acquisition in a pulsed laser experiment, many of the concepts and techniques are applicable to other areas.

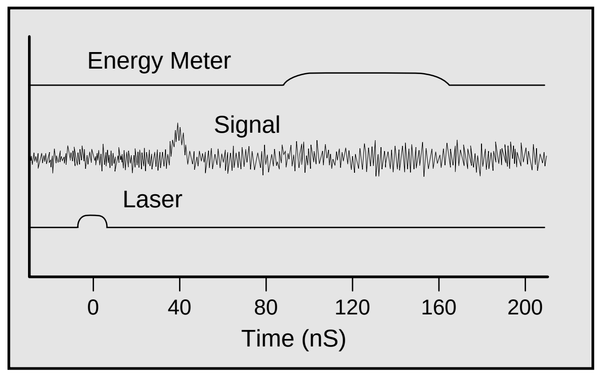
Pulsed Laser Data Acquisition

In this example, excited state atoms in a gas discharge are to be transferred to a different excited state by a pulsed dye laser. The experimental arrangement is shown above. A pulsed dye laser is injected into a gas discharge tube which contains atoms in the initial excited state. A portion of the laser beam is split off and detected by a photodiode in order to trigger the gated integrator. The remainder of the beam is dumped to an energy meter which outputs a signal proportional to the energy in the laser pulse.

The transfer of population will be inferred by the increased fluorescence from the final state as the dye laser frequency is tuned through the transition frequency. The experiment is difficult because the background fluorescence signal from final state atoms excited by the discharge is larger than the laser induced signal. A monochromater with a photomultiplier detects the signal at the wavelength of the final state. As the laser wavelength is tuned, we'd like to see the signal from the photomultiplier look something like this:

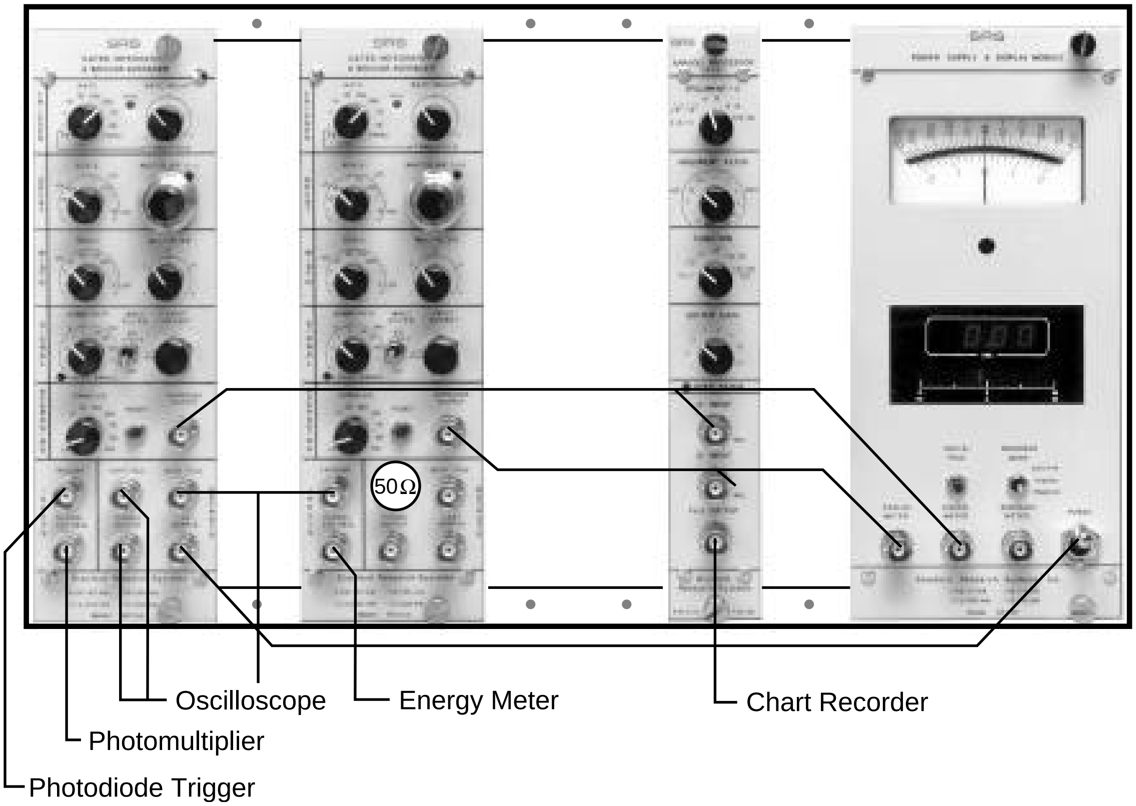


Some other aspects of the experiment can be inferred from the timing diagram below:



1) The laser pulse width is about 8 ns.
2) There is a 40ns delay in the photomultiplier, from the time photons strike the photo-cathode until the anode signal is seen.
3) The laser induced signal is expected to be only 25% of the background and is only present during the laser pulse. The experimenter wants a final Signal/Noise ratio of 5:1.

The boxcar system components needed for this measurement are shown on the following page. One SR250 is used to look at the output of the photomultiplier tube. To maximize the signal to noise ratio, we'd obviously like this integrator to be gated only during the time the signal is present. If we make the gate too wide, we'll be simply integrating more background without any more signal. That's why it's so important to use a gated integrator with low gate jitter such as the SR250. To ensure proper gate timing two equal lengths of cable can be brought from the signal output and the gate output to a dual channel oscilloscope. By overlaying the traces, or by using the ADD mode of the scope, the gate can be accurately placed on top of the signal. This first SR250 will be triggered by a photodiode looking at a small reflection of the laser pulse. By triggering with the laser pulse itself, any jitter between the laser and the laser's own trigger can be ignored.



Since the signal is only 25% of the background, and the desired final signal to noise ratio is 5:1, a 20 fold improvement in signal-to-noise is required. This can be accomplished by averaging over 20^2, or 400, shots. If the laser was firing at 10 Hz, this means that you'd have to wait at least 40 seconds at each point to get a valid measurement–if you scanned through the transition at a faster rate the data would be distorted.

A second SR250 in this experiment is integrating the signal from the energy meter. It is triggered by the busy output of the first boxcar, and will use a much longer gate. The averaging for the second SR250 should be the same as for the first.

We now have two boxcar outputs: one proportional to the signal intensity and one proportional to the laser intensity. Since we expect the signal to be proportional to the laser intensity, the effects of fluctuation and drifts in laser intensity can be removed by normalizing the signal to the laser intensity. This is done in our setup using the SR235 Analog Processor module. The averaged output of the signal SR250 is fed to the "A" input of the SR235. The averaged output of the laser intensity SR250 is sent to the B. The SR235 is set to output the value of 10A/B, or ln(A/B) if additional dynamic range is required. The output of the SR235 is displayed on a stripchart recorder.

Notes on Computer Interfacing

Several improvements in the operation of this experiment can be achieved by controlling the data acquisition with a laboratory computer. The unique ability of the SR250 Gated Integrator Module to provide shot-by-shot data dramatically enhances the instrument's utility in a computer controlled environment. Since the computer can acquire data on each laser shot, you are no longer limited to the exponential averaging provided by the SR250. For instance, the computer could be programmed to take 10 points at each wavelength and linearly average the results. Of course, we discussed earlier that averaging over 400 points was necessary to achieve the desired signal-to-noise improvement. In a computer controlled experiment we could take 10 points at each wavelength, and then average 40 entire scans together. In this manner, the effects of drift in laser intensity would be balanced over each scan instead of concentrated at the end of a single long scan.

The SR245 Computer Interface Module is designed to work with the SR250 to bring you the advantages of computer control. The SR245 can be triggered by the SR250's "BUSY" output and acquire up to 8 channels of analog data. The results can be sent back to your computer via RS-232 or GPIB interfaces.

Overview

The SR250 is a NIM format Gated Integrator/Boxcar Averager that is a self contained gate generator and fast gated integrator with averaging circuitry. Designed for recovering fast analog signals from noisy backgrounds, it is particularly applicable for pulsed laser experiments.

Trigger

Internal Trigger	The rate generator is continuously adjustable from 0.5 Hz to 20 kHz in 9 ranges.
Line Trigger	The gate generator may be triggered from the power mains with adjustable phase.
External Trigger	The gate generator may be triggered by the EXTERNAL TRIGGER INPUT. 1 MΩ input impedance. Trigger threshold adjustable from 0.5 to 2V. The input is protected to ±100 Vdc. The trigger pulse must be over threshold for at least 5 ns and have a rise time shorter than 1 µS.
Manual Trigger	The unit will trigger once if the trigger threshold is scanned through 0 Vdc.
Trigger LED	The Trigger LED blinks with each trigger.

Delay

Delay Scale	Delay scales from 1 ns to 10 ms may be selected.
Delay Multiplier	A 10-turn dial is used to select a delay multiplier from 0 to 10.
Insertion delay	25 ns
Accuracy	2 ns or 5% of the full scale delay, whichever is larger.
Jitter	< 20 ps or 0.01% of the full scale delay, whichever is larger.
Ext. Delay Control	Rear panel input voltage from 0 to 10 VDC overrides the front panel delay multiplier. This input is used by the SR200 Gate Scanner Module or SR245 Computer Interface Module to scan the sample gate in order to record entire waveforms.

Width

Width Scale	1, 3, 10, 30, 100, 300 ns, 1, 3 µs.
Width Multiplier	Continuously adjustable from x1 to x5.
Width Accuracy	2 ns or 20% of full scale, whichever is greater.
Minimum Width	2 ns, FWHM.

Signal

Sensitivity	The sensitivity, volts out/volts in, may be set from 1V/1V to 1V/5mV in a 1, 2, 5 sequence. The sensitivity calibration is accurate to 3% for gate widths longer than 10 ns. The sensitivity decreases to about 50% of nominal at a gate width of 2 ns.
Filter	An input filter allows coupling of DC, AC above 10 Hz, or AC above 10 KHz.
Offset Control	A 10-turn potentiometer adds an input offset of ±0.4 VDC. When using very narrow gates, the offset may need adjustment if the gate width is changed.
Over Range LED	A red LED indicates if the signal is greater than 2 Vdc or if the LAST SAMPLE output is greater than 10 Vdc

Last Sample

Output	A ±10VDC full scale output with an output impedance <1Ω and a 10 mA drive capability. The short circuit output current is limited to about 20 mA. The droop rate at this output is < 0.2% of full scale per second.
Polarity	The polarity of the LAST SAMPLE output may be inverted by a rear panel switch.
Responsivity	95% (no more than 5% of the previous last sample remains)

Averaging

Samples	An exponential moving average may be taken over 1, 3, 10, 30, ... to 10,000 samples or LAST may be selected for no averaging.
Reset	Front panel push button resets the moving average to zero.
Remote Reset	Rear panel BNC input resets the moving average with TTL low or switch closure.
Average Output	BNC output, ±10 VDC full scale, with an output impedance <1Ω, and a 10 mA drive capability. Short circuit current limits at about 20 mA. When there are no triggers the droop rate is <1% per minute on 1 to 30 samples, and < 0.01% per minute for 100 to 10K samples in the average.
Average Polarity and Baseline Subtraction	A rear panel switch selects the polarity of the LAST SAMPLE before it is added to the moving average. This may be used to invert the polarity of the AVERAGE OUTPUT. The switch can also be set to the

TOGGLE position, in order to subtract every other sample from the moving average. By triggering the unit at twice the experiment's repetition rate, the baseline will be sampled on alternate triggers and the baseline will be subtracted from the moving average.

Toggle Output	A rear panel BNC provides a TTL signal that changes state with each trigger. This output is used with the Active Baseline Subtraction feature to indicate if the next sample will be added to, or subtracted from, the moving average. The toggle output is capable of driving 50Ω lines to +2 VDC.

Signal Input and Output

Signal Input	1MΩ input impedance, ±2 VDC usable range, protected to 100 VDC. The input offset drift is <0.5 mV per hour after a 20 minute warm-up. Shot noise at the input is < 0.5 mV. Coherent pickup is < 5 mV, which may be cancelled with the input offset control in fixed gate applications.
Signal Output	The SIGNAL OUTPUT is the input signal delayed by 3.5 ns. This output is used to terminate the input signal, and to precisely time the sample gate with respect to the signal output.

Gate and Busy Outputs

Gate Output	A 200 mV output pulse marks the exact position of the sample gate with respect to the signal output. Position accuracy is ±1 ns. This output must be terminated into a 50Ω load.
Busy Output	This TTL output is used to synchronize the experiment to the internal rate generator (if it is used) or to signal the data acquisition computer that the unit has been triggered and data is ready. The output will drive 50Ω lines to 2 VDC, and is high from the time of the trigger until the unit is ready to accept another trigger input. (A minimum of 45 µs, longer for long delays or gate widths.)

General Specifications

Power Supplies	+24V/135 mA, +12V/380 mA, -12V/230 mA, -24V/150 mA. Approximately 14 Watts. Power from a NIM standard crate or from Stanford Research Systems mainframe model SR280.
Mechanical	Dual width NIM standard enclosure per TID-20893.
Dimensions	2.7"W x 8.714"T x 11.5"D
Warranty	1 year parts and labor on defects in materials or workmanship

Ordering Information

SR250	**Gated Integrator Module**	**$ 2990**

Gated Integrators and Boxcar Averagers

Model SR255 — 100 ps Fast Sampler Module

- ***Gate Widths of 100 ps, 200 ps, 500 ps and 1000 ps***
- ***Droopless Analog Output***
- ***Linear Over Full Range***
- ***Jitter <2 ps rms***
- ***<20 ns Insertion Delay***
- ***Trigger Rates up to 50 kHz***

SR255 Overview

The SR255 Fast Sampler Module is a fast gated integrator providing four discrete user-selected gate widths from 100 ps to 1 ns. All of the necessary electronics are built into this high speed module including an A/D, D/A, and a PROM correction circuit to eliminate the inherent non-linearities in the sampling bridge. Output is provided in both digital and analog form. Input sensitivity can be set from 100 mV/V to 1 V/V, and trigger level selection is provided to ensure compatibility with a broad range of trigger sources. The SR255 is perfect for fast pulsed experiments where the 2 ns minimum gate width of the SR250 is insufficient. The gate delay is controlled by a rear panel voltage input which can be supplied by the SR200 Gate Scanner or the SR245 Computer Interface. Convenient gate view and signal outputs allow precise positioning of the gate– particularly important in applications such as time domain reflectometry or shorted-cable baseline subtraction. The SR255 can be used alone, combined with the SR245 Computer Interface Module for automated data acquisition, or operated with the SR200 Gate Scanner for scanning gate waveform recovery.

SR255 Features

Triggering

The SR255 is triggered by a signal at the 50Ω, DC coupled trigger input. The trigger threshold can be set to -0.5 V, +0.1 V, or +1 V on the front panel. For reliable triggering, the trigger must remain over threshold for at least 5 ns, and not exceed 5 volts. The maximum trigger rate is 50 kHz (reduced when operating on the 1 µs delay scale).

Gate Delay

The delay from trigger to sample is controlled by an analog voltage applied at the rear of the unit. You can select delay ranges of 1, 10, 100, or 1000 nS/V with four switches accessed through the instrument's side panel. In addition to the adjustable delay, there is a fixed insertion delay of about 20 ns. With only 2 ps rms gate jitter, the SR255 makes it easy to set up and maintain precise gate timing.

Gate Width

Four fixed gate widths of 100 ps, 200 ps, 500 ps and 1000 ps can be easily selected. A front panel LED indicates which gate width is being used whenever power is applied.

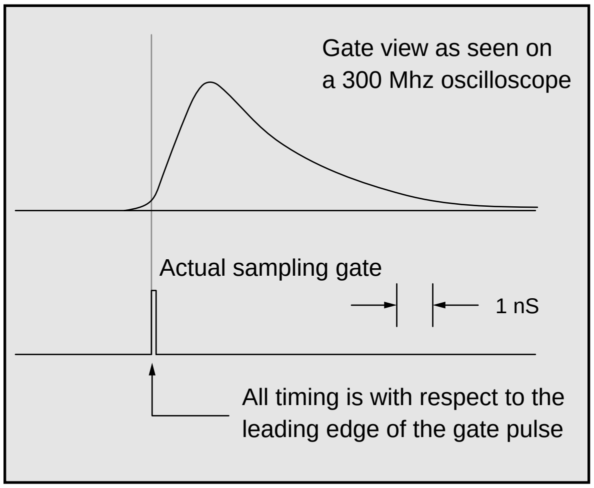
Signal Input

The SR255 is designed to be used with RG58A cable. To achieve an optimally flat response, two meters of RG58A cable should be used with a a BNC to N-Type converter (part #UG 201A/U is provided with each SR255). The frequency response of the SR255's front end has been peaked above 2 GHz to compensate for the losses in 2 meters of cable, and so this length of cable is optimum for the signal input line. This input is passed to the signal out BNC via an internal 300 ps delay line.

The signal output should be terminated in 50Ω with a high quality terminator to minimize reflections and pulse distortion. The signal output aids in synchronizing the sample with the gate and can be used for special applications such as time domain reflectometry or shorted-cable baseline subtraction. The input signal should not exceed the sensitivity selection on the front panel, as the input is only protected to +5 Vdc.

Gate View and Fast Timing

You can use the gate view output to time the sample gate with respect to the signal. The leading edge of the gate view output (50 ps rise time) indicates when the sample gate is being opened, while the output is a pulse of approximately 3 Volts with an exponential decay of about 4 ns.



The gain of the module, volts in/volts out, can be set to 1 V/V, 0.25 V/V, or 0.1 V/V. For example, when 0.1 V/V sensitivity is selected, a 100 mV input will produce a 1 V output. The red overload LED will come on when the output exceeds approximately 1 volt.

Both Analog and Digital Outputs

The SR255 has both analog and digital outputs making it easy to interface in a wide range of experiments. The sample output provides a ±1VDC full scale analog voltage proportional to the value of the signal during the gate interval. Resolution is 0.5% of full scale, and the output can drive up to 10 mA. This output is available 20 µs after the signal is sampled. Each unit is custom linearized with a PROM to ensure excellent linearity and full dynamic range.

Simple to Interface

The Fast Sampler also provides a digital interface through a 15 pin connector on the rear panel of the module. This interface is a parallel binary interface which may be connected to either the 20 pin connector on the circuit board of the SR245 computer interface module or to any 8 bit digital I/O port.

The SR255's 15 pin connector has six digital inputs, eight digital outputs, and a common ground line. The incoming lines serve as module address lines and the out-going lines are tri-state outputs allowing several modules to be daisy-chained together over the same digital bus. Each Fast Sampler in a system can be assigned an address by setting four address switches which are accessed through the right side panel of the module.

Using the SR255 as a Sampling Scope

The SR255 Fast Sampler can be used in conjunction with the SR200 Gate Scanner and an X-Y oscilloscope to provide sampling oscilloscope operation. In a sampling oscilloscope, a narrow sampling gate is scanned over a repetitive waveform in order to recover the shape of the waveform. A sampling “scope” made from the SR255 and SR200 can achieve a resolution of 100 ps and sample rates of up to 50,000 samples per second.

In this application, the SR200 Gate Scanner is used to scan the gate delay over 0 to 10 times the selected delay scale. This scale is set on the SR255 to 1 ns, 10 ns, 100 ns, or 1 µs to provide time-bases with delays of up to 10 µs. (The maximum delay can be extended by the user by increasing capacitor C115 of the 1 µs delay scale). The X-AXIS OUT of the SR200 is used to drive the horizontal (X-axis) of the oscilloscope. The PEN LIFT OUT may be connected to the oscilloscope to blank the CRT during retrace.

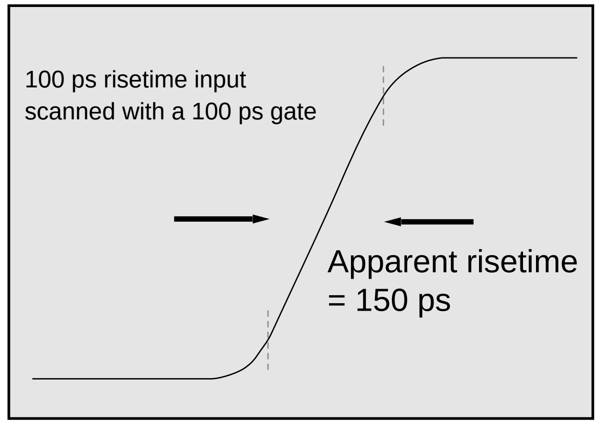
Finally, the SAMPLE OUT of the SR255 Fast Sampler is connected to the vertical (Y-axis) of the oscilloscope to display the sampled signal. The resolution can be changed by changing the Gate Width of the SR255 (100 ps, 200 ps, 500 ps, 1 ns.)

Bandwidth

What happens if we use our “sampling scope” to look at a fast edge? The apparent rise time of a infinitely fast step is roughly equal to the selected gate width. The -3 dB bandwidth of the unit is approximately equal to 0.35 divided by the selected gate width. The following table gives the available gate widths and their corresponding bandwidths.

Gate Width	Bandwidth
1000 ps	350 MHz
500 ps	700 MHz
200 ps	1.7 GHz

Now lets look at what happens when we use our sampling scope to measure the risetime of a real signal. The next figure shows the measurement of a known 100 ps risetime edge using an SR255 “scope” with a 100 ps gate. We’d expect that the resultant pulse shape would be the convolution of the input signal with the SR255's gate. Since widths add roughly in quadrature under convolution, we’d expect the observed risetime to be approximately $\sqrt{(100^2+100^2)}$ or 140 ps. In fact, the actual observed risetime is 150 ps, very close to our prediction.



Noise

From the previous discussion you might think that it’s always best to run with narrow gates to get the best possible bandwidth and step response. This is not necessarily the case. The penalty paid for using narrow gate widths is increased noise. Narrow gates have more noise because of their reduced sampling efficiency and wider bandwidth. Typical noise characteristics are given below.

Gate Width	Peak to Peak Noise	RMS Noise
1000 ps	1.0 mV	200 µV
500 ps	1.8 mV	350 µV
200 ps	3.0 mV	600 µV
100 ps	4.0 mV	800 µV

Overview

The SR255 Fast Sampler Module is a fast gated integrator with four fixed gate widths between 100 ps and 1000 ps. With both analog and digital outputs, less than 2 ps of rms gate jitter, and a scannable gate delay, the SR255 is ideally suited for a variety of short pulse experiments.

Gate Generator

Trigger Input	BNC, 50Ω termination Thresholds: -0.5, +0.1 or +1 Vdc Trigger Rate: dc to 50 kHz
Gate Delay	20 ns + (Delay Range) X (Control Voltage). Delay Range : 1, 10, 100 or 1000 ns/Volt. Control voltage input is on the rear.
Gate Jitter	<2 ps RMS
Gate Width	100, 200, 500, 1,000 ps
Gate View	N-type connector. Leading edge indicates when gate opens.
Gate View Accuracy	±50 ps with respect to signal out.
Gate View Risetime	< 50 ps.

Signal Channel

Signal Input	N-type connector. Characteristic impedance is 50 Ohms. Protected to 5 Vdc. The full scale input level equals the sensitivity setting.
Shot Noise (typical)	200 µV rms on 1000 ps 350 µV rms on 500 ps gate 600 µV rms on 200 ps gate 800 µV rms on 100 ps gate
Sensititviy	0.1, 0.25, or 1.0 Volts full scale. Overange LED indicates signal greater than full scale.
Signal Out	The Signal In is passed to the Signal Out for termination, gate timing and for special applications such as time domain reflectometry.

Outputs

Sample Out	±1 V full scale analog output. Linearized and latched representation of the signal input as sampled during the gate. Resolution is 1/2% of full scale. Output impedance <1 Ohm. 10 mA drive capacity.
Digital Out	Rear panel 8-bit digital interface is addressed as two bytes, an 8-bit data byte for amplitude, an 8-bit status byte for sign, gate width sensitivity, data ready rate error, and overrun status.

General Specifications

Power	+24 V/100 mA, -24 V/120 mA, +12 V/180mA, approximately 8 Watts total.
Connectors	Three N-type to BNC transitions are provided.
Mechanical	Single width standard NIM module
Warranty	One year parts and labor on any defect in materials or workmanship

Ordering Information

SR255	**Fast Sampler Module**	**$ 2990**

Model SR200 — Gate Scanner Module

- ***Gate Scanner for SR250 and SR255***
- ***Forward / Reverse Scans***
- ***Repeat / Single Shot***
- ***Pen Lift Output***

SR200 Overview

The SR200 Gate Scanner Module is designed to automate waveform recovery with the SR250 and SR255 gated integrator modules. Waveform recovery with gated integrators is done by slowly scanning the gate of the integrator over the waveform. Both the SR250 and SR255 modules have external gate delay control inputs. The SR200 provides the adjustable ramp voltage needed to scan the gates using these inputs. The initial and final delays, as well as the scan time, are fully adjustable. Single or repeated scans may be performed in the forward or reverse direction over any portion of the waveform. Scan times from 10 milliseconds to 5 minutes may be selected.

In addition to the delay control output, the SR200 has a 0-10 Volt X-axis ramp output designed to drive the X-axis of a chart recorder or an oscilloscope. A convenient Pen Lift output lifts the chart recorder pen at the end of the scan, or can be used to blank a scope display during the horizontal retrace.

Ordering Information

SR200	**Gate Scanner**	**$ 1000**

SR200 Specifications

Controls

Reverse/Stop/Forward	Selects scan direction or stops the scan.
Single/Reset/Repeat	Selects single or repeated scans or resets to start.
Scan Time	Selects time to complete one scan.
Start Position	Sets the smallest delay in the scan.
Scan Width	Sets the range of delays in the scan.

Outputs

Control Voltage	Rear panel output connects to the SR250 or SR255 Delay Multiplier Input. Impedance <1Ω, 20 mA current limit.
Pen Lift	Logic Signal to lift chart recorder pen or blank an oscilloscope trace.
X-Axis	Analog voltage output scans between 0 and 10.0 Vdc regardless of the start position and scan width settings.

General

Power Supplies	+24V/20mA, +12V/80mA, -12V/0mA, –24V/20mA. 2 W total.
Mechanical	Single Width NIM standard module
Warranty	One year parts and labor on any defect in materials or workmanship.

Gated Integrators and Boxcar Averagers

Model SR235 — Analog Processor Module

- ***Dual Inputs***
- ***Six Complex Output Functions***
- ***Filtering and Gain on Output***

SR235 Overview

The SR235 Analog Processor provides a variety of convenient signal processing functions on up to two inputs. Background subtraction, ratioing, and logarithmic compression are just a few of the functions which can be simply implemented with the SR235. With its many output functions, high accuracy, and variable filtering and gain, the SR235 is the perfect addition for any boxcar system, especially those in which a computer is unavailable to perform signal processing.

The SR235 outputs a voltage proportional to a function of an argument formed from its two inputs (A and B). Allowable arguments are: A, B, $\sqrt{(A^2+B^2)}$, A-B, AxB/10, and 10A/|B|. The functions that can be selected are: x, x^2, $\sqrt{x}$, ln|x|, -dx/dt, and -(dx/dt)/100. Filtering can be performed on the argument with time constants from 0.3 ms to 30 s. The final function output can be amplified with gains of 0.1 to 20 in a 1-2-5 sequence within the SR235's linear output range of ±10 V.

Ordering Information		
SR235	**Analog Processor**	**$ 1500**

SR235 Specifications

A and B Inputs	1 MΩ input impedance. Operating range ±10 Volts, protected to 100 Volts.
Input Offset	< 2 mV.
Argument (X)	A, B, $\sqrt{(A^2+B^2)}$, A-B, AxB/10, and 10A/|B|.
Argument Filter	Select time constants from 0.3 ms to 30 s in a 1, 3, 10 sequence. When the selector is 'OFF' the argument is unfiltered.
Gain	Select from 0.1 to 20 in a 1, 2, 5 sequence
F(x) Output	x, x^2, $\sqrt{x}$, ln|x|, -dx/dt, and -(dx/dt)/100
Frequency Range	-dx/dt to 10 Hz, (-dx/dt)/100 to 1 kHz, $\sqrt{(A^2+B^2)}$ to 20 kHz, and all others from DC to 50 kHz.
Accuracy	Gain, 2%; rms sum, 3%; difference, 1%; multiplication, 2% of full scale; division (with denominator >0.1), 3% of full scale; ln|x|, x^2, $\sqrt{x}$ are accurate to ± 20 mV referenced to the input or at the output, whichever is less; -dx/dt and (-dx/dt)/100 5%.
Power	+24V/120mA, -24V /80mA. 5 W total.
Mechanical	Single width NIM Standard
Warranty	One year parts and labor on any defects in material or workmanship

Gated Integrators and Boxcar Averagers

Model SR245 — Computer Interface Module

- ***Eight Analog Input/Output Ports***
- ***8-Bit Digital Input/Output Port***
- ***Two TTL I/O Ports***
- ***RS-232 and GPIB Interfaces***
- ***3500 Point Sample Memory***
- ***Simple Command Structure***

Ordering Information

SR245	**Computer Interface**	**$ 1500**

SR245 Overview

The SR245 Computer Interface module is a powerful tool for data acquisition, providing both an analog and a digital interface between the computer and your experiment. The eight analog I/O channels can be designated through software as all inputs, all outputs, or as a combination of inputs and outputs. Both inputs and outputs have 13 bits of resolution over the ±10.24 VDC full scale range, with 0.05% accuracy.

Two front panel digital I/O bits are provided for use as counters or triggers, or can be set or read by the computer. Additionally, an 8-bit input and an 8-bit output port are available (on an internal connector) for your own custom digital interfaces.

Both RS-232 and GPIB interfaces are included. Simple commands make the SR245 easy to program from a variety of high level languages–all that's necessary is the ability to send and receive ASCII strings. For example, sending ?5 instructs the module to digitize and send the voltage on the 5th analog input BNC. Other commands allow you to record in the module's 3500 point buffer memory, ramp an analog output at a specified rate (for gate scanning), or read the contents of a digital counter.

Analog Ports

Configuration	Any number of the eight ports may be designated under program control as input ports, the rest default to output ports.
Inputs	1 MΩ input impedance, ±10.24 Vdc range, protected to 40 Vdc. 13 bit resolution (2.5mV). Accuracy: 0.05%. Input offset less than 2.5mV. Maximum A/D rate is 2 kHz.
Outputs	Output impedance <1Ω. Short circuit current limit is 20 mA. 13 bit resolution (2.5mV). Accuracy: 0.05%. Output offset <2.5 mV.

Digital Ports

Type	Two Front Panel I/O TTL bits, One 8-bit digital input port, One 8-bit latched digital output port
Front Panel-Inputs	Input impedances > 100 kΩ. Minimum pulse width is 200 ns. Maximum count rate is 4 MHz. Logic > 3 Vdc, Logic zero <0.7 Vdc. Inputs protected to ±10 Vdc.
Front Panel-Outputs	Can drive loads up to 50 Ω to TTL levels

General

Interfaces	Both IEEE-488 and RS-232 (110 to 19.2 kbaud).
Power	+24V/60mA, 24V/60mA, +12V/20 mA. Approximately 8 Watts total
Mechanical	Single width standard NIM module
Warranty	One year parts and labor on defects in materials and workmanship

Command List

Input/Output Commands

I<n> n=0-8	Designates the first n analog ports as inputs, the remainder become outputs.
?<n> n=1-8	Returns the value of the designated analog port.
?B<n> n=1,2	Returns the value (0 or 1) of the designated digital bit port.
?D	Returns the value of the internal 8 bit digital input port.
?S	Returns the value of the status byte, and clears the status byte.
C	Configures B2 as an input and resets the B2 counter.
?C	Returns the number of pulses occurring at B2 since the previous ?C.
S<n>=<x> n=1-8 x=-10.237 to +10.237	Sets the analog port n (which must be designated as an output) to the value x Volts
SB<n>=<m> n=1,2; m=0,1	Designates digital bit n as output and sets its value to m.
SB<n>=I n=1,2	Designates the selected bit as an input.
SD=<n> n=0-255	Sets the 8-bit digital output port to the value n.
SM=<n> n=0-255	Sets the GPIB SRQ mask to the value n

Trigger Commands

MS	Sets the synchronous mode. Responses to ? commands are returned immediately after the next trigger.
MA	Sets the asynchronous mode (default). Responses to ? commands are returned right after command is received.
T<n> n=1 to 32,767	Designates every nth pulse at B1 as a trigger.
DT	Masks the trigger input so that no triggers are recognized.
ET	Unmasks the trigger input.
PB<n> n=1,2	Outputs a 10µsec TTL pulse at bit port n.
P/<n> n=1-255	Outputs a 10µsec TTL pulse at B2 each nth trigger.

Scan Commands

SC<i>,<k>:<n> i..k=1-8,D	Scans the list i..k of analog ports or digital port for n triggers. Total # of samples may not exceed 3711.
ES	End the current scan immediately and reset the point sending counter.
N	Send the next point of stored scan.
?N	Returns # of points scanned.
A<n>,<i> n,i=1-255	Adds n x 2.5 mV to the value of analog port 8 (must be a positive output) on every ith trigger.
SS<i>,<k>:<n> i..k=1-8,D	Scans the list i..k of analog ports or digital port for n triggers. Data is sent in a 2 byte binary format while the scan is in progress.
X	Sends the data of a stored scan in 2 byte binary format.

Miscellaneous Commands

MR	Master Reset. Returns the SR245 command settings to their default values.
W<n> n=0-255	Introduces a delay of (n x 400 µsec) before sending each character over the RS232.
Z<i>,<k>	Changes the end-of-record characters sent by the SR245 to those specified by the ASCII codes, i...k.

Model SR240 — 300 MHz Preamplifier

- ***DC to 300 MHz Bandwidth***
- ***1.2ns Rise/Fall Time***
- ***2.2nV/√Hz Input Noise***
- ***Four Independent Channels***
- ***Voltage Gains to 125***

SR240 Overview

The model SR240 Fast Preamplifier contains four wide bandwidth, dc coupled amplifiers, each with a gain of 5. The amplifiers can be used independently or cascaded to provide gains of 5, 25, or 125. The fast rise time, low noise, and DC accuracy of the SR240 make it the ideal instrument for use with fast photomultiplier tubes and photodiodes.

The SR240 preamp is useful for amplifying small signals to levels that can be processed by other boxcar system modules. Typically, a signal of at least a few millivolts is required at the input of the SR250. If your detector does not supply this signal level, the SR240 can be used in front of the SR250 to ensure sufficient signal amplitude.

In order to maintain linear operation, input levels should be kept below 200 mV for single channel (gain of 5) operation, 40 mV for dual channel operation (gain of 25), and 8 mV for three-channel operation (gain of 125).

Ordering Information

SR240	**300 MHz Preamplifier**	**$ 1000**

SR240 Specifications

Amplifier

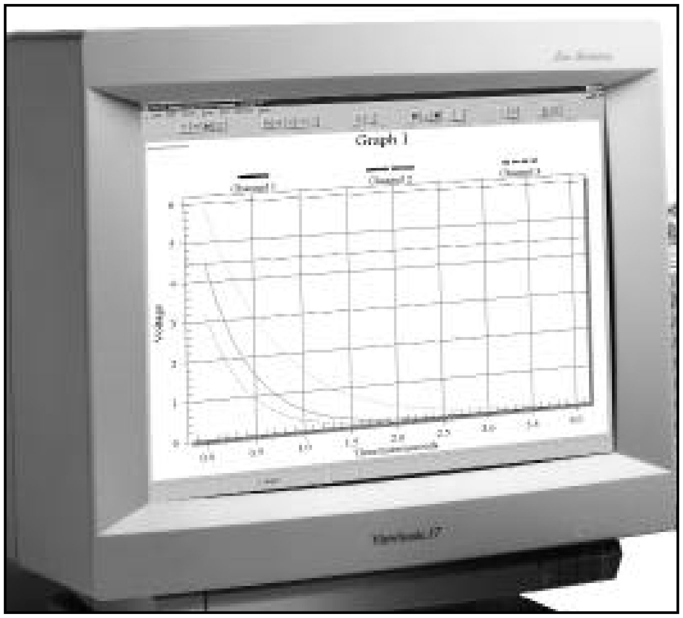
Input	50Ω impedance, dc coupled, BNC connectors.
Outputs	dc coupled, BNC connectors (50Ω termination).
Operating Range	Inputs: ± 200 mV, Outputs: ±1.0 V.
Voltage Gain	5 per channel. Up to 3 channels can be cascaded.
Bandwidth	dc to 300 MHz (-3dB)
Rise/Fall Time	1.2 ns (single channel)
Noise	<50 µV rms referenced to input
Stability	10 µV/°C referenced to input (0-50°C).
Input Offset	±50 µV (adjustable)
Propagation Delay	2.2 ns per channel.
Recovery Time	<4 ns for a X20 overload.
Protection	±3.5 VDC ± 50 V transient.

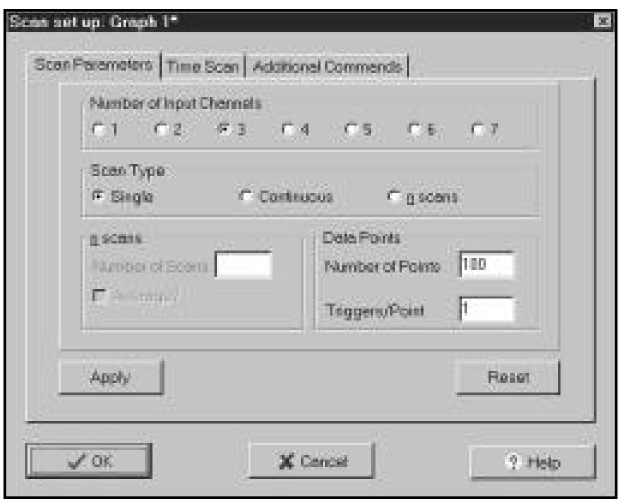
General

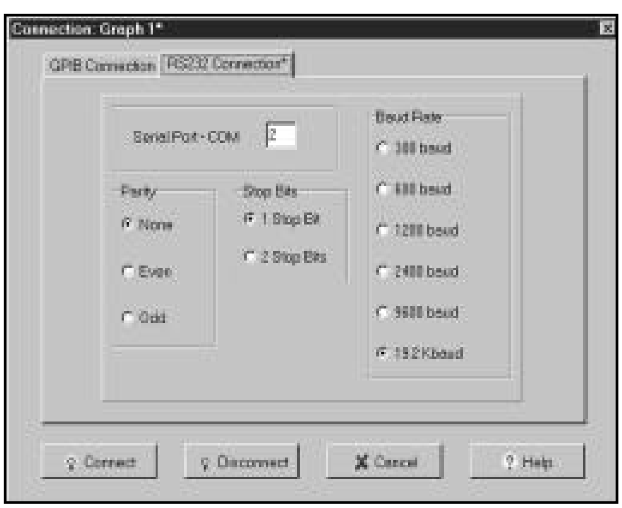
Mechanical	Single width NIM module
Power	+12V/300 mA, -12V/325 mA
Warranty	One year parts and labor on defects in material or workmanship

Gated Integrators and Boxcar Averagers

Model SR272 — Data Acquisition Software







SR272 Features

- ***Data acquisition from the SR250 Boxcar System***
- ***Automatic Scan Logging***
- ***Gate delay scans***
- ***Sample rates up to 1100 samples/s***
- ***Single or multiple shots per bin***
- ***Send commands to other instruments***
- ***RS232 or GPIB interface***

SR272 Requirements

- ***IBM PC or compatible, >66 MHz, >8 MB RAM***
- ***Windows95® Operating System***
- ***GPIB (NI or CEC) or RS232 interface***
- ***SR250 or SR255 Boxcar Averager***
- ***SR245 Computer Interface Module***

Ordering Information

SR272	**Boxcar Software**	**$ 500**

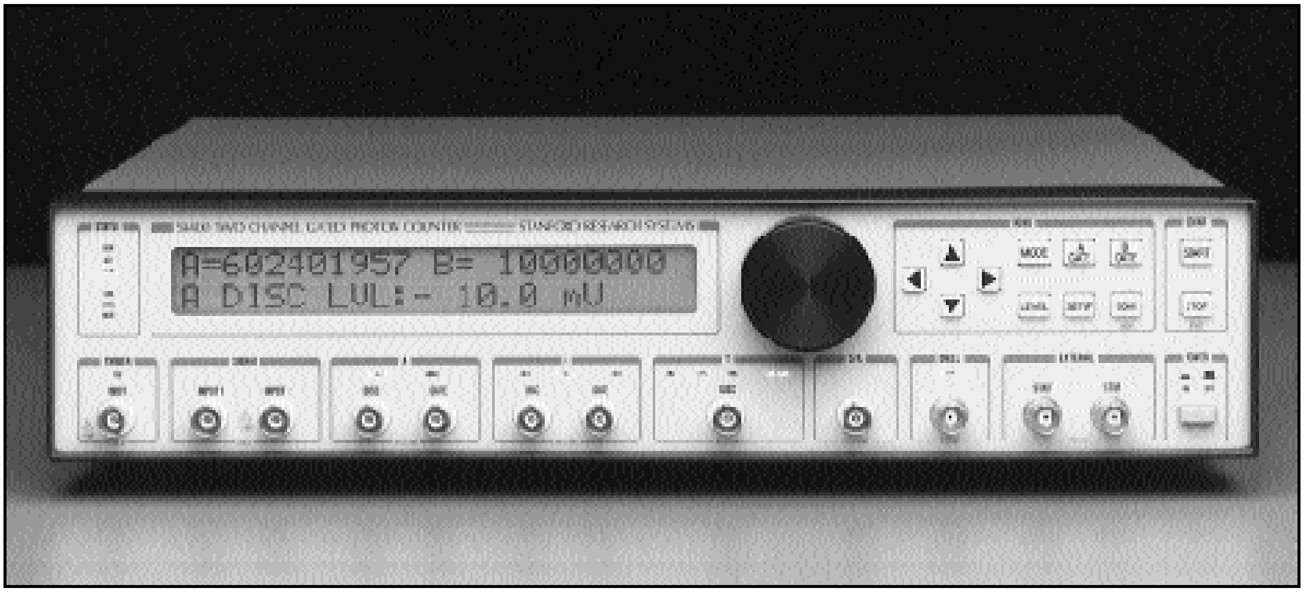
SR272 Overview

The SR272 Boxcar System software from SRS provides simplified data acquistion for the SR250 Gated Integrator and the SR255 Fast Sampler modules. With the SR245 Computer Interface module it provides a complete pulsed data acquisiton system. Up to 7 channels can be scanned simultaneously with fixed or scanning gate delays. Scans may be internally or externally triggered. Flexible averaging modes include scan averaging, or multiple triggers per point averaging.

The software provides all of the functionality of a Win95 application, inluding multiple graphs, zoom and autoscale, intuitive editing, cut and paste options and full Windows help. Data can be stored in various formats, including bitmaps, text files, CGM or MATLAB® files for complete flexibility in off-line analysis. A demonstation version of the SR272 is available at the Stanford Research Systems web site at www.srsys.com.

Photon Counters

Model SR400 — Dual Channel Gated Photon Counter



- ***Two 9-Digit Counters***
- ***200 MHz Counting Rate***
- ***Gated and Continuous Modes***
- ***Two Scanning Gate Generators***
- ***Three Scanning Discriminators***
- ***5 ns Pulse Pair Resolution***
- ***Gate and Discriminator Outputs***
- ***GPIB and RS-232 Interfaces***

SR400 Overview

The SR400 Dual Channel Gated Photon Counter offers a convenient, integrated approach to photon counting that avoids the complexity and expense of older counting systems. No longer is it necessary to mix and match amplifiers, discriminators, gate generators, and counters to get your photon counting experiment up and running. The SR400 combines all these modules into a single, integrated, microprocessor controlled instrument. Complex measurement tasks such as background subtraction, synchronous detection, source compensation, and pile-up correction can all be performed easily with the SR400.

Flexible Counting Modes

At the heart of the SR400 are three 200 MHz counters which can be configured for a virtually unlimited number of counting modes. Counting can be performed for a fixed amount of time, until a certain number of counts have been received or for a fixed number of triggers. Each counter has an associated fast, bipolar discriminator to allow window discrimination and pile-up correction. Fast 50Ω input amplifiers allow 10 mV sensitivity without any external preamplification, or 500 µV sensitivity with the addition of the SR445 preamplifier. Dual gate generators provide counting gates as short as 5 ns or as long as 1 s. Both gates can be scanned in order to measure lifetimes or recover waveforms. A D/A output provides a voltage level proportional to the count value, or log of the count value, for use with a chart recorder.

Interfaces and Software

The SR400 comes complete with RS-232 and GPIB interfaces. All instrument settings can be queried and set via the interfaces. Up to 2000 count values can be stored in an internal buffer, making it simple to take an entire scan and then transfer it to your computer, or to transfer the data as it is taken on a point by point basis. An optional PC compatible software package (SR465) provides data acquisition, display, and data reduction capabilities.

SR400 Features

A block diagram of the SR400 is shown below. There are three fast counters identified as counters A, B, and T. All three counters operate at rates up to 200 MHz. The input to each counter is selected from a number of sources, including the two analog signal inputs, the 10 MHz crystal timebase, and the external trigger input. Counter T can be preset to determine the measurement period. For pulsed experiments, counters A and B may also be synchronized to external events via the two independent gate generators. The gate generator provides gates from 5 ns to 1 s in duration with a delay from an external trigger ranging from 25 ns to 1 s.

Signal Inputs and Discriminators

There are two independent analog signal inputs labelled INPUT 1 and INPUT 2. They are internally terminated into 50Ω. The inputs can accept signals of either polarity up to ±300 mV and are protected to ±5 Vdc. Each input is followed by a DC to 300 MHz amplifier. This allows detection of pulses as small as 10 mV.

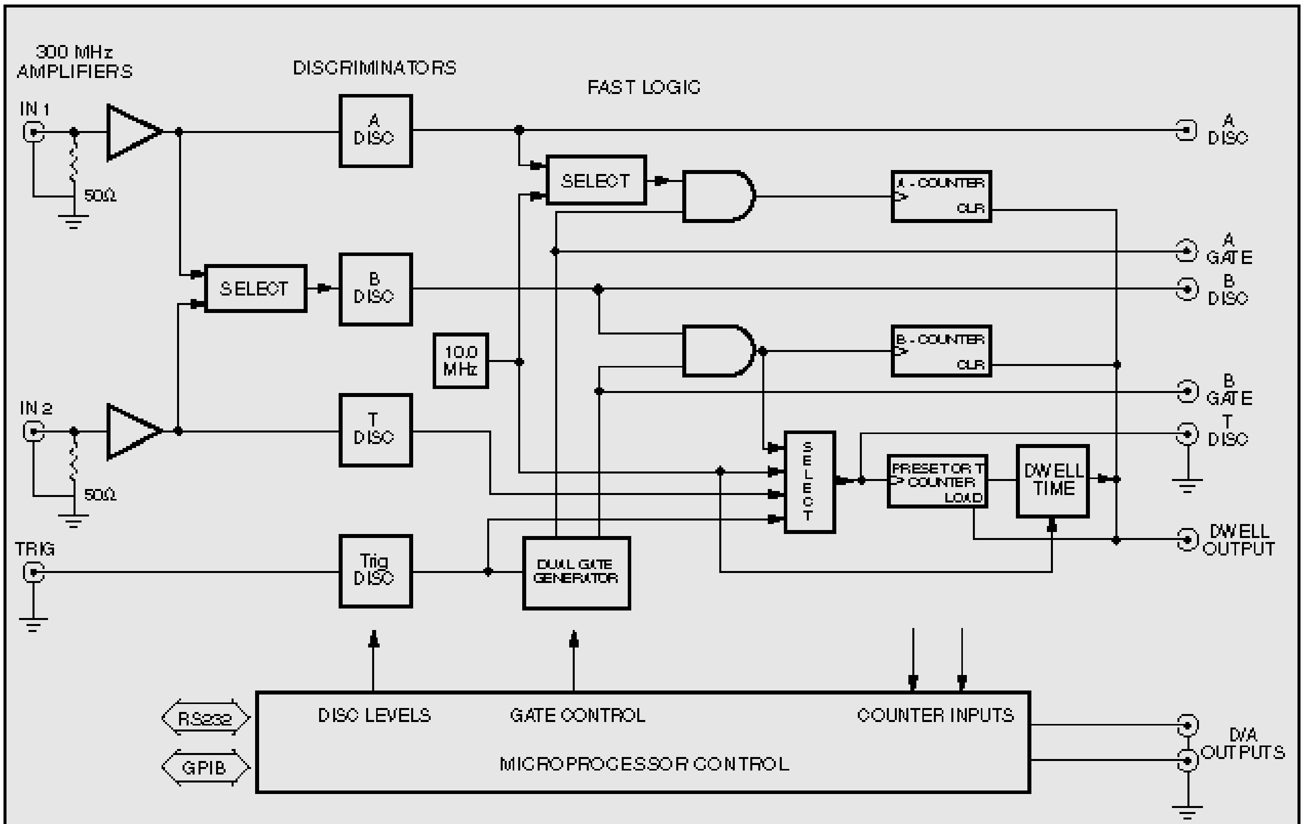
If greater sensitivity is required, the SR445 remote amplifier is available. The SR445 can provide gains from 5 to 125 at a bandwidth of 300 MHz.

There are three discriminators, one for each counter. Each discriminator has selectable slope and a threshold programmable from -300 mV to +300 mV in 0.2 mV steps (referenced to the inputs). Pulse-pair resolution is 5 ns and pulses of either polarity may be detected. Each threshold may be programmed to scan in either direction with selectable step size.

Counting

Each counter's input may be selected from a number of sources. All multiplexing of counter inputs is done internally to provide accurate timing and ease of operation. Input cables should rarely need to be swapped or disconnected to change measurement modes.

The actual inputs to the counters can be viewed as NIM level pulses from the discriminator outputs on the front panel. The discriminated pulses are negative going from 0 to -0.7 V. The falling edge is the active or count-



PHOTON COUNTER BLOCK DIAGRAM

ed edge regardless of the discriminator slope setting. The DISC outputs are very useful when adjusting discriminator thresholds or gate timing.

When the START key is pressed, or a START command is received from the computer interface, or the EXTERNAL START input is pulsed, counter T is enabled to count. The count period begins with the first pulse from T counter's input after START. Once the count period begins, counters A and B count their respective inputs. If the gates are enabled, only the pulses that occur during the gates and fall within the count period are counted.

The count period can be programmed to end in a variety of modes. The SR400 can be programmed to count its input for a fixed number of 10 MHz pulses, i.e. a fixed amount of time. Alternatively, the SR400 can count the number of 10 MHz pulses that occur before a preset number of input pulses are counted. This method has the advantage of spending more time measuring small count rates than large count rates thereby equalizing the signal to noise ratio in the two cases.

The SR400 may be programmed to cycle from one to 2000 count periods in a single scan. At the end of the programmed scan, the counters may be stopped or the scan may be restarted. Consecutive count periods are separated by the dwell time. The dwell time may be set from 2 ms to 60 s. During the dwell time, counting is disabled and data may be transferred or external parameters may be changed. The dwell output provides a TTL output which is high during the dwell time. This can be convenient for interfacing other instruments used in the experiment.

Scanning

In all scan modes, a number of parameters may be scanned. These parameters are the three discriminator thresholds (Pulse Height Analysis), the two gate delays (Boxcar mode), and the two D/A output ports (X-axis of recorder, scope, or analog control of other apparatus).

After each count period, each scanning parameter is adjusted by one step. All changes are made during the DWELL time so that all values are stable during the count periods. The scan limits are determined by the start position, the step size, and the number of periods in a scan. When the counters are reset, all scanned parameters return to their start positions.

Outputs

The front panel can display counts up to $10^9 - 1$. Counters A and B can be displayed as separate counters or combined as A-B or A+B. The front panel D/A output provides an analog output proportional to A, B, A-B, or A+B depending on the counting mode. The scale may be logarithmic (1V/decade) or linear.

If a chart recorder is used for output, the D/A output can drive the Y axis. The X axis can be driven by the recorder itself, or either PORT1 or PORT2 output can provide a ramp voltage to drive the X axis. This latter method allows accurate determination of the X value of each point. If the DWELL output is used as the pen lift, the points will be disconnected.

Interfaces

Built-in RS-232 and GPIB interfaces provide a convenient means of controlling the instrument and retrieving data. While the SR400 is scanning, each of the count values for the A and B counters are stored in an internal buffer. This buffer can be transferred on a point by point basis, or dumped all at once through either interface. A PC compatible program, the SR465, is available to provide a complete ready-to-use data acquisition and analysis program for the SR400. The SR465 is described more fully later on.

Overview

The SR400 is a two channel gated counter that provides discriminators, amplifiers, counters and gates in a microprocessor-based design to greatly simplify data collection for photon and event counting. The architecture and flexibility make it simple to configure for a number of counting modes, and the standard computer interfaces make it easy to collect and analyze data.

Signal Inputs

Bandwidth	DC to 300 MHz
Input impedance	50 Ω
Linear range	± 300 mV (at input)
Input protection	± 5 Vdc, 50 V for 1 µs
Overload recovery	5 ns for <10 µs duration overload

Discriminators

Counters A, B, and T have independent discriminators when counting the signal inputs. All discriminator levels may be set to a fixed level or scanned. Referenced to the signal inputs:

Discriminator range	-300 mV to +300 mV.
Discriminator slope	Rising or Falling
Resolution	0.2 mV
Input offset voltage	1 mV
Minimum pulse input	10 mV
Pulse pair resolution	5 ns.
DISC outputs	NIM levels into 50 Ω
Inhibit input	TTL high stops count

Trigger Input

Impedance	10 kΩ
Threshold	±2.000 Vdc in 1 mV steps
Slope	Rising or Falling
Protection	15 Vdc, 100 V for 1 µs

Gate Generators

Both the A and B gates may be fixed in time or scanned. The gate outputs show the positions of the gates with respect to the discriminator outputs.

Insertion delay	25 ns
Maximum delay	999.2 ms
Minimum gate width	5 ns
Maximum gate width	999.2 ms or CW
Resolution	0.1%, 1 ns minimum
Accuracy	2 ns +1%
Jitter	200 ps rms +100 ppm
Maximum trigger rate	1 MHz
GATE view output	NIM levels into 50 Ω
GATE view error	< 2 ns

Scan and Dwell

The number of count periods or data points in a scan may be set from 1 to 2000. The duration of one count period is determined by the preset condition.
The time between consecutive count periods is the dwell time and can be set from 2 ms to 60 s. The dwell output will be TTL high during the dwell time. This output can be used to trigger external devices. At the end of a scan (of 1 to 2000 count periods) counting may be programmed to stop or start the scan over again. The start key begins the first count period of the programmed scan.
The stop key terminates the current count period and pauses the scan. If scanning, gates and disc levels are held at their current values. The stop key pressed while in a paused condition will reset the scan and all scanned parameters will return to their start values. The start key pressed while paused resumes the scan by starting the next count period.
The dwell time may also be set to external. In this mode, count periods begin with the start key or external start input (TTL rising edge). Count periods terminate with the preset condition, the stop key, or the external stop input (TTL rising edge). A stop key while not counting resets the scan.
All count data is internally buffered for one scan. Data may be read over the computer interfaces during or after a scan.

Display Mode

Continuous	Displays current counter value
Hold	Displays final count value

D/A Output

The front panel D/A Output is proportional to A, B, A-B, or A+B, and is updated at the end of each count period. There are two rear panel D/A outputs, port 1 and port 2. These outputs may be set or scanned from the front panel or via the computer interface.

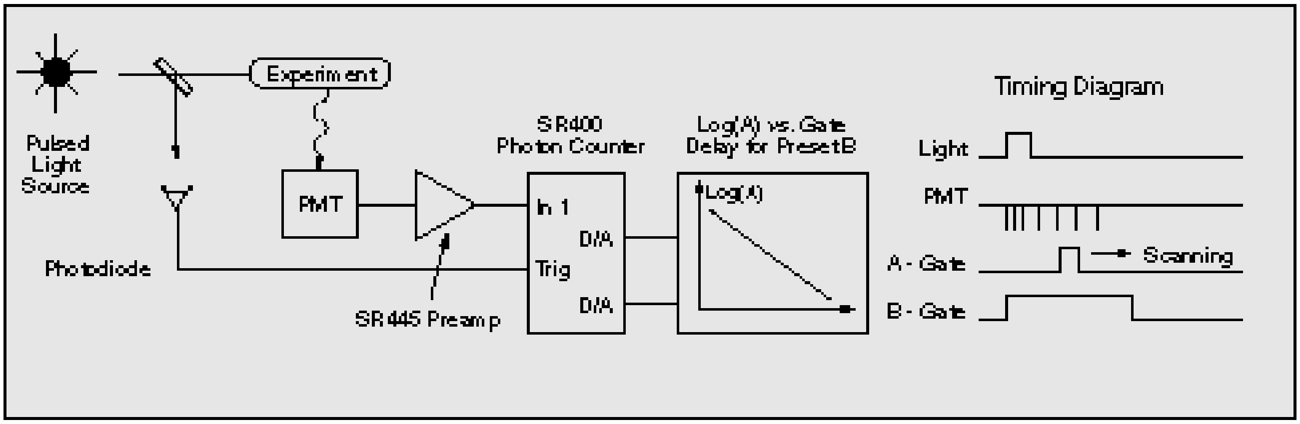
Full scale	± 10 Vdc
Resolution	12 bits (5 mV)
Current rating	10 mA
Output Impedance	< 1 Ω
Accuracy	0.1% + 5 mV

General

Interfaces	IEEE-488 (GPIB) and RS-232
Dimensions:	16" x 13" x 3.5"
Weight:	10 lbs
Power:	35 Watts from 100, 120, 220, or 240 Vac
Warranty:	One year parts and labor on any defects in materials or workmanship

Ordering Information

SR400	**Gated Photon Counter**	**$ 5350**
SR465	**Photon Counter Software**	**$ 500**
SR445	**Quad Preamplifier**	**$ 1100**



Photon Counter Examples

Experiment #1 — (Boxcar) Mode

This experiment uses a scanning gate to measure the lifetime of an excited state pumped by a pulsed laser. The SR400's gate generators are triggered by a photodiode when the laser fires. Counter A counts photons which occur during a narrow gate, while counter B counts the photons during the entire decay.

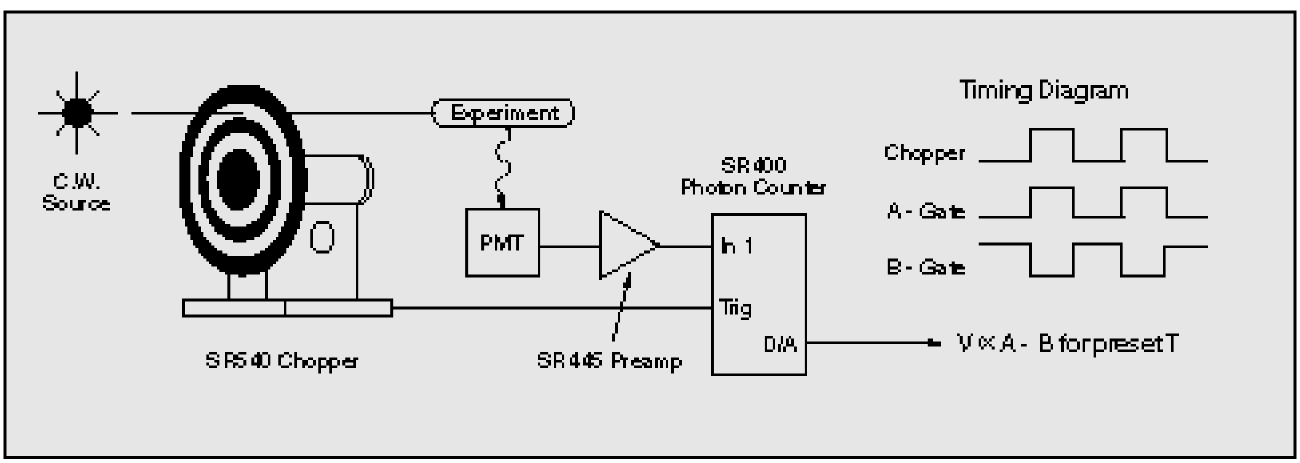
Decay data is normalized to the source intensity by counting until a preset value is reached in counter B. When B reaches its preset condition, the D/A output is set to a voltage proportional to A's count, A's gate is stepped, and a new count interval begins.

By plotting the Log of A's count value vs A's gate delay, the exponential decay curve is linearized and the lifetime of the excited state can be determined from the slope on the chart recorder. If count rates are high and count periods short, then scans may be displayed on an X-Y scope. The D/A output is the Y drive and port 1 or port 2 is the X drive. The dwell output should be used as the blanking pulse

Experiment #2 — Synchronous (Lock In) Mode

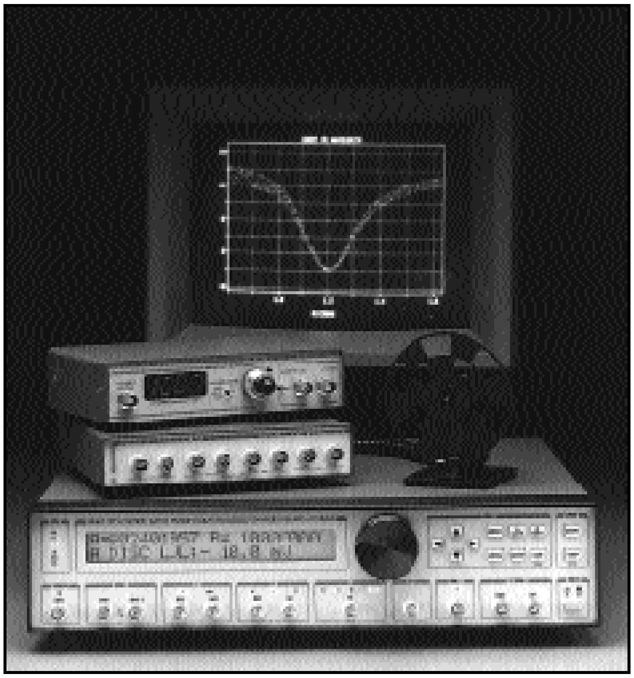
Very small changes in the flux of photons may be measured by synchronous detection. If a signal is fixed in frequency and has a 50% duty cycle, then synchronous photon counting, or photon counting in a 'lock in' mode, can be used. An optical chopper is used to modulate a CW light source. The reference output from the chopper triggers the photon counter's dual gate generator. The A gate is positioned to count photons during the open phase of the chopper and thus counts the signal plus background. The B gate only counts the background, counting pulses only during the closed cycle of the chopper. The difference between the two counts, A-B, is the signal. Accumulating data over many cycles is required to measure the signal since the background rate often far exceeds the signal rate.

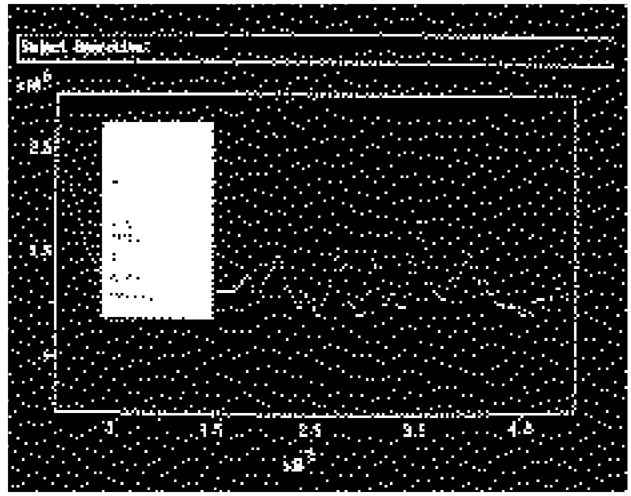
The D/A output is proportional to A-B. Since the background count is subtracted for each chopper cycle, only the signal which is synchronous with the chopper will cause the output of the photon counter to change.



Photon Counters

Model SR465 — Photon Counter Software





- ***Real Time Graphic Displays***
- ***On-Screen Data Reduction***
- ***Linear and Exponential Fitting***
- ***Hardcopy Output and Disk Storage***
- ***Complete Instrument Control***
- ***Works with GPIB or RS-232***

SR465 Requirements

Requirements

DOS v2.1 or greater
640 K RAM
HGC, EGA,CGA or VGA graphics card
GPIB (National or CEC) or RS-232 interface
Hard disk
SR400 Photon Counter

Recommendations

12 MHz 80286 or Faster Processor
Math coprocessor
Dot Matrix Printer
HP-GL plotter

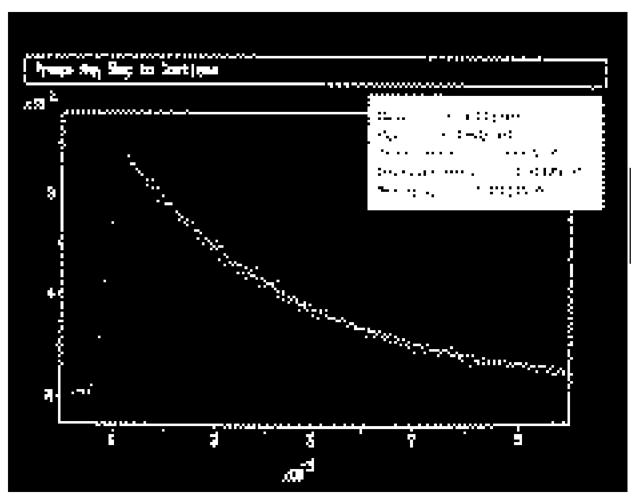
Ordering Information

SR465	**Photon Counter Software (Executable Code Only)**	**$ 500**

SR465 Features

The SR465 is a complete data acquisition, display, and analysis program for use with PC compatible computers and the SR400 photon counter. The program acquires data from the SR400 over RS-232 or GPIB and allows you to plot count values against time, gate delay, or rear panel D/A voltages. A variety of scaling options including autoscale and zoom make it easy to optimize the screen display. Scans can be stored and recalled from disk and the file format is documented to allow your programs easy access to data taken with the SR465.

Once data has been acquired, the SR465 offers a number of analysis options. Savitsky-Golay smoothing, curve fitting to linear, exponential, and gaussian functions, and statistical parameter calculation are all available with the SR465 program.



Photon Counter Preamplifier

Model SR445 — Quad 300 MHz Preamplifier

- ***DC to 300 MHz Bandwidth***
- ***1.2 ns Rise/Fall Time***
- ***2.8 nV/√Hz Input Noise***
- ***Voltage Gains to 125***
- ***Selectable Input Impedance***
- ***Fast Overload Recovery***
- ***Four Independent Channels***
- ***DC Offset Correction***

SR445 Overview

The Model SR445 Fast Preamplifier contains four wide bandwidth DC coupled amplifiers each with a gain of 5. The fast rise time, low noise, and excellent DC accuracy of the SR445 make it an ideal instrument for amplifying the outputs of fast photomultiplier tubes and photodiodes.

The SR445 Fast Preamplifier is useful for amplifying small signals to levels that allow processing by photon counters. Typically, photon pulses should be greater than about 10 mV at the input of the SR400 or SR430. To obtain these signal levels, the four channels may be used independently, or up to three may be cascaded to provide a total gain of 125. To ensure linear operation, the maximum input voltage should be kept below the values shown in the following table:

Number of Channels	Gain	Maximum Input
1	5	200 mV
2	25	40 mV
3	125	8 mV

The input impedance of channel 1 can be increased to approximately 500Ω by a front panel switch. This can improve the sensitivity to signals from current and charge output devices, such as photomultiplier tubes. Each channel has a separate DC offset adjustment allowing you to quickly null out DC errors.

Overview

The SR445 is a DC-300 MHz bandwidth, 50Ω, four channel preamplifier. Each channel has a gain of 5, and up to three channels can be cascaded to provide gains of up to 125. The SR445 is ideal for amplifying low-level outputs from photomultipliers and photodiodes before being detected by the SR400 gated photon counter or the SR430 multichannel scaler.

Amplifier

Input	50Ω impedance, DC coupled, BNC connectors. (Channel 1: 50Ω or 500Ω)
Outputs	DC coupled, BNC connectors (50Ω termination).
Operating Range	Inputs: ± 200 mV, Outputs: ±1.0 V.
Voltage Gain	5 per channel. Up to 3 channels can be cascaded.
Bandwidth	DC to 300 MHz (-3dB)
Noise	< 50 µV rms referenced to input (2.8 nV/√ Hz).
Stability	10 µV/°C referenced to input (0-50°C).
Input Offset	± 50µV (adjustable)
Propagation Delay	2.2 ns per channel.
Rise/Fall Time	1.2 ns (single channel)
Recovery Time	< 4 ns for a x20 overload.
Protection	± 3.5 Vdc, ± 50 V transient.

General

Mechanical	7.7" x 6.7 " x 2"
Power	16 W, 100/120/220/240 VAC
Warranty	One year parts and labor on any defects in material or workmanship.

Ordering Information

SR445	**300 MHz Quad Preamplifier**	**$ 1100**

Photon Counters

Model SR430 — 5 ns Multichannel Scaler/Averager

- ***5 ns to 10 ms Bin Width***
- ***100 MHz Count Rate***
- ***1 k to 32 k Bins Per Record***
- ***Built In Discriminator***
- ***No Interchannel Dead Time***
- ***On Screen Data Analysis***
- ***Menu Based User Interface***
- ***Hardcopy Output to Printers and Plotters***
- ***DOS Compatible 3.5" Drive***
- ***GPIB and RS-232 Interfaces***

SR430 Overview

The SR430 is the first multichannel scaler combining amplifiers, discriminators, bin clocks, and data analysis in a single integrated instrument. The SR430's 5 ns time resolution makes it easily the fastest commercially available multichannel scaler. With its many features and its easy-to-use menu driven interface, the SR430 easily performs measurements that are far more difficult with other less capable instruments.

What is a Multichannel Scaler/Averager?

Multichannel Scalers can be thought of as time-resolved photon counters. A trigger starts a bin clock with adjustable bin widths from 5 ns to 10 ms. The instrument records how many photons arrive in each bin. Between 1k and 64k bins can be counted in each time 'record', following the trigger. Since all bins are counted on each trigger, the SR430 acquires a waveform much faster than an equivalent single channel photon counter with a scanning gate. The 'Averager' part of the instrument's name refers to its ability to add the data from successive records. Even if in a single record there are only a few photons per bin, by taking repetitive shots and adding, you can watch the signal grow out of the noise. Multichannel scaler/averagers are useful in a variety of applications where it is necessary to count events as a function of time: LIDAR, time of flight mass spectroscopy, and fluorescence decay measurements are just a few examples.

Built-in Analysis Capabilities

The SR430 is not just a data acquisition instrument–a complete range of data analysis features are included. Savitsky-Golay smoothing, curve fitting to linear, exponential, and gaussian functions and statistical calculations are just some of the SR430's capabilities.

SR430 Features

Input and Discriminator

The analog signal input has a 50 Ω input impedance and can accept signals of either polarity up to ±300 mV. Input signals should be at least 10 mV–if the signal level is less than this the SR445 preamplifier should be used. The input is followed by a discriminator with a selectable slope and a threshold adjustable between ±300 mV. The discriminator output is a NIM level (0 to –.7 V, active low) signal available at the front panel. The discriminator is always active, even when the SR430 is not taking data, in order to simplify setting the threshold level.

Trigger Timing

A valid front panel trigger pulse starts the SR430's data acquisition cycle. After a fixed insertion delay of 45 ns, the front panel SYNC/BUSY output will go high to indicate the start of the first bin. At the same time, the BIN CLK output will begin to toggle its output state at the boundary of each individual time bin. The SR430 provides 20 fixed internal bin widths ranging from 5 ns to 10.486 ms. Alternatively, an external bin clock can be provided to the SR430 allowing you to define your own bin size. The minimum bin width with an external bin clock is 250 ns.

The number of bins in each record is adjustable from 1k (1024) to 32k in 1k increments. Data acquisition in the SR430 is seamless–there is no dead time between bins. Once the selected number of bins has been recorded, the BIN CLK output stops toggling and the SR430 either adds or subtracts the result of the current record from the accumulated bin totals. During this time, the SYNC/BUSY output remains high indicating that the SR430 is not yet ready for another trigger. The total time that SYNC/BUSY remains high is given by the formula:

$$T_{busy} = (N_{bins} x T_{bin}) + (N_{bins} x 250ns) + 150\ \mu s$$

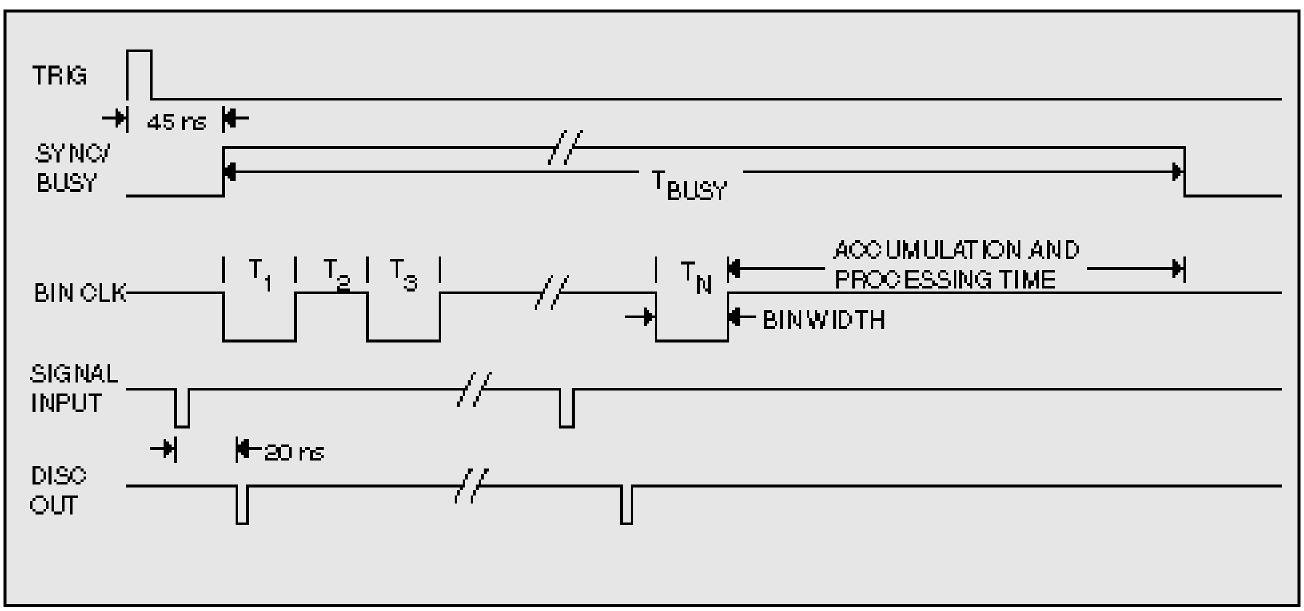
The first term in this equation represents the data acquisition interval, i.e. the time during which discriminated pulses are counted by the SR430. The last two terms represent the additional time over the data acquisition time required by the SR430 to process the results of each record. The total time SYNC/BUSY is high limits the maximum trigger rate of the instrument. If triggers are received at a faster rate the unit will indicate a RATE error and ignore the trigger.

Accumulation

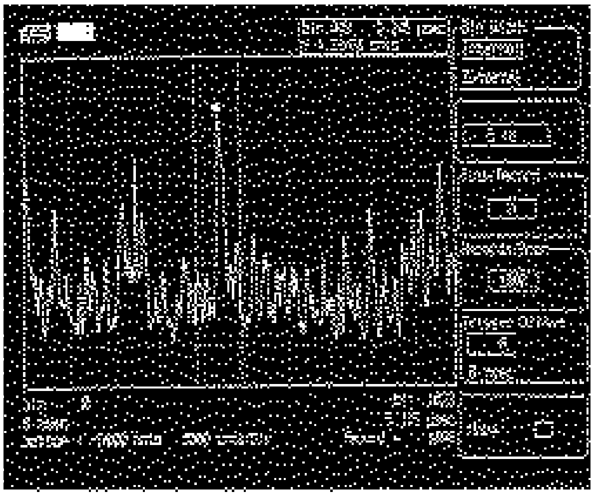
The SR430 can be programmed to accumulate between one and 64k records, or set to free run. Each record can be added or subtracted from the current accumulator totals. The instrument can be set to toggle between add and subtract every N records, or an external toggle input can select the polarity of the next record. A rear panel inhibit input allows you to selectively prevent the accumulation of any given record. The screen display is updated continuously as records are accumulated, providing a live, real-time, display of the data.

Data Display

The 7" CRT display allows complete flexibility in dis-



playing your results. Between 8 and 16k bins can be displayed on the screen at any time, and complete horizontal and vertical zooming and scrolling features are provided. An "Autoscale" key quickly optimizes the screen for the current data with a single keypress. A fast, responsive, on-screen cursor lets you read the maximum, minimum or mean data value from a selected range of the graph.

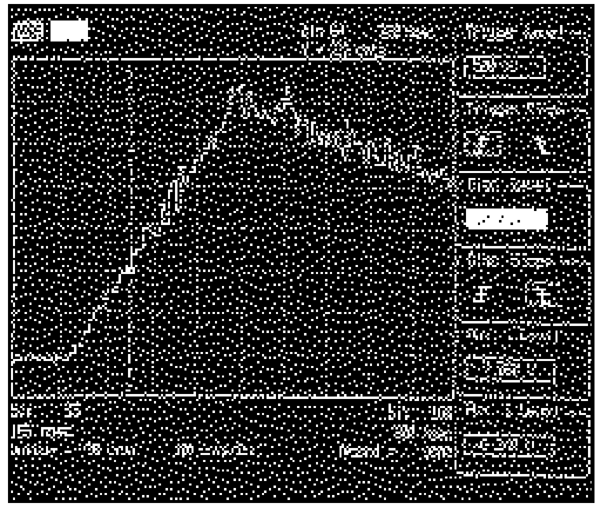


Menus and Softkeys

The SR430 is based on a simple menu oriented user interface. Each menu groups related instrument functions and defines softkeys to control those functions. The instrument settings are changed by pressing the softkeys or by turning the front-panel knob. A PC keyboard can be used to enter text information in menu fields as well. Complete context sensitive help is provided for all menus and softkeys. The remote command list is even provided on a help screen as an aid in programming the SR430.

Data Analysis

The SR430's extensive capabilities don't stop with data acquisition. Savitsky-Golay smoothing can be applied to any portion of the data with selectable smoothing intervals. Gaussians, exponentials or straight lines can be fit to arbitrary regions of the display, allowing you, for instance, to quickly determine decay lifetimes. Basic statistical parameters can be calculated for data regions including total number of counts, mean number of counts and variance. Basic arithmetic operations, including addition, subtraction, multiplication, division, logs, and square roots, can be applied to the current data or to data in disk files.



Built-in Disk Drive

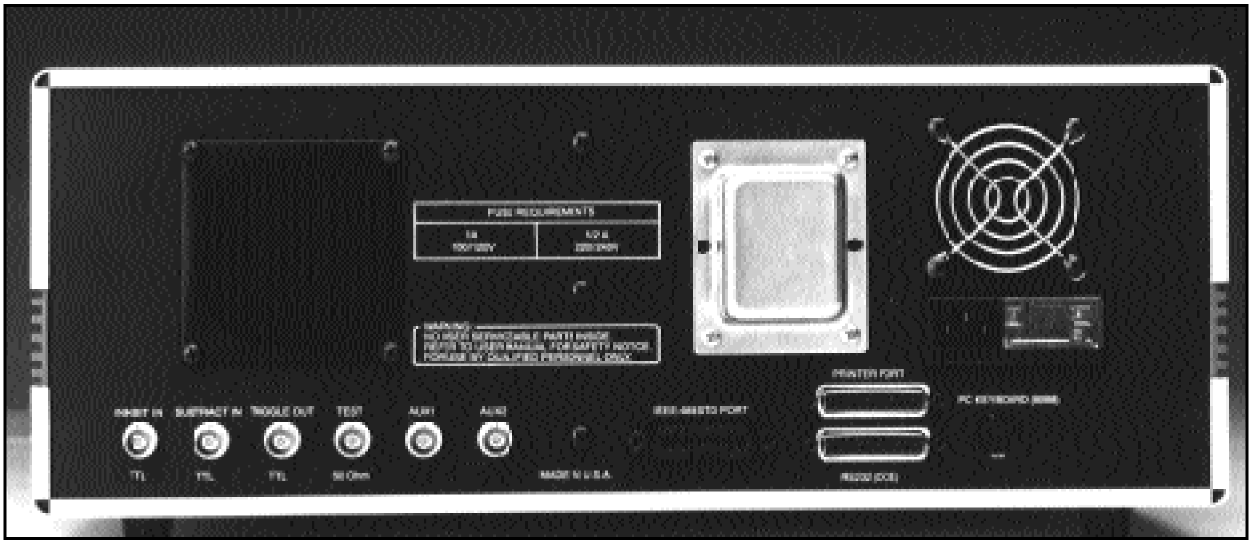
The SR430 has a built-in 3.5" DOS compatible floppy disk drive to simplifiy data transfer between the instrument and your computer. Unlike the disk drives found on some manufacturer's instruments, the SR430 uses a standard DOS 720k format that can be read on any PC or Macintosh (Reading the files with a Macintosh requires Apple File Exchange). Both data files and settings files can be stored so you can store complete instrument setups for a variety of situations and recall them instantly.

Hardcopy Output

Hardcopy output is available from the SR430 in a variety of forms. A standard Centronics printer port lets you dump the screen to dot-matrix or LaserJet compatible laser printers at any time. Additionally, the SR430 can plot its display on any HP-GL compatible plotter via the RS-232 or GPIB interface.

Complete Programmability

Both RS-232 and GPIB interfaces are standard on the SR430 ensuring compatibility with the widest range of laboratory computers. All instrument settings and functions can be read and set via the interfaces. A complete list of all characters received and transmitted over the interfaces can be displayed on the CRT display–an invaluable aid when debugging your programs. Numerous modes are available for downloading the count data to your computer including ASCII transfer, binary transfer, and a special fast binary dump mode which transfers data continuously over the GPIB interface as it is being acquired by the instrument.



SR430 Rear Panel

Overview

The SR430 Multichannel Scaler/Averager is a counting instrument with a discriminated front end, capable of counting pulses at a 100 MHz rate. Each new trigger starts a record whose data is added to the accumulation of all of the previous records. The double buffered design ensures that there is no dead time between bins and that no pulses are missed at the bin boundaries. Up to 32,767 counts can be accumulated per bin, and up to 65,535 scans can be added together. A number of integrated capabilites provide flexibility for a wide range of experiments, including: autoscale, smoothing, curve fitting, non-volatile setup memories, data storage to disk, push button hardcopy to graphics printers and plotters, and IEEE488 and RS-232 interfaces.

Signal Input

Bandwidth	DC to 250 MHz
Input impedance	50 Ω
Linear range	±300 mV (at input)
Input protection	±5 Vdc, 50 V for 1 µs
Overload recovery	5 ns for <10 µs duration overload

Discriminator

Discriminator range	-300 mV to +300 mV
Resolution	0.2 mV
Slope	Positive or Negative
Accuracy	2 mV + 1%
Min. pulse amplitude	10 mV
Pulse pair resolution	10 ns (typical)
DISC view output	NIM level into 50 Ω There is a 20 ns insertion delay from the signal input to the discriminator output.

Trigger Input

Impedance	10 kΩ
Threshold	-2.000 V to +2.000 V in 1 mV steps
Slope	Rising or Falling
Protection	15 Vdc, 100 V for 1 µs

Internal Time Bins

Bin width	5 ns, 40 ns, 80 ns, 160 ns, 320 ns, 640 ns, 1.28 µs, 2.56 µs, ...10.486 ms. (10ns and 20 ns bins are not available)
Accuracy	1 ns + 20 ppm of bin width
Jitter (rms)	100 ps + 10 ppm of delay from SYNC/BUSY output (Time bins are synchronous with the SYNC/BUSY output).
Indeterminacy	2.5 ns pk-pk with respect to the TRIGGER input
Insertion delay	45 ns from trigger to first bin. Rising edge of SYNC/BUSY output occurs at the beginning of the first bin. However, signal pulses arriving 25 ns after the trigger will be counted in the first bin.

Externally Clocked Time Bins

EXT BIN CLK Input	Rising edge triggers next time bin.
Maximum frequency	4 MHz (250 ns minimum bin width)
Minimum time high	100 ns
Minimum time low	100 ns
Insertion delay	Rising edge of SYNC/BUSY output occurs at first rising edge of EXT BIN CLK after TRIGGER. The beginning of the first bin occurs at the same time. (Time bins are synchronous with the SYNC/BUSY output).

Counters/Accumulation

Bins per record	1k to 16k in 1k increments (1024 to 32,704 including Trigger Offset)
Max. count rate	100 MHz
Max. count	32,767 per bin per trigger
Records/accumulation	1 to 64k (or free run)
Max. accumulation	32,767 per bin in Add mode, ±16,383 per bin in Toggle, or External mode.
Add/Subtract	Records may be added or added and subtracted (Toggle between add and subtract every N triggers where N is programmable). External subtract input may also control the toggle.

Trigger Rate

Minimum Trigger Time	Tp = Record time + Accumulation time + Overhead, or, Tp = (Number of Bins x Bin Width) + (Number of Bins x 250 ns) +150 µs. SYNC/BUSY output is high for Tp after each trigger. When SYNC/BUSY returns low, the next record may be triggered. Triggers received while SYNC/BUSY is high are ignored.

Outputs

DISC	NIM level into 50 Ω. Low whenever the signal input exceeds the discriminator level with the correct slope.
SYNC/BUSY	TTL level. Rising edge is synchronous with the first time bin of each record. Use this edge to trigger the experiment. Remains high until re-armed for next trigger.
BIN CLK OUTPUT	NIM level into 50 Ω. Each transition is a bin boundary. Active only while a record is being acquired. Timing skew relative to the DISC output is less than 2 ns.
TOGGLE	TTL level. Indicates whether the next record will be added to or subtracted from the accumulation. (Internal toggle mode)

TEST	50 MHz NIM output into 50Ω. Used to test counters.
AUX1,AUX2	General purpose analog outputs
Full Scale	±10 V
Resolution	5 mV
Output Current	10 mA
Output Impedance	<1Ω
Accuracy	0.1% + 10 mV

Inputs

SIGNAL	Analog 50Ω input.
TRIGGER	10 kΩ input.
BIN CLK INPUT	TTL input. Rising edge triggers next time bin.
ACC. INHIBIT	TTL input, sampled each trigger. If high, causes the current record to be ignored (not accumulated).
SUBTRACT	TTL input, sampled each trigger. If high, causes the current record to be subtract-ed from the accumulation (in external toggle mode).

General

Interfaces	IEEE-488, RS-232, and Centronics printer standard. All instrument functions can be controlled and read through the IEEE-488 and RS-232 interfaces.
Data Transfer	16 kBins in 500 ms
Hardcopy	Screen dumps to Epson compatible dot matrix or HP LaserJet printers (parallel). Plots to HPGL compatible plotters (serial or IEEE-488).
Disk	3.5" MSDOS compatible format, 720kbyte capacity. Storage of data and setups.
Power	60 Watts,100/120/220/240 Vac, 50/60Hz
Dimensions	17"W X 6.25"H X 16.5"L
Rackmounts	Included
Weight	30 lbs
Warranty	One year parts and labor on any defects in material or workmanship

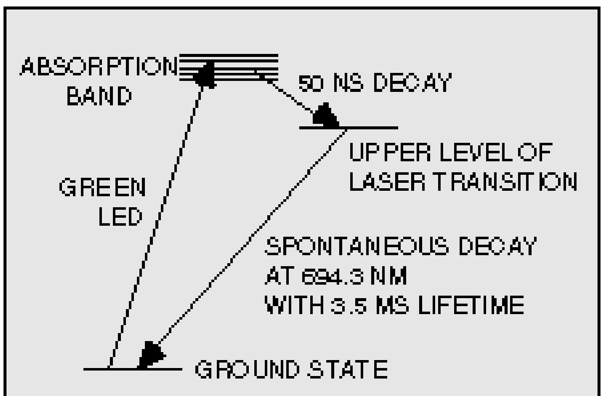
Ordering Information

SR430	**Multichannel Scaler**	**$ 7950**
SR445	**Quad 300 MHz Preamplifier**	**$ 1100**
O760H	**Carrying Handle**	**$ 100**

Multichannel Scaler Application: Fluorescence Decay of Ruby

This experiment is typical of time resolved photon counting experiments. A pulsed light source is used to pump atoms to an excited state. Fluorescent decay from the excited state is observed, allowing the lifetime of the upper state to be measured.

The energy level diagram of ruby is shown below. There are absorption bands around 400 nm and 550 nm. The Cr+++ ions which absorb light at these wavelengths decay in about 50 ns to the upper state of the well known laser transition. This state has a lifetime of about 3.5 ms, and decays to the ground state by emitting a photon at 694.3 nm.



ENERGY LEVELS FOR RUBY

The absorption band at 550 nm overlaps the emission line of a green LED. In this example experiment, a pulsed green LED is used to quickly populate the excited state, and decays from the excited state are seen through a bandpass interference filter centered on ruby's 694.3 nm emission line.

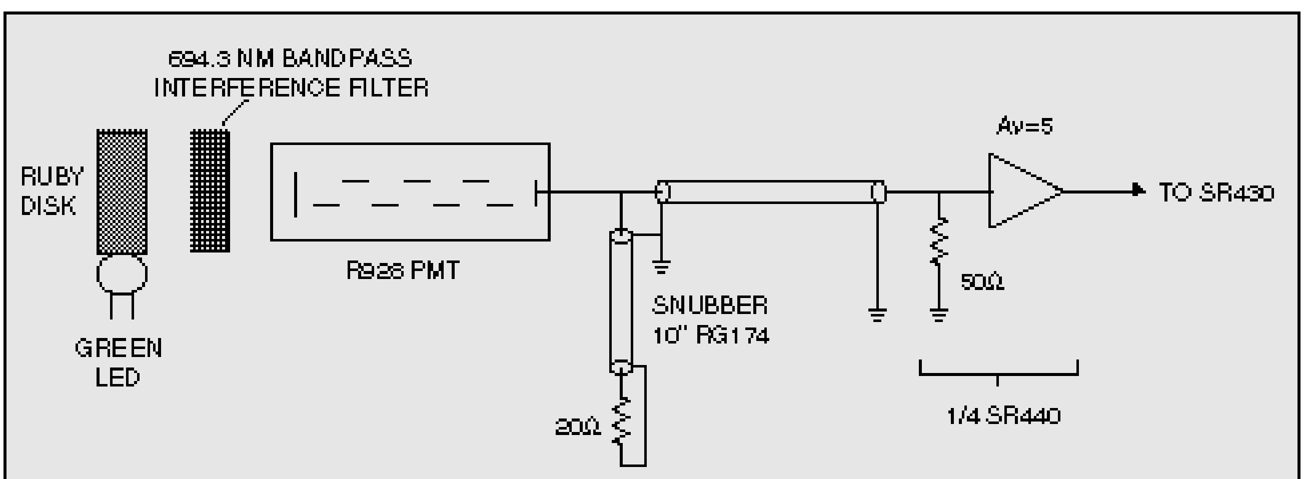
Apparatus

The experimental set-up is shown below. The green LED is glued to the edge of a 1 cm diameter, 3 mm thick, ruby disk. The ruby disk is viewed through the bandpass interference filter by a Hamamatsu R928 PMT. This side-on PMT was selected for its high gain, fast rise time, and good red sensitivity.

The phototube base uses a tapered voltage divider, with about 3x the normal interstage voltage between the photocathode and the first dynode. This helps to narrow the pulse height spectrum for single photon events. The lower dynodes are bypassed, and 100Ω resistors are used between the dynodes and their bypass capacitors to reduce ringing in the anode signal. A snubber network consisting of a 10 inch piece of RG174 terminated into 20 Ohms is used to further reduce anode ringing and reduce the falltime of the output current pulse.

Operation

The PMT is operated at the maximum rated high voltage (1250 Vdc). The output pulses have a mean amplitude of 20 mV into 50 Ohms. To increase the pulse height to 100 mV, one amplifier in the SR445 preamp provides a gain of 5 with a 300 MHz bandwidth. The discriminator threshold is set to 20 mV. When viewed



APPARATUS FOR DEMONSTRATION EXPERIMENT

with a 300 MHz oscilloscope, it is apparent that this threshold setting will count the majority of output pulses, but will not count anode rings or amplifier noise.

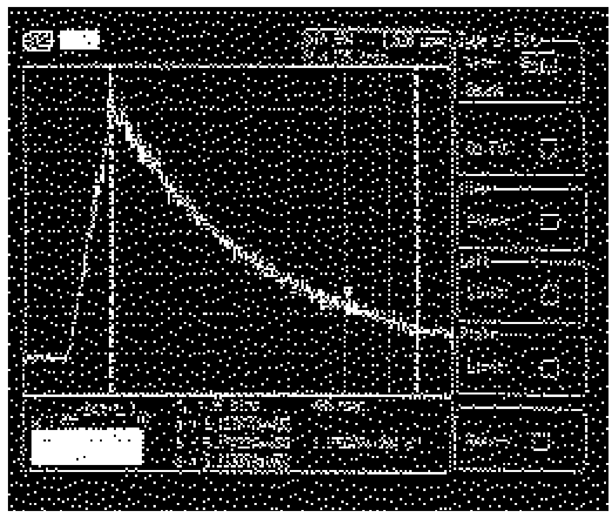
The green LED is flashed at a 40 Hz rate for about 1 ms. During this time, population integrates in the upper level of the laser transition. Spontaneous decays from the upper level are counted by the SR430.

Instrument Configuration

The multichannel scaler is triggered by the same pulse which flashes the LED. The bin width is 20.48 µs and the record length is 1k bins. The records per accumulation is set to 100. Thus each record takes approximately 21 ms of real time to acquire which is sufficient to measure the 3.5 ms lifetime of the excited state. The pulse rate of 40 Hz will not generate rate errors. A summary of the SR430 setup parameters is shown below.

Data Acquisition

After the SR430 Levels and Mode menus have been setup, data acquisition may begin. Pressing the [START] key starts the first record. Data accumulates on the screen until all 100 records have been acquired. When data acquisition is complete, the math menu may be used to fit an exponential curve to the data to measure the lifetime directly. Finally, the data curve is printed or plotted and stored to disk. A picture of an actual decay curve obtained with an SR430 is shown below along with the exponential fit to the data. The decay time measured by the SR430 is 3.5 ms, quite close to the actual value.

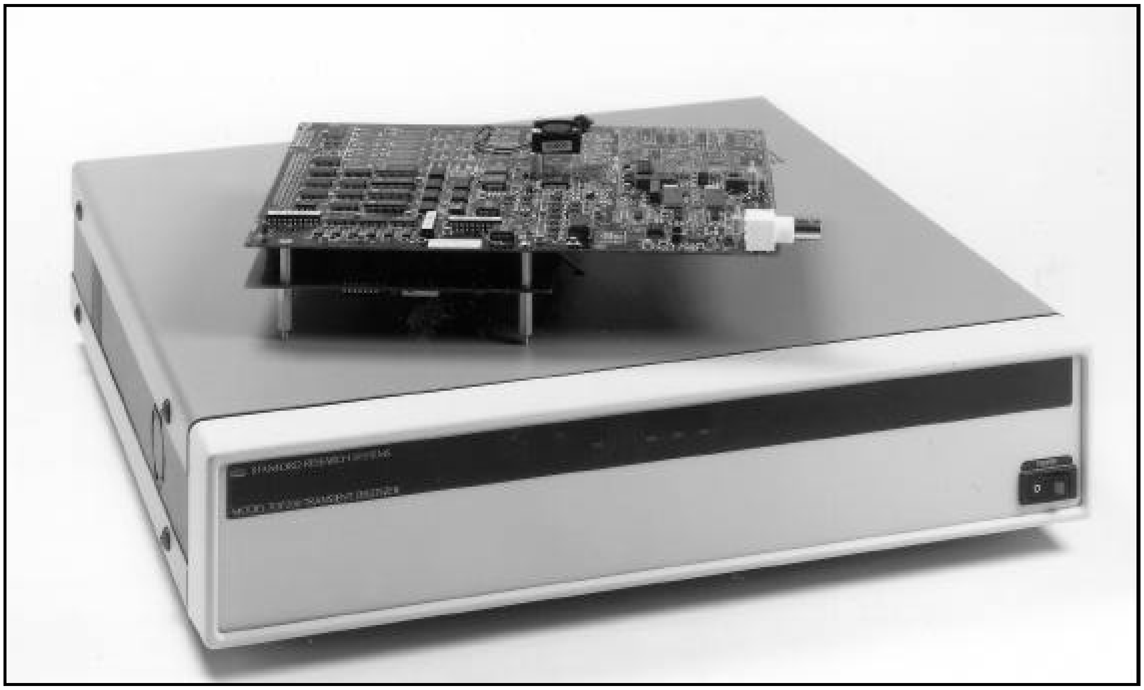


SR430 Configuration for Ruby Experiment

Levels:	**Trigger Level**	**+1.000 V**	**Trigger threshold set to +1.000V**
	Trigger Slope	**RISE**	**Trigger on rising edge of LED trigger**
	Disc Level	**-20.0 mV**	**Disc threshold set to -20.0 mV**
	Disc Slope	**FALL**	**Discriminate negative pulses**
Mode:	**Bin Clk Source**	**Internal**	**Internal bin time base**
	Bin Width	**20.48 µs**	**20.48 µs bins**
	Bins/Record	**1k**	**1024 bins/record**
	Records/Scan	**100**	**Accumulate 100 records**
	Trigger Offset	**0**	**Start data at bin #0**
	Accumulate Mode	**Add**	**Add all records to accumulation**

Transient Digitizers

TD250/TD500 Transient Digitizers



- ***Up To 500 MSample/s Sample Rates***
- ***8-bit Resolution***
- ***Single or Dual Inputs***
- ***Up to 40 dB Linear Input Gain***
- ***Up to 1 Msample On-board Memory***
- ***Available with or without Enclosure and Power Supply***
- ***Logarithmic and Linear Amplifiers***
- ***Low Trigger Jitter (<2ns)***
- ***Up To 1.3Mbyte/s Data Transfer via GPIB***

TD250/TD500 Overview

The SRS TD series transient digitizers are low-cost digitizers optimized for time-of-flight and other spectroscopy applications but well suited to a variety of digitzer applications. The SRS TD digitizers are an excellent low-cost alternative to Digital Storage Oscilloscopes (DSO's) in many applications.

The core of the TD250 is a 250 MHz, 8 bit analog to digital converter (ADC) with 512 Kbytes of high speed memory providing a 2 ms total record length. The TD500 adds another digitizer and another 512K of memory providing 500 MSample/s digitization of a single input or 250 MS/s digitization of two channels when the optional second input channel is included.

Both the TD250 and TD500 are available as free standing printed circuit boards requiring external power, or in a rack-mountable enclosure with power supply. The instrument is controlled via a GPIB connection to the host computer.

Front End

The front ends of the TD Series contain software configurable variable gain amplifiers. Optimized for spectroscopic applications, the input signals are treated as unipolar, eliminating the need for a sign bit. The on-board attenuator can be user specified to match detector output. 16-bit DACs allow convenient input offset correction via software. Spark gaps and clamp diodes protect valuable components from high voltages.

In the Linear mode, 9 software selectable gain settings are available: 0 to 30 dB in 3.3 dB steps. The logarithmic amplifier, which covers 3 orders of magnitude of input range, provides much better performance when wide dynamic range is required and eliminates the need for separate measurements with high and low gain. This can greatly improve the throughput of your experiement and elimate the need to waste time searching for the optimum gain setting.

Low Trigger Jitter

The trigger circuit in the TD Series uses clock picking to achieve 2 ns jitter from the free running 4 ns clock. There is no insertion delay, so the first point is acquired at the trigger. The trigger channel is software controlled for both slope (rising or falling edge) and level (+2V to -2V) .

System Integration

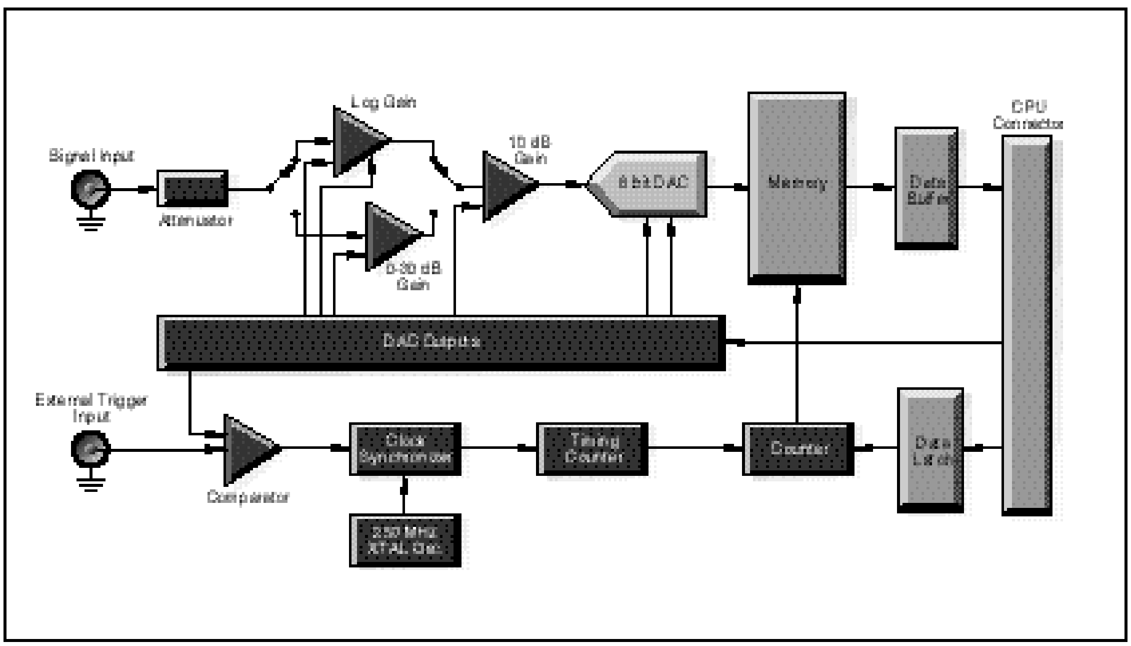
Since the TD Series Digitizers do not require mounting inside a PC they can be placed at the optimum position next to the detector, minimizing cable noise pickup. This also eliminates bulkhead BNC connectors and other signal-compromising components. The standard GPIB interface to the host computer assures compatibility with virtually any host computer system.

Fast DMA Transfer

The TD Series supports the high speed GPIB IEEE-488.2 transfer mode, which sends data to the host computer at speeds up to 1.3 Mbytes per second. In bench tests, the instruments have been clocked on a P75 running Windows 95 at over 750 Kbytes per second. This means that full length (1/2 megabyte) transients can be transferred in less than 2 seconds. The software allows you to transfer any segment of the data memory.

Easy Programmability

The TD Series is controlled through a high level text based command language. There are no jumpers or switches to set. The user can use any language and operating system for control and data acquisition.



TD Series Block Diagram

TD Series Rear Panel (With Optional Enclosure)

TD Series Specifications

Interface	All board functions and data transfer are via IEEE-488 (GPIB) interface. Ahost platform with an IEEE-488 controller is required. The CPU also provides an RS-232 interface.
Data Transfer Rate	Host dependent. Transfer rates as high as 600 ksamples/sec can be achieved (via GPIB).
Power	External power supply required. -5.2, +5.2 V @ 2A
Size	10.5"L x 7.75"W x 2.0"H

Digitizer

Input Range	0 to -2.0 V
Resolution	8 bits
Accuracy and Linearity	<1 bit
Sample Rate	250 MSample/sec
Aperture Jitter	<10 ps
SNR	42 dB for 100 MHz full scale signal (10 dB linear gain setting).
Memory Depth	512 k samples (software configurable with 4-sample resolution)

Input

Input Impedance	50 Ω
Input Attenuation	-0.78 dB*
Input Range	0 to -0.7 V (diode clamped) Signal MUST be negative.
Input Protection	Plasma breakdown device and diode clamping.
Linear Gains	10, 14, 17, 20, 24, 27, 30, 34, 37, 40 dB. Full scale input is (-2.0V/gain)x(input attenuation). Input MUST be negative.
Logarithmic Gain	300 mV full scale input. 0.23 dB/bit resolution and (2.0 dB log conformance.
Bandwidth	200 MHz

Trigger

Triggering	External trigger only. Data is all post-trigger.
Sample Uncertainty	Sample clock starts within 2 ns of trigger.
Input Impedance	10 kΩ*
Input Attenuation	0 dB*
Input Range	5.0 V (diode clamped)
Trigger Level	2.5 V (20 mV resolution)
Trigger Polarity	Rising or falling edge.
Pulse Width	> 2 ns

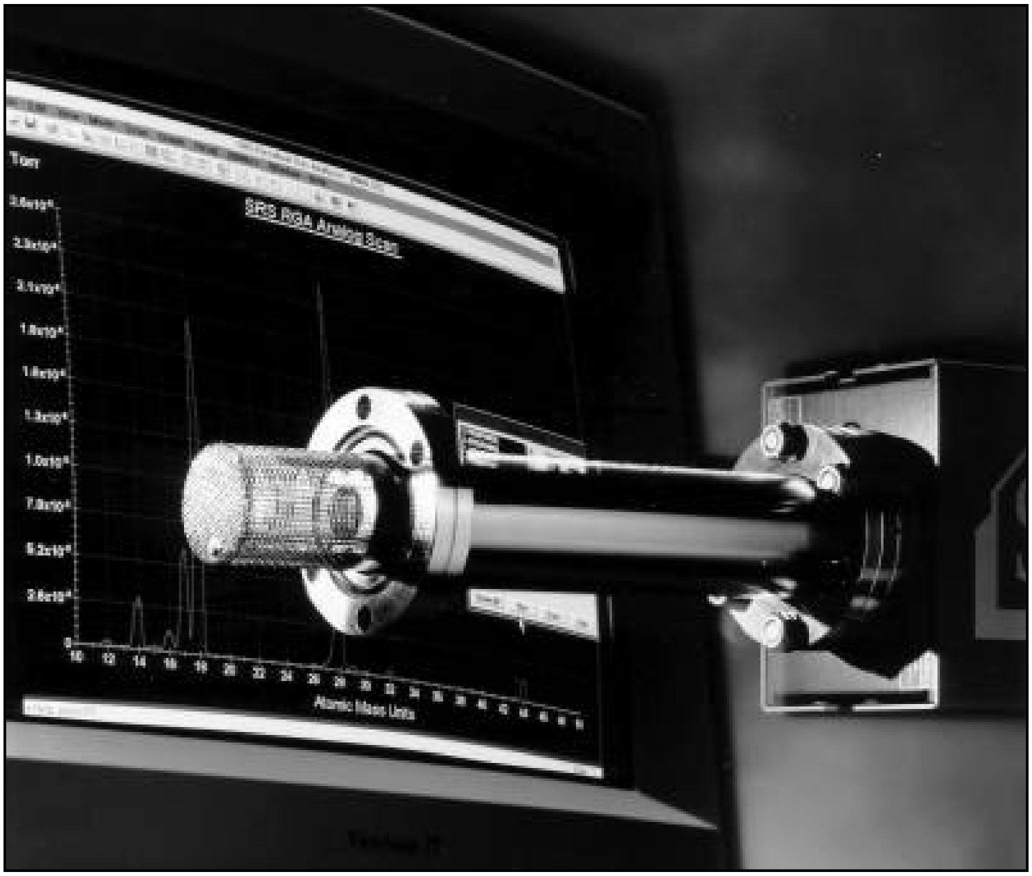
*May be changed by the user in hardware.

Ordering Information

TD250	**250 MS/s Digitizer**	**$ 2000**
TD500	**500 MS/s Digitizer**	**$ 3000**
Option 01	**Second Input for TD**	**$ 250**
Option 02	**Enclosure and Power Supply**	**$ 150**

Residual Gas Analyzers

RGA 100/200/300 Residual Gas Analyzers



- ***100, 200, and 300 amu Systems***
- ***Better than 1 amu Resolution***
- ***6 Orders of Magnitude Dynamic Range***
- ***Detectable Partial Pressures to 10^{-14} Torr***
- ***Real-time RGA Windows® Software***
- ***Mass Spectra, Leak Detection, and Pressure vs. Time Modes***
- ***Multi-head Operation***
- ***Field Replaceable Electron Multiplier and Filament***
- ***Rackmount Computer System Available***

RGA100/200/300 Overview

The RGA100/200/300 residual gas analyzers from Stanford Research Systems offer exceptional performance and value. These RGA's provide detailed gas analysis of vacuum systems at about half the price of competitive models. Each RGA system comes complete with a quadrupole probe, electronics control unit (ECU) and a real-time Windows® software package that is used for data acquisition and analysis as well as probe control.

Flexible Operation

Each RGA probe is programmable through its RS-232 port with a simple set of ASCII commands. You can program the RGA probe directly, or use Stanford Research Systems extensive Windows® software package that is included with both RGA models. A turn–key system that includes a rack-mountable computer system is also available.

The Competitive Choice

Whether your application involves vacuum diagnostics, process control, or sampling of high pressure gasses, the RGA100/200/300's combination of features, high performance and low cost makes it the obvious choice.

RGA100/200/300 Features

Rugged Probe Design

The RGA's probe consists of an ionizer, quadrupole mass filter and a detector. The simple design has a small number of parts which minimizes outgassing and reduces the chances of introducing impurities into your vacuum system. The probe assembly is rugged and mounts onto a standard 2 3/4 inch CF flange. It is covered with a stainless steel tube with the exception of the ionizer which requires just 2 1/2 inches of clearance in your vacuum system – about the same as a standard ion gauge. The probe is designed using self-aligning parts so it can easily be reassembled after cleaning.

Compact Electronics Control Unit

The densely packed electronics control unit (ECU) contains all the necessary electronics for controlling the RGA head. It is powered by either an external +24 VDC (2.5 A) power supply or an optional, built-in power module which plugs into an AC outlet. LED indicators provide instant feedback on the status of the electron multiplier, filament, electronics system and the probe. The ECU can easily be removed from the probe for high temperature bakeouts.

Unique Filament Design

A long-life dual Thoriated-Iridium (ThO2Ir) filament is used for electron emission. Dual thoriated filaments last much longer than single filaments, maximizing the time between filament replacement. Unlike other designs, RGA100/RGA200 filaments can be replaced by the user in a matter of minutes.

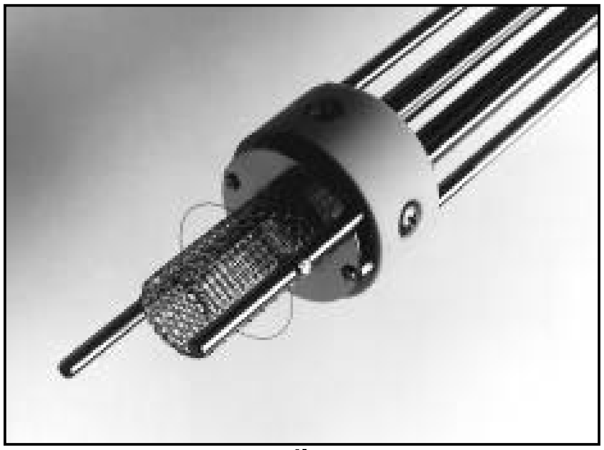
Continuous Dynode Electron Multiplier

A Faraday cup detector is standard with both 100 and 200 amu systems allowing partial pressure measurements from 10^{-4} to 5×10^{-11} torr. For increased sensitivity and faster scan rates an optional electron multiplier is offered that detects partial pressures down to 5×10^{-14} torr. This state-of-the-art macro multi-channel continuous dynode electron multiplier (CDEM) offers increased longevity and stability and can also be installed by the user – a first for RGAs.

Useful features

Both RGAs have a built-in degassing feature. Using electron impact desorption, the ion source is thoroughly cleaned, greatly reducing the ionizer's contribution to background noise.

A firmware-driven filament protection feature constantly monitors (675 Hz) for over pressure. If over pressure is



RGA Filament

detected the filament is immediately shut off, preserving its life.

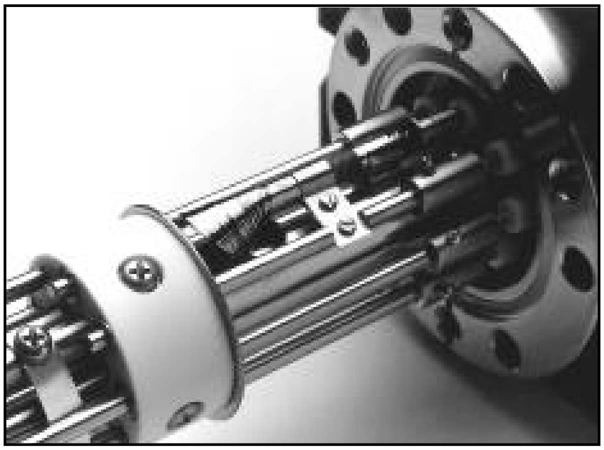
A unique temperature compensated logarithmic electrometer detects ion current from 10^{-7} to 10^{-15} amps in a single scan with better than 2% precision. This huge dynamic range means you can make measurements of small and large gas concentrations simultaneously.

Complete Programmability

Communication with computers is made via the RS-232 interface. Analog and histogram (bar) scans, leak detection, total pressure, and probe parameters are all controlled and monitored through a high level command set. This allows easy integration into preexisting processing programs.

RGA Windows Software

The RGA systems are supported with a real-time Windows® software package that runs on IBM compatible



Continuous Dynode Electron Multiplier

PCs (486 or greater). The intuitive graphical user interface allows measurements to be made quickly and easily. The program is fully interactive giving the user complete control of the graphical display. Screens can be split for dual mode operation, scales can be set to linear or log format, and data can be scaled manually or automatically. Data is captured and displayed in real-time or scheduled for acquisition at a given time interval for long-term data logging. Features include user selectable units (Torr, mBar, Pascals and Amps), programmable audio and visual alarms, and comprehensive, on-line help.

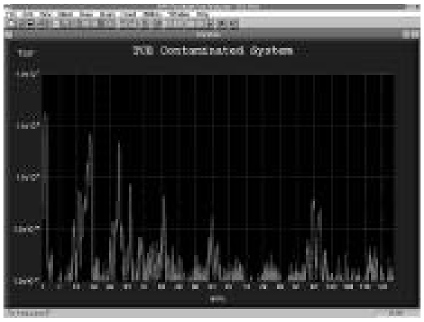
The software also allows complete RGA head control with easy mass scale tuning, sensitivity calibration, ionizer setup and electron multiplier gain adjustment. For further analysis, data files can be saved in ASCII format for easy transfer into spreadsheets. Graphic images can be saved as META files or copied to the clipboard for importing directly into other Windows® programs. The software also provides password protection for locking out head parameters so that casual users can't alter important settings.

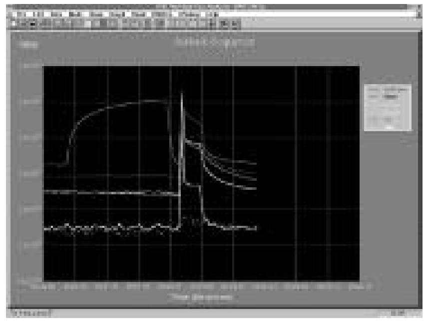
Multiple-head Operation

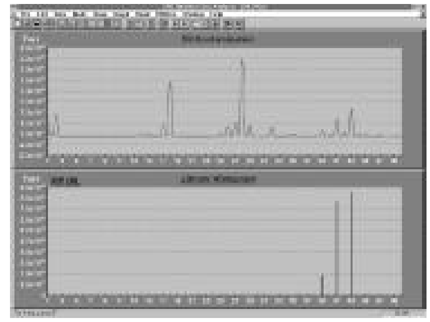
The software supports multiple head operation when more than one RGA is needed. Up to eight ECUs can be monitored from the software.

Computer Rackmount Enclosure

RGA Rear Panel with and without Option O2 Power Module



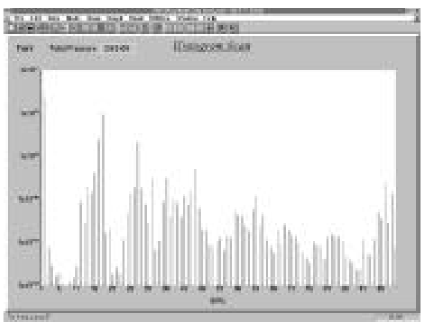


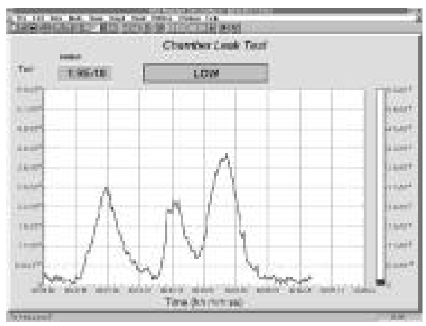


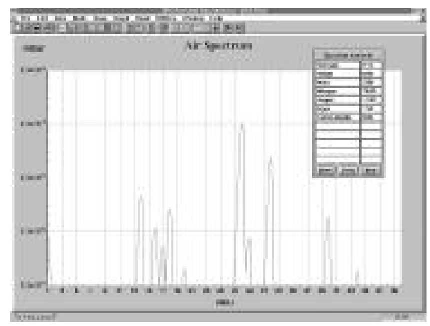
Analog mode *provides a line graph representation of the acquired mass spectrum (partial pressure vs. mass number). Span, resolution and noise floor can each be set. Scans can be single-shot, timed or taken continuously. Total pressure is available in analog and most other modes of operation.*

Pressure vs. time *presents a strip chart of partial pressures for selected masses and provides a complete time history of your data. Complete scrolling and zoom control is available even while data is being acquired. This mode is most often used for monitoring process trends.*

Library mode *contains a comprehensive list of gases that can be used to compare against the current spectrum. A search mode allows you to select up to 12 masses and identify and display (numerically and graphically) the intensity of all gases that contain these masses.*



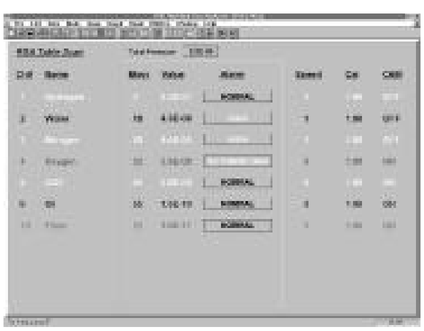


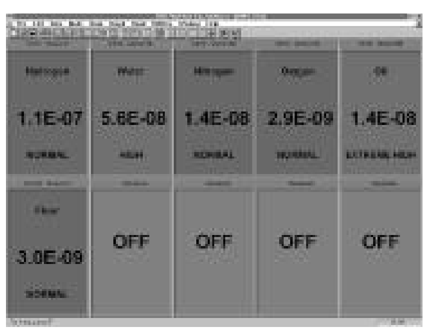


Histogram mode *displays a bar graph of partial pressure vs. mass allowing the spectrum to easily be interpreted. This mode is often used for quick and easy vacuum analysis. The screen can be split for viewing two modes of operation simultaneously.*

Leak detection mode *monitors a particular mass number (not just Helium) over time, and combines many features of the previous modes. A vertical bar graph provides a visual reference for viewing changes in intensity from a distance. A programmable audible tone, large numeric read-out, and visual alarm are also provided.*

Analysis utility *provides an approximation of the composition of gases being monitored by the RGA. Since more than one gas can contribute to a particular amu's partial pressure, the analysis mode is extremely useful in determining the make-up of complex gases. Up to 12 common gases can be selected for the analysis.*





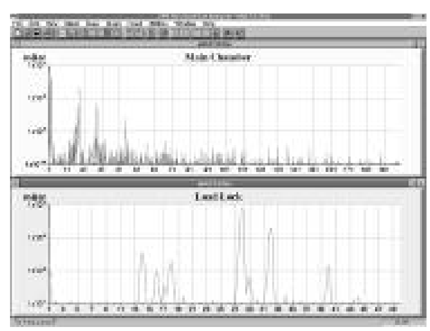


Table mode *provides a readout of mass, scaling factor and true partial pressure. The display shows the peak heights and alarm status of up to 10 masses. The electron multiplier can be independently set on or off for each mass. This allows the user to view minor species even in the presence of high total pressure.*

Annunciator mode *is provided for conveniently monitoring up to 10 masses. If a particular mass has tripped its preset alarm, the large box will turn red indicating a problem. An audible alarm will also be present until the mass falls back within its preset limits. This mode is most often used for Go/No-Go testing.*

Multi-head operation *is available when more than one RGA is needed for analysis. Up to eight heads can be monitored simultaneously from the software.*

SRS Tech Note

About RGA Dynamic Range

Dynamic range is defined as the ratio between the smallest signal that can accurately be measured and a full scale signal. Residual gas analyzers typically offer 3 to 4 orders of magnitude dynamic range. SRS RGAs use a logarithmic amplifier in the detector to achieve more than 6 orders of magnitude dynamic range. Figure 1 at the right shows a mass spectrum of 99.999% nitrogen, which was measured using the Faraday cup (FC) detector. The partial pressure scale covers seven decades (10^{-4} to 10^{-11} Torr) and data is acquired in a single scan without range changes. The effect of outgassing of the vacuum chamber has been removed using the background subtraction feature of our software. The RGA has sufficient resolution to detect $^{15}N_2$, which is naturally present at 15 ppm, even though it is only 2 amu away from a peak that is five decades larger.

Figure 1 shows a high pressure of 4 x 10^{-5} torr ($^{14}N_2$), which is close to the saturation limit of the detector (1 x 10^{-4} Torr), and a noise floor of about 1 x 10^{-10} Torr. From these values we can determine that the detection limit of the RGA is a few ppm. Switching to the electron multiplier detector (CDEM) we can improve the signal detection limit of the RGA. Using the software's table mode (see figure 2), the RGA can be set for FC or CDEM detection for up to 10 masses. The Faraday cup is used for major species and the electron multiplier for minor species. Note the dynamic range has been significantly improved (8 orders of magnitude) with the noise floor now at about 10^{-13} Torr.

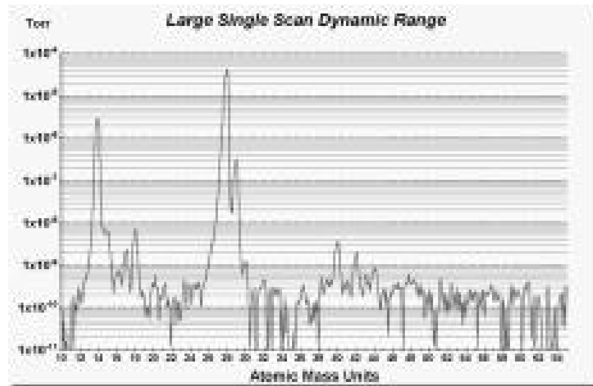


FIGURE 1

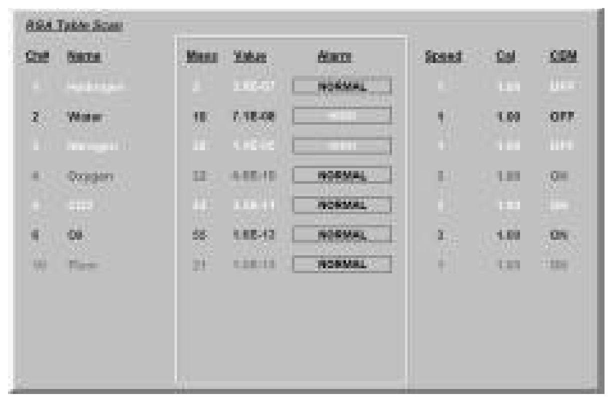
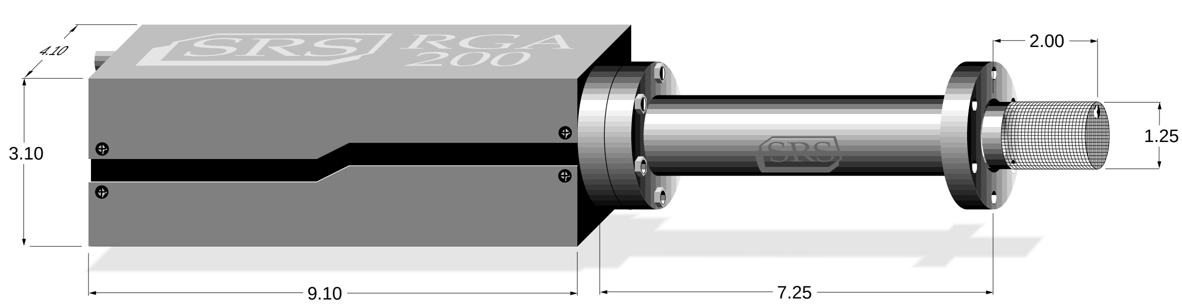


FIGURE 2

RGA Dimentional Drawing



Operational Specifications

Mass Range:	
RGA100	1 to 100 amu
RGA200	1 to 200 amu
RGA300	1 to 300 amu
Mass filter type	Quadrupole
Detector type:	
Faraday cup (FC)	standard
Electron multiplier (CEM)	optional
Resolution (per AVS std. 2.3)	Greater than 0.5 amu @ 10% peak height. Adjustable to constant peak width throughout the mass range.
Sensitivity (A/Torr)	$2x10^{-4}$ (FC), <200 (CEM).†(User adjustable throughout high voltage range.)
Minimum detectable partial pressure	$5x10^{-11}$ Torr (FC).† $5x10^{-14}$ Torr (CEM).†
Operating pressure range	10^{-4} Torr to UHV (FC) 10^{-6} Torr to UHV (CEM)
Bakeout temperature (without ECU)	350° C (FC) 300° C (CEM)
Total press. meas.	Always available

Note† Measured with N_2 @ 28 amu with 1 amu full peak width, 10% height, 70 eV electron energy, 12 eV ion energy and 1 mA electron emission current.

Ionizer

Design	Open ion source. cylindrical symmetry, electron impact ionization.
Material	SS304 construction
Filament	Thoriated Iridium (dual) with firmware protection. Built-in 1 to 10 W degas ramp-up. Field replaceable.
Electron energy	25 to 105 V, programmable.
Ion energy	8 or 12 V, programmable.
Focus voltage	0 to 150 V, programmable.
Electron emission current	0 to 3.5 mA, programmable.

Application Information

Application Note #7, “Vacuum Diagnosis With RGA’s” and #8, “Using RGA’s to Sample High Pressure Gasses” contain useful information on how RGA’s can be incorporated into practical vacuum systems.

General

Probe dimension	8.75" from flange face to top of ionizer
Probe insertion	2.0"
Probe mounting flange	2.75" CF
Minimum tube I.D.	1.375"
ECU dimensions	9.1" x 4.1"x 3.1". Easily separated from the probe for bakeout.
LED indicators	Power ON/OFF, filament ON/OFF, degas ON/OFF, Elec. mult. ON/OFF, RS-232 Busy, Error, Overpressure, Burnt Filament.
Warm-up time	Mass stability ±0.1 amu after 30 minutes.
Computer interface	RS-232C, 28,800 Baud with high level command set.
Software	Windows® based application. Requires 486 or better.
Power Requirement	24 VDC @ 2.5 Amps. Male DB9 connector. Optional 110/120/220/240 VAC (50/60 Hz) adapter.
Weight	6 lbs.
Warranty	One year parts and labor on materials and workmanship.

Ordering Information

RGA100	**100 amu system with RGA Windows® software**	**$3750**
RGA200	**200 amu system with RGA Windows® software**	**$4500**
RGA300	**300 amu system with RGA Windows® software**	**$6000**
Opt 01	**Electron Multiplier (with HVPS)**	**$1500**
Opt 02	**Built-in power module (For AC Line Operation)**	**$250**
Opt 03	**Electron Multiplier Ion Counting Output**	**250**
Opt 04	**Max insertion nipple**	**$400**
O100HJR	**200°C Heater Jacket**	**$395**
O100RF	**Replacement filament**	**$200**
O100EM	**Replacement Electron Multiplier**	**$1000**
O100RI	**Replacement Ionizer**	**$450**
O100TR	**Computer Output Card Cable, Relay Card**	**$595**
O100TS	**Computer Output Card Cable, TTL Screw Terminal Card**	**$290**
O100RM	**Rack mount enclosure (19") (for computer and monitor)**	**$500**
O100CS	**Computer System: Pentium w/ HD, CD Rom, 15" VGA Monitor, Windows 95®**	**$2000**

Closed Ion Source Gas Analyzers

Model CIS100/200/300— Closed Ion Source Gas Analyzers

- ***100, 200, and 300 amu Systems***
- ***1 ppm Detection with Electron Multiplier***
- ***1 mTorr Operation***
- ***Real-time RGA Windows Software***
- ***Multi-head Operation***
- ***Field Replaceable Electron Multiplier and Filament***
- ***RGA and CIS Mode***

CIS100/200/300 Overview

The new Closed Ion Source Gas Analyzers (100, 200 and 300 amu mass range) extend the versatility of Stanford Research Systems line of Gas Analyzers. With better than 1 ppm detection limit, direct sampling at mTorr pressure and a user friendly real-time Windows software package, the CIS systems will satisfy your most demanding application requirements. On-line process monitoring and control, verification of process gas purity at the point of use, high vacuum residual gas analysis, and process equipment leak checking are some of the areas where these systems will prove indispensable.

Why Use a CIS ?

Unlike open ion source RGAs, the CIS system can be used to directly sample process gases at up to 10 mTorr pressures, and provide a 1 ppm or better detection limit for commonly monitored background gases like water, nitrogen etc. Higher sample pressure also leads to better sensitivity and faster response times. Background gases inside the probe are maintained at levels similar to that of the open ion source configuration. This translates into an additional dynamic range of two orders of magnitude. The Tech Note at the end of this section explains more details of the Closed Ion Source.

Compact Design

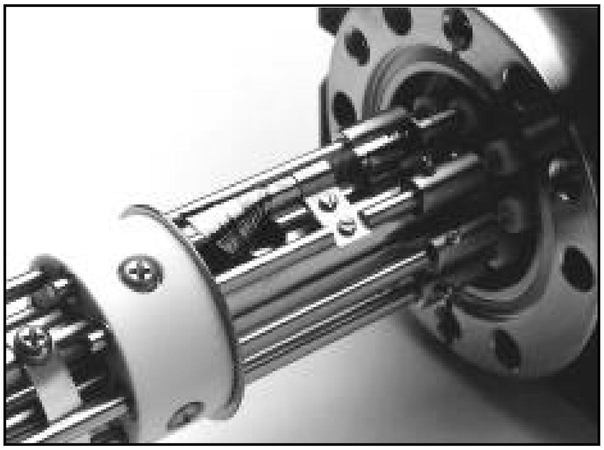
The probe consists of a quadrupole mass spectrometer with a CIS ionizer mounted inside a 2.75” Conflat Tee (CIS Cover Tee). The control unit mounts directly on the probe’s feedthru flange and contains all the necessary electronics for operating the instrument. The side port of the CIS Cover Tee provides a connection for the differential pumping system that keeps the quadrupole mass analyzer and the filament at high vacuum. The system can be connected directly to a process chamber through its standard CIS Mounting Flange (2.75” CF connection). The unit is self-aligning, with a simple, robust design. Cleaning and reassembly of the probe, and replacement of the filament and electron multiplier can be performed in the field, with no need to return the unit.

Gold plated ionizer

The entire ionizer is made of gold plated stainless steel. This reduces outgassing and background levels in the ionization region, improves the long-term stability and permits operation while exposed to reactive and corrosive gases. A Tungsten filament is used, which resists corrosive and reactive gases such as WF_6 and silane, and leads to extended lifetime. In addition, the probe is gas tight (there is a pressure difference between the process chamber and the probe), which means that commonly interfering species are prevented from contaminating the ionizer, and the spectra obtained while scanning are free of overlaps.

Versatility

The CIS systems can also be used in a so-called “RGA mode”. In this mode, the unit has a lower minimum



Continuous Dynode Electron Multiplier

detectable partial pressure, but a lower maximum operating pressure as well. The RGA mode is used, for example, in the first stage of a sputtering process when the chamber is evacuated to a low pressure and the quality of the vacuum is checked for leaks and harmful contaminants. The unit can then be switched to a CIS mode, for sampling directly at higher pressure, for a better signal to noise ratio.

Choice of Detectors

The CIS series analyzers come standard with a Faraday cup detector which provides between 1 ppm and 10 ppm detection. To guarantee detectibility below 1 ppm the optional Continuous Dynode Electron Multiplier must be included.

Complete programmablity

A standard RS-232 interface is provided along with a complete programming reference. All probe parameters can be controlled and monitored, and data can be acquired for use in pre-existing applications.

Windows software

The CIS systems are supported with a real-time Windows® software package that runs on IBM compatible PCs (486 or greater). The intuitive graphical user interface allows measurments to be made quickly and easily. The program is fully interactive giving the user complete control of the graphical display. Screens can be split for dual mode operation, scales can be set to linear or log format, and data can be scaled manually or automatically. Data is captured and displayed in real-time or scheduled for acquisition at a given interval for long term data logging. Features include user selectable units (Torr, mBar, Pascals, Amps, ppm, percent), programmable audio and visual alarms, and comprehensive, on-line help. The software also allows complete CIS head control with easy mass scale tuning, sensitivity calibration, ionizer setup and electron multiplier gain adjustment. For further analysis, data files can be saved in ASCII format for easy transfer into spreadsheets. Graphic images can be saved as META files or copied to the clipboard for importing into other Windows® programs. The software also provides password protection for locking out head parameters so that casual users can't alter important settings.The software supports multiple head operation when more than one CIS is needed. Up to eight ECUs can be monitored by the software.

Pumping Requirements

The CIS100/200/300 requires connection to a pumping system with a pumping speed of at least 70 L/s and a base pressure of 10^{-10} torr. The connection port is a 2.75" Conflat flange. Option O100TDP provides a turbo pump which mounts directly to the CIS head along with a diaphragm roughing pump. When this option is ordered the entire analyzer including the quadrapole spectrometer and pumps are assembled, tested, and calibrated at the factory. Users can provide their own pumping station, however it is the user's repsonsibility to ensure that the pumping system does not damage or limit the performance of the instrument.

SRS Tech Note

Closed vs. Open Ion Source Gas Analyzers

The SRS Residual Gas Analyzer (RGA) uses an Open Ion Source (OIS) configuration for the ionization of the gas molecules. The OIS penetrates into the process chamber and is "open" to all the gaseous molecules in the rarefied vacuum environment. The pressure in the ionizer is the same as in the rest of the surrounding vacuum and also the same as in the quadrupole mass filter and ion detector. The upper pressure limit for the operation of an OIS gas analyzer is 10^{-4} Torr; however, the pressure range can be shifted to higher levels (i.e. 10^{-3} to 10 Torr) with the help of a differentially-pumped Partial Pressure Reduction gas inlet system (PPR) consisting of a restriction and a vacuum pump package. OIS gas analyzers have adequate sensitivity and dynamic range to detect part-per-million (PPM) level contaminants in principle; however, interferences from process gases and background interferences from the sensor itself (i.e. outgassing from the quadrupole and detector assembly) can make the detection of PPM levels of some common residual impurities, such as water, difficult in practice.

In applications requiring the measurement of pressures between 10^{-4} and 10^{-2} Torr, the problem of background and process gas interferences to the mass spectra can be significantly reduced by replacing the traditional OIS PPR configuration described above with a Closed Ion Source (CIS) gas sampling system. Diagrams of the two systems are shown on the next page.

The CIS Ionizer sits on top of the quadrupole mass filter replacing the more traditional OIS used in conventional RGAs. It consists of a short, gas-tight tube with two very small openings for the entrance of electrons and the exit of ions. Ions are produced by electron impact directly at the process pressure (i.e. mTorr range) while, at the same time, a pumping system keeps the filament and the rest of the quadrupole assembly at pressures below 10^{-5} Torr through differential pumping (i.e. two decades of pressure reduction).

Because the sampling pressure in the CIS is typically two decades higher than that of the rest of the sensor's vacuum system, the signal-to-background ratio is significantly increased relative to OIS PPR configuration. This is particularly important when measuring common residual gases, such as water

In order to best illustrate the difference in performance between the two ionization configurations, we use as an example the measurement of water impurity levels in a 10^{-3} Torr Ar sputtering process using both an OIS and a CIS gas analyzer.

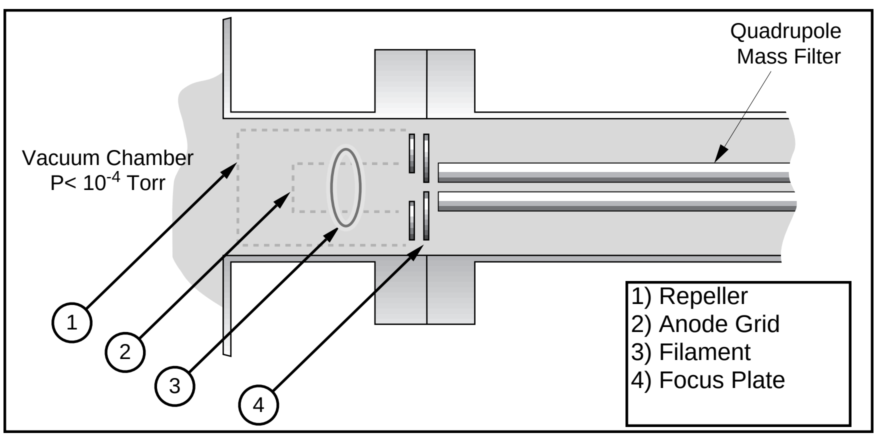
An OIS cannot be exposed directly to 1 mTorr of process pressure. A PPR must be used to step the total pressure down to about 10^{-5} Torr at the ionizer, corresponding to a two-decade pressure reduction factor. The pressure drop, brings a 1 PPM level of water in the process chamber (corresponding to a 10^{-9} Torr partial pressure of the impurity) to a partial pressure level in the mass spectrometer of about 10^{-11} Torr, well within the detection limit of a typical RGA. However, with the mass spectrometer isolated from the process gases, the residual pressure in the quadrupole chamber is, at best, in the order of 10^{-9} Torr with most of that being water. This water background level is one hundred times larger than the 10^{-11} Torr corresponding to the one PPM of water coming from the process chamber. The obvious conclusion is that the water vapor concentration in the process gas cannot be reliably detected or measured to better than 100 PPM under these "common" operating conditions.

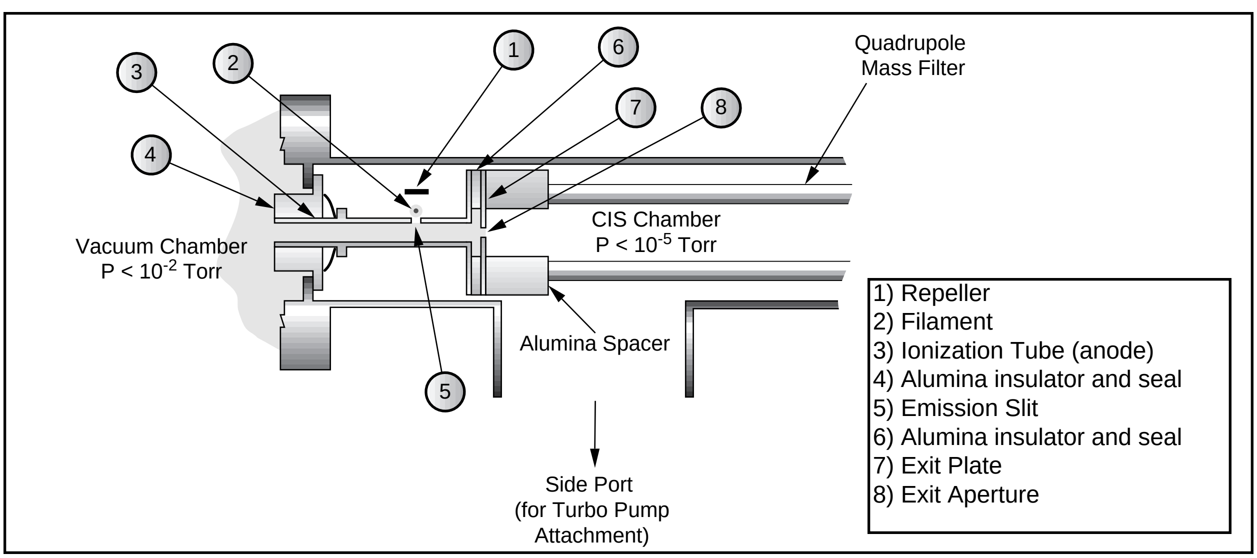
Even though the RGA is intrinsically capable of performing sub-PPM measurements, it is not always easy to find places in the residual mass spectrum of the RGA where the background is readily in the PPM levels.

In the CIS system, the Ar gas is ionized directly at 10^{-3} Torr without any intermediate pressure reduction. At the same time, the background water signal is the same or less than in the OIS configuration, since it is due to water molecules backstreaming into the CIS ionizer from the quadrupole assembly (10^{-9} Torr partial pressure level). This background water signal now corresponds to a 1 PPM water level in the process gas. The obvious conclusion is that the water vapor concentration in the process gas can now be reliably detected and measured to better than one PPM levels: A two decade improvement over the OIS water detectability limit.

The combination of direct sampling and differential pumping provides the potential for PPM and sub-PPM detection limits for even the most pervasive residual gases. For other common interferences, such as organic contaminants or reaction by-products of the filament, the gas tight design of the source reduces the visibility of the ionization region to those gases providing a very clean residual gas spectrum, free of many of the spectral overlaps that are common in OIS PPR setups.

The ability of the CIS Quadrupole Gas Analyzer to sample gases directly in the mTorr range and to provide PPM level detectability across its entire mass range has made the the instrument the detector of choice in semiconductor processing applications such as PVD, CVD and etching.





Operational

Mass range	
CIS100	1 to 100 amu
CIS200	1 to 200 amu
CIS300	1 to 300 amu
Mass filter type	Quadrupole
Detector type	Faraday cup (FC) – standard Electron multiplier (CEM) – optional
Resolution (per AVS std. 2.3)	Better than 0.5 amu @ 10% peak height. Adjustable to constant peak width throughout the mass range.
Max. operating temp.	200°C (FC), 100°C (CEM)
Bakeout temperature	350°C (without ECU)

Ionizer

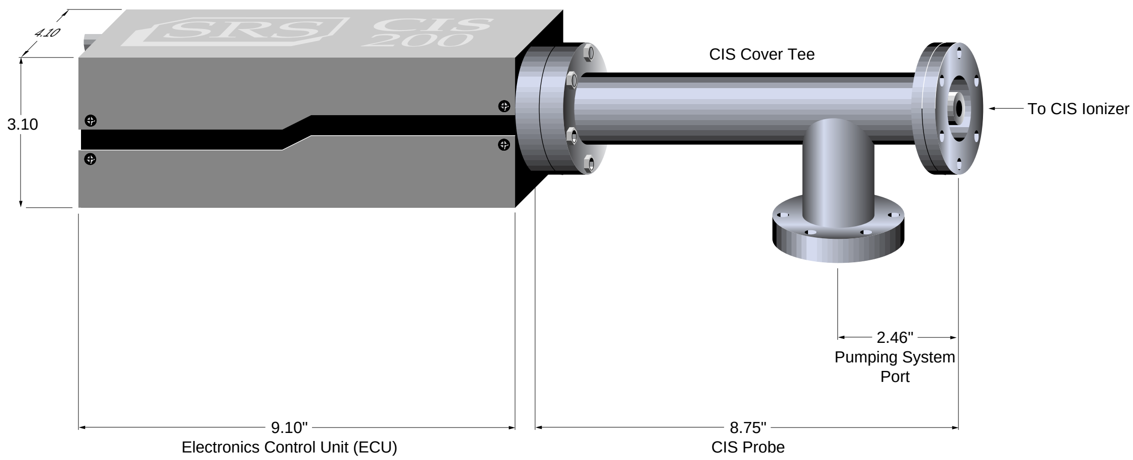
Design	Closed ion source. cylindrical symmetry, electron impact ionization.
Material	SS304, Gold plated
Filament	Tungsten with firmware protection. Built-in 1 to 10 W degas ramp-up. Field replace-able. Optional: Thoriated Iridium filament.
Electron energy	25 to 105 V, programmable.
Ion energy	8 or 12 V, programmable.
Focus voltage	0 to 150 V, programmable.
Electron emission current	0 to1.0 mA, programmable.

General

Probe dimensions	see next page
Probe mounting flange	2.75" CF
ECU dimensions	9.1" x 4.1"x 3.1". Easily separated from the probe for bakeout.
LED indicators	Power ON/OFF, filament ON/OFF, degas ON/OFF, Elec. mult. ON/OFF, RS-232 Busy, Error, Overpressure, Burnt Fila-ment.
Warm-up time	Peak height ±2% after 3 minutes. Mass stability ±0.1 amu after 30 minutes.
Computer interface	RS-232C, 28,800 Baud with high level command set.
Software	Windows® based application. Requires 486 or better.
Power Requirement	24 VDC @ 2.5 Amps. Male DB9 connec-tor. Optional 120 VAC adapter.
Weight	6 lbs.
Warranty	One year parts and labor on materials and workmanship.

CIS Parameter	Programmable range	RGA mode	CIS mode
Electron emission current (mA)	0 to 1	0.5	0.05
Electron energy (eV)	25 to 105	70 eV	70 or 35 eV
Ion energy (eV)	4 or 8	4 or 8	4 or 8
Extraction Voltage (V)	0 to -150	-50 typical	-50 typical
Sensitivity (for N_2 @ 28 amu)		10^{-5} A/Torr	10^{-6} A/Torr
Linear range upper limit (Torr)		10^{-4}	2×10^{-3}
MDPP (Torr)		10^{-9}	10^{-8}

Notes:
1. The CIS tests were performed by attaching a 70 l / s hybrid turbomolecular pump, backed up by a high performance diaphragm pump to the extra port in the CF Tee of the CIS probe.
2. The RGA mode sensitivities reported were calculated for N_2 at $<10^{-5}$ Torr.
3. The CIS mode sensitivities were calculated for N_2 at 1-5 x 10^{-4} Torr.
4. MDPP (Minimum Detectable Partial Pressure) is determined by measuring baseline levels for FC detection in the presence of $^{28}N_2$ at 10^{-5} torr (RGA mode) and 10^{-3} Torr (CIS mode). Up to 3 orders of magnitude improvement in detectability is possible if the CDEM is turned on.



CIS Dimentional Drawing

Ordering Information

CIS100	**100 amu system with RGA Windows® software**	**$5750**
CIS200	**200 amu system with RGA Windows® software**	**$6500**
CIS300	**300 amu system with RGA Windows® software**	**$8000**
Opt 01	**Electron Multiplier (with HVPS)**	**$1500**
Opt 02	**Built-in power module (For AC Line Operation)**	**$250**
O100RFW	**Replacement Tungsten Filament (Qty. 5)**	**$100**
O100RFT	**Replacement ThO_2Ir Filament (Qty. 2)**	**$125**
O100HJC	**200°C Heater Jacket**	**$395**
O100EM	**Replacement Electron Multiplier**	**$1000**
O100RIC	**Replacement Ionizer kit (includes filament)**	**$900**
O100TDP	**Turbo pump and Diaphragm Pump**	**$9975**
O100TR	**Computer Output Card Cable, Relay Card**	**$595**
O100TS	**Computer Output Card Cable, TTL Screw Terminal Card**	**$290**
O100RM	**Rack mount enclosure (19") (for computer and monitor)**	**$500**
O100CS	**Computer System: Pentium, Hard Drive, CDROM,15” VGA Monitor, Windows 95®**	**$2000**

Process Monitoring Systems

PPR100/200/300 — Vacuum Process Monitoring Systems

- ***Inlet pressure selectable in decades from 10 mTorr to 10 Torr***
- ***Flow rate of 3 x 10^{-5} Torr l/s with pressure reduction inlet active***
- ***2 sec. response time at 0.1 Torr inlet pressure (scales linearly with pressure)***
- ***Quadrupole mass spectrometers with 100, 200 or 300 amu range***
- ***6 orders of magnitude dynamic range in a single scan***
- ***Windows® based software for fast data acquisition and control***

PPR100/200/300 Overview

The PPR Vacuum Process Monitoring System is designed for inline process monitoring and diagnosis. Two paths are provided to the RGA: a high conductivity path for monitoring base vacuum and a pressure reducing path for monitoring the process at operating pressure. The pressure reducing path contains a micro-hole orifice, which is designed to operate at one of the following pressures: 0.01, 0.1, 1, or 10 Torr. It reduces the sample pressure to the operating pressure of the RGA (about 10^{-6} Torr). This pressure drop is maintained by the pumping system, which consists of a hybrid turbomolecular pump and a diaphragm pump. Both pumps are oil free and will not contaminate your process. The inlet assembly that attaches to your process chamber is pictured above. The full system also includes a controller, diaphragm pump and a Windows® based software program for data acquisition and control.

The software is used to operate the instrument in various modes, including analog scan, histogram mode and pressure vs. time mode. An electron multiplier option provides greater sensitivity and higher scan speeds. The PPR system is shipped completely assembled and calibrated, and is ready to attach to your vacuum process chamber.

Performance

Gas flow	~ 3 x 10^{-5} mbar l /s with pressure reduction inlet active
Response time	2 s at 0.1 mbar inlet pressure (scales linearly with pressure)
Startup time	8 minutes nominal

Connections

Inlet	2.75 inch CF flange, rotatable with through holes
Inlet to controller	6 foot cable (provided)
Inlet to backing pump	6 foot, 1/4 inch ID x 7/16 inch OD flexible hose (provided)
Computer interface	RS-232C (28,800 baud, 9 pin D connector)

Pumps

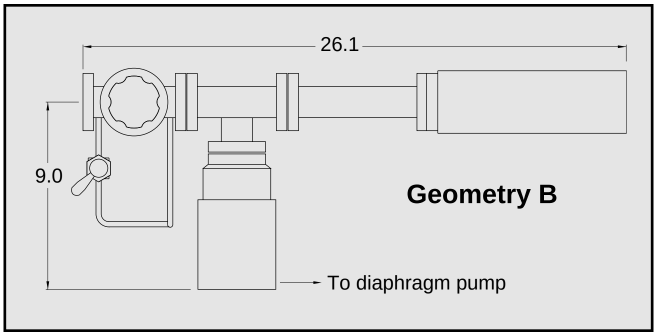
High Vacuum	Hybrid turbomolecular/drag pump, 70 liter/s, ultimate pressure 2 x 10^{-9} mbar
Backing	Diaphragm pump with ultimate pressure less than 1 mbar. Protection class IP44
Cooling	Requires forced air cooling

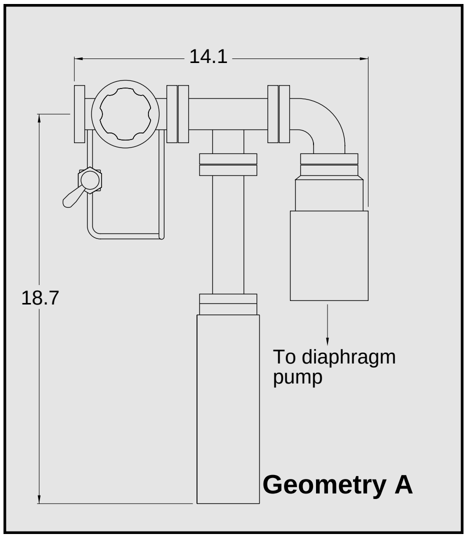
General

Power requirements	110 VAC/60 Hz or 220 VAC/50 Hz (not field selectable), less than 300 W
Dimensions	Vary with configuration (see sample configurations below)
Weight	Inlet (mounted on chamber) 7 kg. (16 lbs.)

Inlet Geometries

The inlet assembly is offered in various geometries to match your space requirements. Shown are two typical configurations. (dimensions in inches.) Contact SRS for complete mechanical drawings or descriptions of addional geometries





Ordering Information

PPR100	**100 amu PPR System**	**$ 17,500**
PPR200	**200 amu PPR System**	**$ 18,250**
PPR300	**300 amu PPR System**	**$ 19,750**
Option 01	**Electron Multiplier**	**$1500**
O100RO	**Replacement Orifice**	**$ 50**
O100HJR	**200°C Heater Jacket**	**$395**
O100RF	**Replacement filament**	**$200**
O100EM	**Replacement Electron Multiplier**	**$1000**
O100RI	**Replacement Ionizer**	**$450**
O100TR	**Computer Output Card Cable, Relay Card**	**$595**
O100TS	**Computer Output Card Cable, TTL Card**	**$290**
O100CS	**Computer System: Pentium, CDROM, 15” VGA Monitor, Windows ‘95**	**$ 2000**
O100RM	**19” Computer Rack Mount**	**$ 500**

- ***100, 200 or 300 amu systems***
- ***Pressures from 10 mbar to 1 bar***
- ***Response time of less than 1 second***
- ***Compact, transportable package***
- ***Better than 1 amu resolution***
- ***Six orders of magnitude dynamic range in a single scan***
- ***Real-time Windows® based gas analysis software***
- ***RS-232 computer interface***

QMS100/200/300 Overview

QMS Series gas analyzers offer efficient, cost effective solutions for gas analysis and monitoring in vacuum or atmospheric pressure processes. These mass spectrometers simplify the task of on-line process monitoring, leak detection, troubleshooting and analysis of gas species.

On-line monitoring

The QMS system continuously samples gas at low flow rates (several milliliters per minute) making the instrument ideal for on-line analysis. The inlet can be equipped to sample at pressures from above atmospheric to as low as 10 mbar. Data is acquired continuously, as opposed to batch sampling which is employed by gas chromatographs. The QMS system has a fast response time which means that a change in composition at the inlet can be detected in less than a half second. Complete spectra can be recorded in under one minute and individual masses can be measured at rates up to 4 points per second.

Compact, user friendly design

An advanced quadrupole mass spectrometer design, coupled with state -of-the-art pumping technology, allows the entire system to be packaged in a small, transportable system. The sampling inlet valves and pumps are controlled from the front panel and operation does not require a detailed understanding of the quadrupole or the principles of the pressure reducing inlet. The unit can be operated either standing upright or laid on its side. The pumps contain no oil or other liquids and therefore gravity has no effect on them. Capillaries are used as sampling probes and are available in a wide variety of materials and sizes for different applications.

Principle of operation

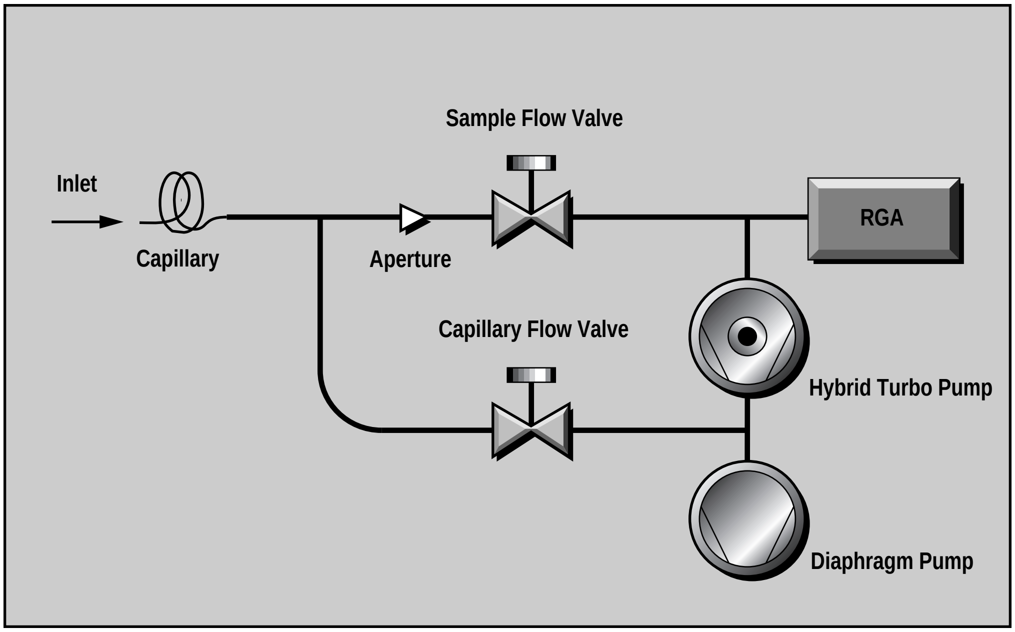
The QMS uses a two stage pressure reducing inlet to sample gases at high pressure. This allows the use of a residual gas analyzer (which operates at high vacuum) as the gas detector. Different capillaries allow the system to sample at higher or lower pressures. A large flow is drawn through the capillary which drops the pressure 3 decades. A hybrid turbomolecular/drag pump draws a small amount of gas through an aperture, which reduces the pressure to about 1e-6 Torr, while most of the flow is bypassed directly to a diaphragm pump. Solenoid valves control the gas flow. The entire system is controlled by a microcontroller, which ensures correct operation of the pumps and valves, while the software controls the spectrometer and acquires data in several different modes.

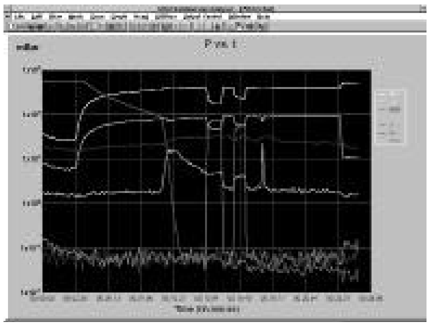
Bypass flow configuration

The bypass flow stream and the sample flow stream are recombined at the diaphragm pump, which makes it possible to use only one backing pump. This contributes to the system's small size. The bypass flow configuration is also critical to the fast response time of the QMS. Single stage pressure reduction with leak valves would lead to extremely small flow rates and unreasonably long sample times. The bypass flow on the other hand gives an exceptional response time of less than 0.2 seconds from the tip of the capillary to the RGA chamber.

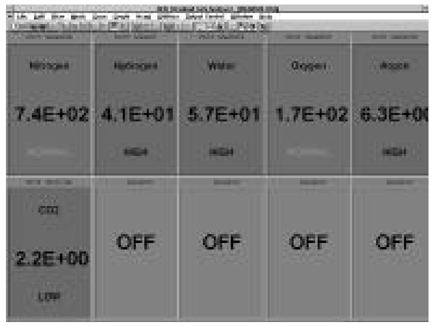
RGA Windows® software

The QMS systems are supported with a real-time Windows® software package that runs on IBM compatible PCs (486 or greater). The intuitive graphical user interface allows measurements to be made quickly and easily. Data is captured and displayed in real-time or scheduled for acquisition at specified time intervals. Features include analog and histogram modes, pressure vs. time scans, leak detection, audio and visual alarms, relay output options and comprehensive on-line help. Probe parameters can also be controlled and monitored through a high level command set which allows easy integration into pre-existing processing programs. A standard RS-232 interface is used as the data link to the host computer.

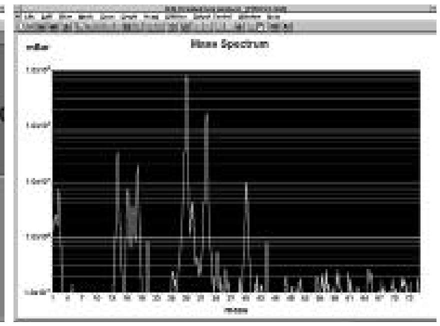




__Pressure vs. time mode__ presents a strip chart display of partial pressures for selected gases and provides a complete history of data.



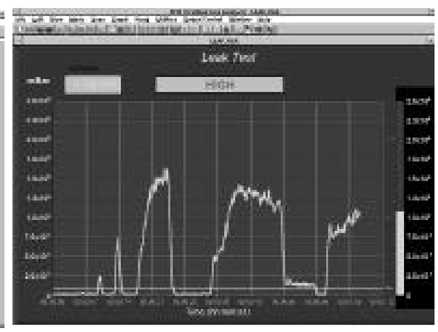
__Annunciator mode__ monitors up to 10 gases. User set limits allow Go/No-Go testing in conjunction with audible alarms and/or an optional relay board.



__Analog mode__ presents the entire spectrum and displays the results as partial pressure vs. mass number. Gas monitoring can be continuous or timed.

__Library and Histogram mode__ are combined to show the capabilities of split screen operation. The __Histogram mode__ displays a bar graph of partial pressure vs. mass number. This mode is most often used for vacuum analysis. The __Library mode__ contains a comprehensive list of gases and a search mode that allows the selection of up to 12 masses for identification.

__Table mode__ provides a readout of mass, scaling factor and true partial pressure for numerical analysis. Peak heights and alarm status of 10 masses can be displayed. The electron multiplier can be set on or off for each mass allowing the detection of minor species even in the presence of high total pressure.



__Leak Detection mode__ monitors a particular mass number over time. A programmable audible tone that changes pitch proportionally with partial pressure is useful in detecting the location of a leak.

Inlet

Type	Capillary: available in stainless steel, PEEK, and glass lined plastic.
Flowrate	1 to 10 milliliter per minute at atmospheric pressure.
Pressure	Selectable from 10 mbar to 1 bar.

Mass Spectrometer

Type	Quadrupole
Detectors	Faraday cup and electron multiplier
Range	100, 200 or 300 atomic mass units (amu)
Resolution	Less than 0.5 amu at 10% of peak height.
Detection limit	Less than 1 ppm
Operating pressure	5 x 10^{-6} mbar

Connections

Inlet	1/4" Ultra-Torr® fitting
Computer interface	RS-232C, 28.8k baud, DB9 connector. 25' RS-232 cable included.

System

High vacuum pump	Hybrid turbo-molecular
Diaphragm pump	Advanced low pressure

Materials

Construction	SS304 and SS316
Insulators	Alumina, ceramic
Seals	Viton®, buna-N, and nitrile butyl rubber
Miscellaneous	Aluminum, Tygon®

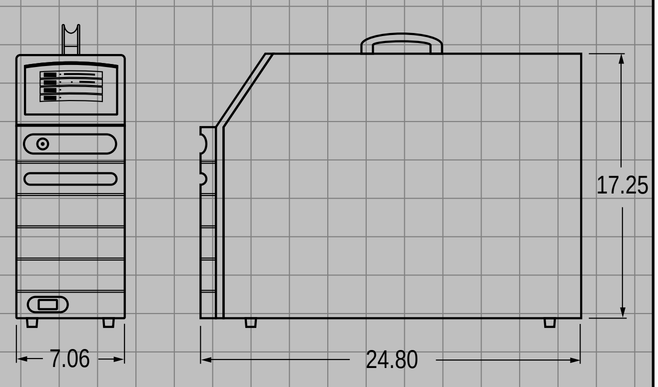
Software

RGA Windows®	Included, Runs under Windows® 3.1, Windows 95® and Windows NT® on 486 or higher

General

Startup time	5 minutes from full stop
Maximum ambient operating temperature	35 °C
Power requirement	110 VAC @ 60 Hz or 220 VAC @ 50 Hz (not field selectable)
Dimensions	See below
Weight	35 kg (75 lbs.)
Warranty	1 year parts and labor on materials and workmanship.

QMS 100/200/300 Dimensions



Ordering Information

QMS100	**100 amu Gas Analyzer**	**$23,000**
QMS200	**200 amu Gas Analyzer**	**$23,750**
QMS300	**300 amu Gas Analyzer**	**$25,250**
Option01	**High Vacuum Rear Port**	**$500**
O100BV	**Bypass Valve with pressure Reduction Orifice**	**$3000**
O100RO	**Replacement Orifice for Bypass Valve**	**$50**
O100TR	**Computer Output Card Cable, Relay Card**	**$595**
O100TS	**Computer Output Card Cable, TTL Card**	**$290**
O100CS	**Computer System: Pentium, Hard Drive, CDROM,15" VGA Monitor, Windows 95®**	**$2000**
O100RM	**19" Computer Rack Mount**	**$500**

Application Note #1

About FFT Spectrum Analyzers

What is an FFT Spectrum Analyzer?

FFT Spectrum Analyzers, such as the SR760, SR770, and SR780 take a time varying input signal, like you would see on an oscilloscope trace, and compute its frequency spectrum.

Fourier's theorem states that any waveform in the time domain can be represented by the weighted sum of sines and cosines. The FFT spectrum analyzer samples the input signal, computes the magnitude of its sine and cosine components, and displays the spectrum of these measured frequency components.

Why Look at a Signal's Spectrum?

For one thing, some measurements which are very hard in the time domain are very easy in the frequency domain. Consider the measurement of harmonic distortion. It's hard to quantify the distortion of a sine wave by looking at the signal on an oscilloscope. When the same signal is displayed on a spectrum analyzer, the harmonic frequencies and amplitudes are displayed with amazing clarity. Another example is noise analysis. Looking at an amplifier's output noise on an oscilloscope basically measures just the total noise amplitude. On a spectrum analyzer, the noise as a function of frequency is displayed. It may be that the amplifier has a problem only over certain frequency ranges. In the time domain it would be very hard to tell.

Many of these measurements were once done using analog spectrum analyzers. In simple terms, an analog filter was used to isolate frequencies of interest. The signal power which passed through the filter was measured to determine the signal strength in certain frequency bands. By tuning the filters and repeating the measurements, a spectrum could be obtained.

The FFT Analyzer

An FFT spectrum analyzer works in an entirely different way. The input signal is digitized at a high sampling rate, similar to a digitizing oscilloscope. Nyquist's theorem says that as long as the sampling rate is greater than twice the highest frequency component of the signal, then the sampled data will accurately represent the input signal. In the SR7xx, (we'll use SR7xx to refer to any of the SR760/SR770/SR780 analyzers) sampling occurs at 256 kHz. To make sure that Nyquist's theorem is satisfied, the input signal passes through an analog filter which attenuates all frequency components above 156 kHz by 90 dB. This is the anti-aliasing filter. The resulting digital time record is then mathematically transformed into a frequency spectrum using an algorithm known as the Fast Fourier Transform or FFT. The FFT is simply a clever set of operations which implements Fourier's theorem. The resulting spectrum shows the frequency components of the input signal.

Now here's the interesting part. The original digital time record comes from discrete samples taken at the sampling rate. The corresponding FFT yields a spectrum with discrete frequency samples. In fact, the spectrum has half as many frequency points as there are time points. (Remember Nyquist's theorem). Suppose that you take 1024 samples at 256 kHz. It takes 4 ms to take this time record. The FFT of this record yields 512 frequency points, but over what frequency range? The highest frequency will be determined by the period of 2 time samples or 128 kHz. The lowest frequency is just the period of the entire record or 1/(4 ms) or 250 Hz. Everything below 250 Hz is considered to be DC. The output spectrum thus represents the frequency range from DC to 128 kHz with points every 250 Hz.

Advantages of FFT Analyzers

The advantage of this technique is its speed. Because FFT spectrum analyzers measure all frequency components at the same time the technique offers the possibility of being hundreds of times faster than traditional analog spectrum analyzers. In the case of a 100 kHz span and 400 resolvable frequency bins, the entire spectrum takes only 4 ms to measure. To measure the signal with higher resolution the time record is increased, but again, all frequencies are examined simultaneously, providing an enormous speed advantage.

In order to realize the speed advantages of this technique we need to do high speed calculations. And, in order to avoid sacrificing dynamic range, we need high resolution A/D converters. SRS spectrum analyzers have the processing power and front end resolution needed to realize the theoretical benefits of FFT spectrum analyzers.

Dual Channel FFT Analyzers

One of the most common applications of FFT spectrum analyzers is to measure the transfer function of a mechanical or electrical system, that is, the ratio of the output spectrum to the input spectrum. Single channel analyzers such as the SR760, cannot measure transfer functions. Single channel analyzers with integrated sources, such as the SR770, can measure transfer functions, but only by assuming that the input spectrum of the system is equal to the spectrum of the integrated source. In general, to measure a general transfer function a two-channel analyzer, such as the SR780, is required. One channel measures the spectrum of the input, the other measures the spectrum of the output, and the analyzer performs a complex division to extract the magnitude and phase of the transfer function. Because the input spectrum is actually measured and divided out, you're not limited to using a predetermined signal as the input to the system under test—any signal will do.

Frequency Spans

Before we continue, let's clarify a couple of points about our frequency span. We just described how we arrived at a DC to 128 kHz frequency span using a 4 ms time record. Because the signal passes through an anti-aliasing filter at the input, the entire frequency span is not useable. The filter has a flat response from DC to 100 kHz and then rolls off steeply from 100 kHz to 156 kHz. No filter can make a 90 dB transition instantly. The range between 100 kHz and 128 kHz is therefore not useable and the actual displayed frequency span stops at 100 kHz. There is also a frequency bin labeled 0 Hz (or DC). This bin actually covers the range from 0 Hz to 250 Hz (the lowest measurable frequency) and contains the signal components whose period is longer than the time record (not only DC). So our final displayed spectrum contains 400 frequency bins. The first covers 0 - 250 Hz, the second 250 - 500 Hz, and the 400th covers 99.75 - 100.0 kHz.

Spans less than 100 kHz

The length of the time record determines the frequency span and resolution of our spectrum. What happens if we make the time record 8 ms or twice as long? Well, we ought to get 2048 time points (sampling at 256 kHz) yielding a spectrum from DC to 100 kHz with 125 Hz resolution containing 800 points. But the SR7xx places some limitations on this. One is memory. If we keep increasing the time record, then we would need to store more and more points. (1 Hz resolution would require 256 k values.) Another limitation is processing time. The time it takes to calculate an FFT with more points increases more than linearly.

To overcome this problem, the analyzer digitally filters and decimates the incoming data samples (at 256 kHz) to limit the bandwidth and reduce the number of points in the FFT. This is similar to the anti-aliasing filter at the input except the digital filter's cutoff frequency can be changed. In the case of the 8 ms record, the filter reduces the bandwidth to 64 kHz with a filter cutoff of 50 kHz (the filter rolls off between 50 and 64 kHz). Remember that Nyquist only requires samples at twice the frequency of the highest signal frequency. Thus the digital filter only has to output points at 128 kHz or half of the input rate (256 kHz). The net result is the digital filter outputs a time record of 1024 points effectively sampled at 128 kHz to make up an 8 ms record. The FFT processor operates on a constant number of points and the resulting FFT will yield 400 points from dc to 50 kHz. The resolution or linewidth is 125 Hz.

This process of doubling the time record and halving the span can be repeated by using multiple stages of digital filtering. The SR7xx can process spectra with a span of only 191 mHz with a time record of 2098 seconds if you have the patience. However, this filtering process only yields baseband measurements (frequency spans which start at DC).

Starting the span somewhere other than DC

In addition to choosing the span and resolution of the spectrum, we may want the span to start at frequencies other than DC. It would be nice to center a narrow span around any frequency below 100 kHz. Using digital filtering alone requires that every span start at DC. We need to frequency shift, or heterodyne, the input signal. Multiplying the incoming signal by a complex sine will frequency shift the signal. The resulting spectrum is shifted by the frequency of the complex sine. If we incorporate heterodyning with our digital filtering, we can shift any frequency span so that it starts at DC. The resulting FFT yields a spectrum offset by the heterodyne frequency.

Heterodyning allows the analyzer to compute zoomed spectra (spans which start at frequencies other than DC). The digital filter processor can filter and heterodyne the input in real time to provide the appropriate filtered time record at all spans and center frequencies. Because the digital signal processors in the SR7xx are so fast, you won't notice any calculation time while taking spectra. All the signal processing calculations, heterodyning, digitally filtering, and Fourier transforming, are done in less time than it takes to acquire the data, so the SR7xx can take spectra seamlessly, i.e. there is no dead time between one time record and the next.

Measurement Basics

An FFT spectrum is a complex quantity. This is because each frequency component has a phase relative to the start of the time record. (Alternately, you may wish to think of the input signal being composed of sines and cosines.) If there is no triggering, then the phase is random and we generally look at the magnitude of the spectrum. If we use a synchronous trigger then each frequency component has a well defined phase.

Spectrum

The spectrum is the basic measurement of an FFT analyzer. It is simply the complex FFT. Normally, the magnitude of the spectrum is displayed. The magnitude is the square root of the FFT times its complex conjugate. (Square root of the sum of the real (sine) part squared and the imaginary (cosine) part squared). The magnitude is a real quantity and represents the total signal amplitude in each frequency bin, independent of phase.

If there is phase information in the spectrum, i.e. the time record is triggered in phase with some component of the signal, then the real (cosine) or imaginary (sine) part or the phase may be displayed. The phase is simply the arctangent of the ratio of the imaginary and real parts of each frequency component. The phase is always relative to the start of the triggered time record.

Power Spectral Density (PSD)

The PSD is simply the magnitude of the spectrum normalized to a 1 Hz bandwidth. This measurement approximates what the spectrum would look like if each frequency component were really a 1 Hz wide piece of the spectrum at each frequency bin.

What good is this? When measuring broadband signals such as noise, the amplitude of the spectrum changes with the frequency span. This is because the linewidth changes so the frequency bins have a different noise bandwidth. The PSD, on the other hand, normalizes all measurements to a 1 Hz bandwidth and the noise spectrum becomes independent of the span. This allows measurements with different spans to be compared. If the noise is Gaussian in nature, then the amount of noise amplitude in other bandwidths may be approximated by scaling the PSD measurement by the square root of the bandwidth. Thus the PSD is displayed in units of V/√Hz or dBV/√Hz.

Since the PSD uses the magnitude of the spectrum, the PSD is a real quantity. There is no real or imaginary part or phase.

Time Record

The time record measurement displays the filtered and decimated (depending on the span) data points before the FFT is taken. In the SR760 and SR770 this information is available only at full span. In the SR780, time records can be displayed at all spans. For baseband spans (spans that start at DC), the time record is a real quantity. For non-baseband spans the heterodyning discussed earlier transforms the time record into a complex quantity which can be somewhat difficult to interpret.

Two-Channel Measurements

As we discussed earlier, two-channel analyzers, such as the SR780, offer additional measurements such as transfer function, cross-spectrum, coherence and orbit. These measurements, which only apply to the SR780, are discussed below.

Transfer Function

The transfer function is the ratio of the spectrum of channel 2 to the spectrum of channel 1. For the transfer function to be valid, the input spectrum must have amplitude at all frequencies over which the transfer function is to be measured. For this reason, broadband sources such as noise, or periodic chirps, are often used as inputs for transfer function measurements.

Cross Spectrum

The cross spectrum is defined as:

cross spectrum = FFT2 conj(FFT1)

The cross spectrum is a complex quantity which contains magnitude and phase information. The phase is the relative phase between the two channels. The magnitude is simply the product of the magnitudes of the two spectra. Frequencies where signals are present in both spectra will have large components in the cross-spectrum.

Orbit

The orbit is simply a two dimensional display of the time record of channel 1 vs. the time record of channel 2. The orbit display is similar to an oscilloscope displaying a “Lissajous” figure.

Coherence

Coherence measures the percentage of power in channel 2 which is caused by (phase coherent with) power in the input channel. Coherence is a unitless quantity which varies from 0 to1. If the coherence is 1, all the power of the output signal is due to the input signal. If the coherence is 0, the input and output are completely random with respect to one another. Coherence is related to signal to noise ratio (S/N) by the formula:

$$S/N = \gamma^2/(1-\gamma^2)$$

where γ^2 is the traditional notation for coherence.

Correlation

The SR780 analyzer also computes auto and cross-correlation. Correlation is a *time* domain measurement which is defined as follows:

$$\text{Auto Correlation}(\tau) = \int x^*(t)x(t-\tau)dt$$

$$\text{CrossCorrelation}(\tau) = \int x^*(t)y(t-\tau)dt$$

where x and y are the channel 1 and channel 2 input signals and the integrals are over all time. It is clear that the auto correlation at a time t is a measure of how much overlap a signal has with a delayed- by-t version of itself, and the cross-correlation is a measure of how much overlap a signal has with a delayed-by-t version of the other channel. Although correlation is a time-domain measurement the SR780 uses frequency-domain techniques to compute it in order to make the calculation faster.

Spectrum

The most common measurement is the spectrum and the most useful display is the Log Magnitude. The Log Mag display graphs the magnitude of the spectrum on a logarithmic scale using dBV as units.

Why is the Log Mag display useful? Remember that the SR7xx has a dynamic range of about 90 dB.below full scale. Imagine what something 0.01% of full scale would look like on a linear scale. If we wanted it to be 1 centimeter high on the graph, the top of the graph would be 100 meters above the bottom. It turns out that the log display is both easy to understand and shows features which have very different amplitudes clearly.

Of course the analyzer is capable of showing the magnitude on a linear scale if you wish.

The real and imaginary parts are always displayed on a linear scale. This avoids the problem of taking the log of negative voltages.

Phase

In general, phase measurements are only used when the analyzer is triggered. The phase is relative to the start of the time record.

The phase is displayed in degrees or radians on a linear scale from -180 to +180 degrees. There is no phase "unwrap” on the SR760 and SR770. The SR780 can display unwrapped phase, however, which is very useful, for instance, in displaying the phase of filter transfer functions which may vary over hundreds or even thousands of degrees.

The phase of a particular frequency bin is set to zero if neither the real nor imaginary part of the FFT is greater than 0.012% of full scale (-78 dB below f.s.). This avoids the messy phase display associated with the noise floor. (Remember, even if a signal is small, its phase extends over the full 360 degrees.)

Watch out for phase errors

The FFT measurement can be thought of as N band-pass filters, each centered on a frequency bin. The signal within each filter shows up as the amplitude of each bin. If a signal's frequency is between bins, the filters act to attenuate the signal a little bit. This results in a small amplitude error. The phase error, on the other hand, can be quite large. Because these filters are very steep and selective, they introduce very large phase shifts for signals not exactly on a frequency bin.

On full span, this is generally not a problem. The bins are 250 Hz apart and most synthesized sources have no problem generating a signal right on a frequency bin. But when the span is narrowed, the bins move much closer together and it becomes very hard to place a signal exactly on a frequency bin.

Windowing

What is windowing? Let's go back to the time record. What happens if a signal is not exactly periodic within the time record? We said that its amplitude is divided into multiple adjacent frequency bins. This is true but it's actually a bit worse than that. If the time record does not start and stop with the same data value, the signal can actually smear across the entire spectrum. This smearing will also change wildly between records because the amount of mismatch between the starting value and ending value changes with each record.

Windows are functions defined across the time record which are periodic in the time record. They start and stop at zero and are smooth functions in between. When the time record is windowed, its points are multiplied by the window function, time bin by time bin, and the resulting time record is by definition periodic. It may not be identical from record to record, but it will be periodic (zero at each end).

In the Frequency Domain

In the frequency domain, a window acts like a filter. The amplitude of each frequency bin is determined by centering this filter on each bin and measuring how much of the signal falls within the filter. If the filter is narrow, then only frequencies near the bin will contribute to the bin. A narrow filter is called a selective window — it selects a small range of frequencies around each bin. However, since the filter is narrow, it falls off from center rapidly. This means that even frequencies close to the bin may be attenuated somewhat. If the filter is wide, then frequencies far from the bin will contribute to the bin amplitude but those close by will not be attenuated significantly.

The net result of windowing is to reduce the amount of smearing in the spectrum from signals not exactly periodic with the time record. The different types of windows trade off selectivity, amplitude accuracy, and noise floor.

The SR7xx offers several types of window functions including Uniform (none), Flattop, Hanning, Blackman-Harris, and Kaiser.

Uniform

The uniform window is actually no window at all. The time record is used with no weighting. A signal will appear as narrow as a single bin if its frequency is exactly equal to a frequency bin. (It is exactly periodic within the time record). If its frequency is between bins, it will affect every bin of the spectrum. These two cases also have a great deal of amplitude variation between them (up to 4 dB).

In general, this window is only useful when looking at transients which do not fill the entire time record.

Hanning

The Hanning window is the most commonly used window. It has an amplitude variation of about 1.5 dB (for signals between bins) and provides reasonable selectivity. Its filter rolloff is not particularly steep. As a result, the Hanning window can limit the performance of the analyzer when looking at signals close together in frequency and very different in amplitude.

Flattop

The Flattop window improves on the amplitude accuracy of the Hanning window. Its between-bin amplitude variation is about .02 dB. However, the selectivity is a little worse. Unlike the Hanning, the Flattop window has a wide pass band and very steep rolloff on either side. Thus, signals appear wide but do not leak across the whole spectrum.

Blackman-Harris

The Blackman-Harris window is a very good window to use with the SR760 and SR770. It has better amplitude accuracy (about 0.7 dB) than the Hanning, very good selectivity and the fastest filter rolloff. The filter is steep and narrow and reaches a lower attenuation than the other windows. This allows signals close together in frequency to be distinguished, even when their amplitudes are very different.

Kaiser

The Kaiser window, which is available on the SR780 only, combines excellent selectivity and reasonable accuracy (about 0.8 dB for signals between exact bins). The Kaiser window has the lowest side-lobes and the least broadening for non-bin frequencies. Because of these properties, it is the best window to use for measurements requiring a large dynamic range. On the SR760 and SR770, the Blackman-Harris window is the best large dynamic rage window.

Averaging

The SR7xx analyzers supports several types of averaging. In general, averaging many spectra together improves the accuracy and repeatability of measurements.

RMS Averaging

RMS averaging computes the weighted mean of the sum of the squared magnitudes (FFT times its complex conjugate). The weighting is either linear or exponential.

RMS averaging reduces fluctuations in the data but does not reduce the actual noise floor. With a sufficient number of averages, a very good approximation of the actual random noise floor can be displayed.

Since RMS averaging involves magnitudes only, displaying the real or imaginary part or phase of an RMS average has no meaning. The RMS average has no phase information.

Vector Averaging

Vector averaging averages the complex FFT spectrum. (The real part is averaged separately from the imaginary part.) This can reduce the noise floor for random signals since they are not phase coherent from time record to time record.

Vector averaging requires a trigger. The signal of interest must be both periodic and phase synchronous with the trigger. Otherwise, the real and imaginary parts of the signal will not add in phase and instead will cancel randomly.

With vector averaging, the real and imaginary parts as well as phase displays are correctly averaged and displayed. This is because the complex information is preserved.

Peak Hold

Peak Hold is not really averaging, instead, the new spectral magnitudes are compared to the previous data, and if the new data is larger, then the new data is stored. This is done on a frequency bin by bin basis. The resulting display shows the peak magnitudes which occurred in the previous group of spectra.

Peak Hold detects the peaks in the spectral magnitudes and only applies to Spectrum, PSD, and Octave Analysis measurements. However, the peak magnitude values are stored in the original complex form. If the real or imaginary part or phase is being displayed for spectrum measurements, the display shows the real or imaginary part or phase of the complex peak value.

Linear Averaging

Linear averaging combines N (number of averages) spectra with equal weighting in either RMS, Vector or Peak Hold fashion. When the number of averages has been completed, the analyzer stops and a beep is sounded. When linear averaging is in progress, the number of averages completed is continuously displayed below the Averaging indicator at the bottom of the screen.

Auto ranging is temporarily disabled when a linear average is in progress. Be sure that you don't change the input range manually. Changing the range during a linear average invalidates the results.

Exponential Averaging

Exponential averaging weights new data more than old data. Averaging takes place according to the formula,

$$\text{New Average} = (\text{New Spectrum} \bullet 1/N) + (\text{Old Average}) \bullet (N-1)/N$$

where N is the number of averages.

Exponential averages "grow" for approximately the first 5N spectra until the steady state values are reached. Once in steady state, further changes in the spectra are detected only if they last sufficiently long. Make sure that the number of averages is not so large as to eliminate the changes in the data that might be important.

Real Time Bandwidth and Overlap Processing

What is real time bandwidth? Simply stated, it is the frequency span whose corresponding time record exceeds the time it takes to compute the spectrum. At this span and below, it is possible to compute the spectra for every time record with no loss of data. The spectra are computed in "real time". At larger spans, some data samples will be lost while the FFT computations are in progress.

For all frequency spans, the SR7xx can compute the FFT in less time than it takes to acquire the time record. Thus, the real time bandwidth of the SR7xx is 100 kHz. This includes the real time digital filtering and heterodyning, the FFT processing, and averaging calculations. The SR7xx employs two digital signal processors to accomplish this. The first collects the input samples, filters and heterodynes them, and stores a time record. The second computes the FFT and averages the spectra. Since both processors are working simultaneously, no data is ever lost.

The SR780 accomplishes its high-speed processing with a single advanced technology floating-point DSP chip.

Averaging Speed

How can you take advantage of this? Consider averaging. Other analyzers typically have a real time bandwidth of around 4 kHz. This means that even though the time record at 100 kHz span is only 4 ms, the "effective" time record is 25 times longer due to processing overhead. An analyzer with 4 kHz of real time bandwidth can only process about 10 spectra a second. When averaging is on, this usually slows down to about 5 spectra per second. At this rate it's going to take a couple of minutes to do 500 averages.

The SR7xx, on the other hand, has a real time bandwidth of 100 kHz. At a 100 kHz span, the analyzer is capable of processing 250 spectra per second. In fact, this is so fast, that the display can not be updated for each new spectra. The display only updates about 6 times a second. However, when averaging is on, all of the computed spectra will contribute to the average. The time it takes to complete 500 averages is only a few seconds. (Instead of a few minutes!)

Overlap

What about narrow spans where the time record is long compared to the processing time? The analyzer computes one FFT per time record and can wait until the next time record is complete before computing the next FFT. The update rate would be no faster than one spectra per time record. With narrow spans, this could be quite slow.

And what is the processor doing while it waits? Nothing. With overlap processing, the analyzer does not wait for the next complete time record before computing the next FFT. Instead it uses data from the previous time record as well as data from the current time record to compute the next FFT. This speeds up the processing rate. Remember, most window functions are zero at the start and end of the time record. Thus, the points at the ends of the time record do not contribute much to the FFT. With overlap, these points are "re-used" and appear as middle points in other time records. This is why overlap effectively speeds up averaging and smooths out window variations.

Typically, time records with 50% overlap provide almost as much noise reduction as non-overlapping time records when RMS averaging is used. When RMS averaging narrow spans, this can reduce the measurement time by a factor of two.

Overlap Percentage

The amount of overlap is specified as a percentage of the time record. 0% is no overlap and 99.8% is the maximum (511 out of 512 samples re-used). The maximum overlap is determined by the amount of time it takes to calculate an FFT and the length of the time record and thus varies according to the span.

The SR760/SR770 always try to use the maximum amount of overlap possible. This keeps the display updating as fast as possible. Whenever a new frequency span is selected, the overlap is set to the maximum possible value for that span. If less overlap is desired, then use the Average menu to enter a smaller value. On the widest spans (25, 50 and 100 kHz), no overlap is allowed.

The SR780 uses a slightly different system for specifying the overlap. The overlap entered by the user is the “requested overlap.” The instrument attempts to make the actual overlap as close as possible to the requested overlap. The SR780 computes and displays the actual overlap so that it is obvious when it differs from the requested overlap.

Octave Analysis

The magnitude of the normal spectrum measures the amplitudes within equally divided frequency bins. Octave analysis computes the spectral amplitude in logarithmic frequency bands whose widths are proportional to their center frequencies. The bands are arranged in octaves with either 1, 3, or 12 bands per octave (1/1, 1/3, or 1/12 octave analysis). Octave analysis measures spectral power closer to the way people perceive sound, that is, in octaves.

The actual method used to calculate octave measurements differs for each of the analyzers. In the SR780, the input data passes into a bank of parallel digital filters. The filter center frequencies and shapes are determined by the type of octave analysis, either full (1/1), 1/3 or 1/12 octave. and comply with ANSI s1-11-1986, Order 3, Type 1-D. The output of each filter is rms averaged to compute the power and displayed as a bar-type graph. This is a real-time measurement of the power within each band and is the only available octave measurement. Since the bands are spaced logarithmically, octave displays always have a logarithmic X-axis.

Band Center Frequencies

The center frequency of each band is calculated according to ANSI standard S1.11 (1986). The shape of each band is a third-order Butterworth filter whose bandwidth is either a full, 1/3, or 1/12 octave. The full octave bands have band centers at:

$$\text{Center Freq: } = 1 \text{ kHz} \times 2^{n}$$

The 1/3 octave bands have center frequencies given by:

$$\text{Center Freq: } = 1 \text{ kHz} \times 2^{(n-30/3)}$$

Finally, the SR780 only can calculate octave power in 1/12 octave bins whose center frequencies are at:

$$\text{Center Freq: } = 1 \text{ kHz} \times 2^{1/24} 2^{n/12}$$

Swept Sine Measurements

The SR780 contains an additional measurement mode, the swept sine mode, which is useful for making measurements with high dynamic range. A swept sine measurement is basically a sine sweep which steps through a specified sequence of frequency points. At each point the source maintains a constant frequency and the inputs measure only signals at this frequency. After each point has been measured, the source moves on to the next point in the sequence. Unlike the FFT which measures many frequencies at once, swept sine measures one frequency at a time. As we'll see, this technique is somewhat slower, but leads to increases in dynamic range.

Transfer functions can be measured using the FFT mode or the swept-sine mode. However, if the transfer function has a large variation within the measurement span, then the FFT may not be the best measurement technique. It's limitation comes from the nature of the chirp source that must be used. The FFT measures the response at all frequencies within the span simultaneously, thus the source must contain energy at all of the measured frequencies. In the time record, the frequency components in the source add up and the peak source amplitude within the time record generally exceeds the amplitude of each frequency component by about 30 dB. Since the input range must be set to accommodate the amplitude peak, each component is measured at -30 dB relative to full scale. This effectively reduces the dynamic range of the measurement by about 30 dB! If the transfer function has a variation from 0 to -100 dB within the measurement span, then each bin of the FFT must measure signals from -30 dBfs to -130 dBfs. Even with a large number of vector averages, this proves difficult, especially with large measurement spans.

Swept sine measurements, on the other hand, can optimize the measurement at each frequency point. Since the source is a sine wave, all of the source energy is concentrated at a single frequency, eliminating the 30 dB chirp dynamic range penalty. In addition, if the transfer response drops to -100 dBV, the input range of Channel 2 can auto range to -50 dBV and maintain almost 100 dB of signal to noise. In fact, simply optimizing the input range at each frequency can extend the dynamic range of the measurement to beyond 140 dB.
For transfer functions with both gain and attenuation, the source amplitude can be optimized at each frequency. Reducing the source level at frequencies where there is gain prevents overloads and increasing the amplitude where there is attenuation preserves signal to noise. To optimize the measurement time of sweeps covering orders of magnitude in frequency, the detection bandwidth can be set as a function of frequency. More time can be spent at lower frequencies and less time at higher frequencies. In addition, frequency points can be skipped in regions where the response does not change significantly from point to point. This speeds measurements of narrow response functions.

Application Note #2

Making Measurements with the SR620 Time Interval Counter

In order to realize the full potential of the SR620 Universal Time Interval and Frequency Counter a brief look at its several modes of operation and some of their applications is appropriate. It is also important for a user to understand the performance specifications of the SR620 in order to draw reasonable conclusions from experimental data. How accurate is a particular measurement? What errors can be expected? This application note defines, in detail, the performance specifications and terminology for each mode of operation. This information will enable the user to fully understand the capabilities and limitations of the instrument.

The first section of this application note describes the type of measurements that the SR620 can perform and gives some application examples. It will give the reader an appreciation for the wide range of applications in which the SR620 Universal Time Interval and Frequency Counter can be used.

The second section (SR620 Specification Guide) explores the specifications of the SR620 in detail. A precise definition of measurement accuracy (resolution and error) along with explanations of technical terminology is given for each mode of operation.

Applications of the SR620

The SR620 Universal Time Interval and Frequency Counter has been used to measure everything from the propagation delays of integrated circuits to the distance to the moon. Its versatility and affordable price have made the SR620 an instant success in the engineering and scientific marketplaces. The SR620 can be used to measure time interval, frequency, period, pulse width, phase, rise and fall time and will also do event counting. Statistical calculations including mean, standard deviation, Allan variance, minimum and maximum are performed on up to one million samples in all modes of operation. In addition to displaying the statistical data on the 16 digit LED display, distribution graphs of the data, in histogram and strip chart format, can be displayed on an X-Y oscilloscope. Application examples of the SR620 are discussed below.

Time Measurement

The SR620 measures the time interval between two independent signals, A and B. An example using this mode of operation is measuring the electrical length of a cable. The cable can be configured as end to end or single ended with the remote end shorted to ground or left open. Using the built-in 1 kHz reference signal as a stimulus, the propagation delay from one end of the cable to the other or between the incident and reflected rising edge of the pulse can be measured. Knowing that electricity travels at approximately one foot per 1.5 nanoseconds, the cable length can easily be calculated.

Another time interval application is the measurement of propagation delays of integrated circuits. Again, the 1 kHz reference source can be used to excite the experiment and the time delay from the input to the output of the integrated circuit can be measured.

Pulse Width Measurement

Magnetic and optical memory disk data is stored using different modulation schemes to minimize disk real-estate and maximize signal to noise ratio. For example C.D. players use 3-11 modulation to obtain very high disk density. This scheme produces data patterns with nine different pulse widths (corresponding to 3, 4, 5, ...11 consecutive 0's). The SR620 can be used to measure these pulse widths and their variations and display them graphically in histogram form on an X-Y oscilloscope.

Rise and Fall Time Measurements

When analyzing the transition time required for a least significant bit change in a DAC (digital to analog converter), the 10-90% rise time of that transition is of importance. Once the rise time has been established, the small signal frequency response of the DAC can be calculated (BW = 0.35 / rise time). The SR620 allows the user to set the start and stop voltage thresholds for maximum flexibility in rise and fall time measurements so that any part of a transition may be analyzed.

Frequency and Period Measurements

When measuring the quality of a reference frequency source the jitter (standard deviation or Allan variance) is often of significance. The SR620 will analyze the source over a set gate time and then display a distribution curve of the data showing the mean frequency, minimum and maximum frequencies and the jitter revealing the quality of the source. Frequency is measured as $N/(\Delta t)$ and Period is measured as $(\Delta t)/N$ where N is the number of cycles and Δt is the elapsed time to complete N cycles.

Phase Measurements

When characterizing operational amplifiers it is useful to know the phase verses frequency relationship. The SR620 can measure the difference in phase between the input and output at different frequencies so that a Bode plot can be constructed.

Event Counting

Used in conjunction with a discriminator, the SR620 can function as a photon counter that counts electrical pulses from a PMT (photomultiplier tube). It can count at a rate of up to 200 MHz. In another application, the SR620 has proven to be a cost effective way to count canned goods traveling on a conveyer belt as they pass a check point.

These examples represent only a few of many possible applications of the SR620. A full description of how to perform these and other types of measurements is given in the SR620's operation and service manual.

SR620 Specification Guide

This section provides an explanation of the specifications of the SR620 and their effect on the accuracy and resolution of a measurement.

Statistical Functions

The SR620 can display statistical information about the measurement of N samples. The SR620 computes and reports the mean, standard deviation or root Allan variance, minimum, and maximum values seen during the measurement. The equation for the statistical functions are given by:

$$\text{standard deviation} = \sqrt{\frac{n\sum_{i=1}^{n} x_i^2 - \left(\sum_{i=1}^{n} x_i\right)^2}{n(n-1)}}$$

$$\text{root Allan variance} = \sqrt{\frac{\sum_{i=1}^{n-1} (x_{i+1} - x_i)^2}{2(n-1)}}$$

Least Significant Digit (LSD)

The LSD is the smallest displayed increment in a measurement. The SR620 has a 4 ps single-shot LSD and thus the smallest amount that two single-shot time interval measurements may differ by is 4 ps.

Resolution

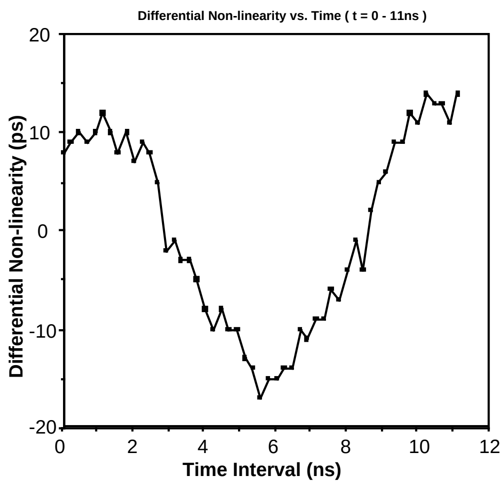
Resolution is the smallest difference in a measurement that the SR620 can discern. That is, the smallest statistically significant change which can be measured by the SR620. Resolution is of primary interest in comparing readings from the same instrument. The instrument resolution is limited by many things including short-term timebase stability, internal noise, trigger noise,etc.. Because these processes are random in nature, resolution is specified as an rms value rather than a peak value. This rms value is the standard deviation of the measured value. The SR620's single-shot resolution is typically 25 ps rms. This number can be improved by averaging over many measurements, or in the case of frequency and period measurements, increasing the gate time. The single-shot LSD is always smaller than the single-shot resolution.

Error

Error is defined as the difference between the measured value and actual value of the signal being measured. The error in a measurement is of primary concern when the absolute value of the parameter being measured is important. Error consists of the random factors mentioned above and systematic uncertainties in the measurement. Systematic uncertainties include timebase aging, trigger level error, insertion delay, etc.. Systematic errors may always be measured and subtracted from subsequent measurements to reduce the error. The SR620's absolute error is typically less than 0.5 ns for time interval measurements less than 1 ms.

Differential Non-linearity

Absolute error is of interest in determining how far a value is from the actual value. Often only the relative accuracy (the difference between two measurements) is important. Differential non-linearity is a measurement of the relative accuracy of a measurement and is specified as the maximum time error for any given relative measurement. The SR620's differential non-linearity is typically ±50 ps. That means if the time interval is changed by some amount the SR620 will report that change correctly to within ±50 ps. Graphs 1 and 2 show the SR620's typical differential non-linearity as a function of time interval. Graph 1 shows the non-linearity over the time range of 0 to 11 ns. The deviations are due to the residual non-linearity of the time-to-amplitude converters used to interpolate a fraction of one 90 MHz clock tick. This curve repeats every 11.11 ns. Graph 2 shows the non-linearity over the time range of 0 to 11 ms. For times greater than 11 ms the non-linearity is dominated by the timebase error.



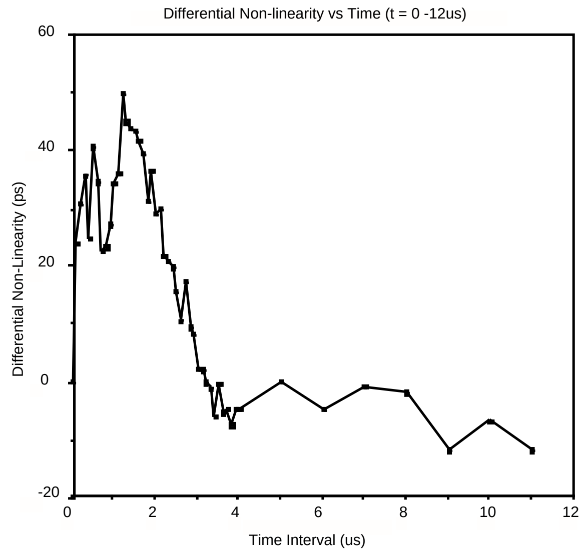
Graph 1: Differential Non-linearity for time differences of 0 to 11 ns showing the residual non-linearity of the time-to-amplitude converters.

Timebase Specifications

The specifications of the timebase affect both the resolution and error of measurements made with the SR620. A timebase may be specified by two parameters: its short-term stability and its long-term stability.

Short-term Stability

The short-term stability of an oscillator is a measure of the changes in the output frequency of the oscillator on a short time scale (seconds or less). These changes in the frequency are usually random and are due to internal oscillator noise, output level modulation,etc.. These random changes in frequency affect the resolution of the measurement just as other internal noise does. The The short-term stability of an oscillator is a measure of the changes in the output frequency of the oscillator on a short time scale (seconds or less). These changes in the frequency are usually random and are due to internal oscillator noise, output level modulation,etc.. These random changes in frequency affect the resolution of the measurement just as other internal noise does. The



Graph 2: Differential Non-linearity for time differences of 0 to 11 ms

short-term stability of an oscillator is usually characterized by specifying either its Allan variance or its phase noise. The SR620's timebase short-term stability is specified by its Allan variance. Typical values for several gate times are:

	standard oscillator	oven oscillator
1.0s gate	$1.2x10^{-10}$	$2.3x10^{-11}$
10s gate	$2.3x10^{-10}$	$1.0x10^{-11}$
100s gate	$1.25x10^{-9}$	$4.2x10^{-11}$
spec. limit at 1s	$2.0x10^{-10}$	$5.0x10^{-11}$

The resolution of the SR620 is specified as:

resolution = $[(25\text{ ps})^2 + (\Delta t \text{ x short-term stability})^2]^{1/2}$

Where Δt is the time interval and the resolution is given in ps rms. For time intervals greater than 125 ms (standard oscillator) or 500 ms (oven oscillator) the short-term stability of the timebase will dominate the resolution limit of the SR620.

Long-term Stability

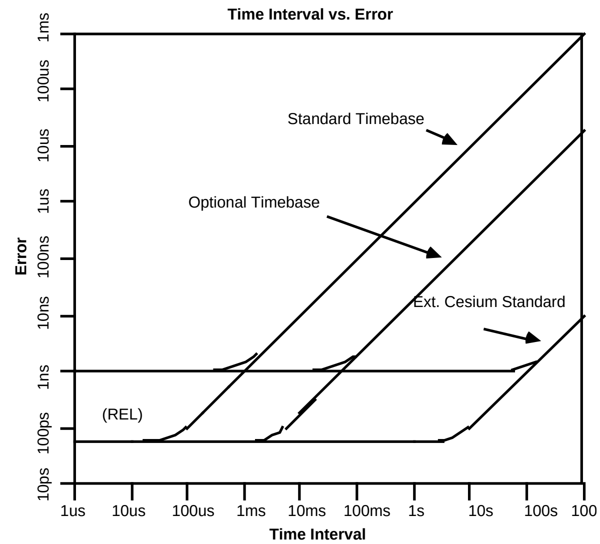
The long-term stability of an oscillator is a measure of its changes in frequency over long time intervals (hours, days, months, or years). It is the long term stability of the timebase that will ultimately limit the absolute accuracy of the SR620 and determines the calibration interval necessary to maintain a desired error limit. The long-term stability consists of two components: oscillator aging and oscillator temperature response. The aging of an oscillator is the change in frequency over time due to physical changes in the components (usually the crystal) and is usually specified as a fractional frequency change over some measurement period. Temperature response is due to changes in the oscillator characteristics as a function of ambient temperature and is specified as a fractional frequency change over some temperature range. The timebase for the SR620 is specified as:

	Standard Oscillator	Oven Oscillator
Aging	1.0×10^{-6}/yr	5.0×10^{-10}/day
Temp. Response	1.0×10^{-6} (0-50 °C)	5.0×10^{-9} (0-50 °C)

For example, the oven oscillator 30 days after calibration may have drifted at most 30 x 5×10^{-10} x 10 MHz = 0.15Hz. Also, a worst case temperature variation must be assumed when evaluating the worst case error. The optional oscillator must be assumed to be at worst 5 ppb in error because the conditions when the SR620 was calibrated are unknown. This worst case error is not a good estimate of the actual oscillator drift under most conditions.

External Timebases

The SR620 has a rear panel input that will accept either a 5 or 10 MHz external timebase. The SR620 phase-locks its internal timebase to this reference. The phase-locked loop has a bandwidth of about 20 Hz and thus the characteristics the the SR620's clock, for measurement times longer than 50 ms, become that of the external source. For shorter measurement times the clock characteristics are not important compared to the internal jitter (25 ps rms) of the SR620. Thus, if the signal from a Cesium clock is input into a SR620 with a standard TCXO oscillator the short-term and long-term stability of the SR620 will become that of the Cesium clock. This is illustrated in graph 3.



Graph 3: Error vs. Time Interval for Various Timebases

Trigger Input Specifications

There are two ways that the inputs can affect the resolution and accuracy of a measurement. The first is called trigger jitter and is due to random noise on the A and B input signals and the trigger input buffers. This random noise causes the input to trigger at a time different than it otherwise would in the absence of noise. Because this is a random process this affects the resolution just as the other random noise sources do. Trigger timing jitter can be minimized by careful grounding and shielding of the input and by maximizing the input slew rate. Note, however, that the slew rate is limited by the SR620's 1ns input rise time. The trigger timing jitter can be described by the equation:

$$\text{Trigger Timing Jitter} = \frac{\sqrt{\left(E_{internal}\right)^2 + \left(E_{signal}\right)^2}}{\text{Input Slew Rate}}$$

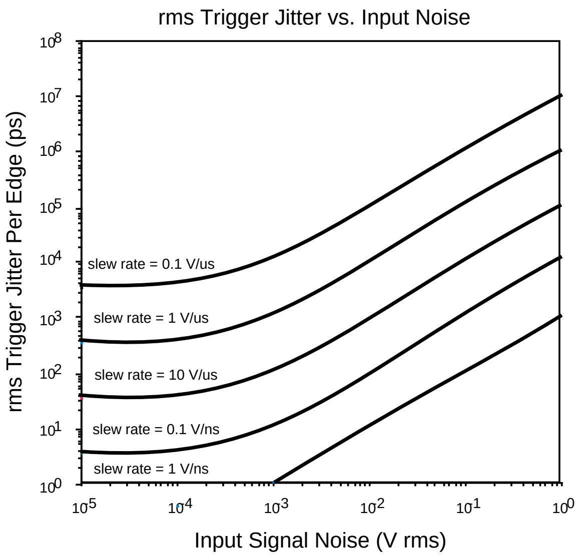
where

$$E_{internal} = \text{internal input noise } (350\mu\text{V rms typical})$$
$$E_{input} = \text{input signal noise}$$

If the trigger level is set to a value other than the intended value the time interval measured will be in error. This error (trigger level timing error) is a systematic error that affects only the error of the measurement and not its resolution. The SR620's trigger thresholds are set to an accuracy of 15 mV + 0.5% of value. The effect this has on the measurement is given by:

$$\text{Trigger Level Timing Error} = \frac{15\,\text{mV} + 0.5\%\ \text{of setting}}{\text{Input Slew Rate}}$$

Graphs 4 and 5 show the effects of trigger timing jitter and trigger timing level error on resolution and error. These graphs are applicable to all measurements, not just time intervals.

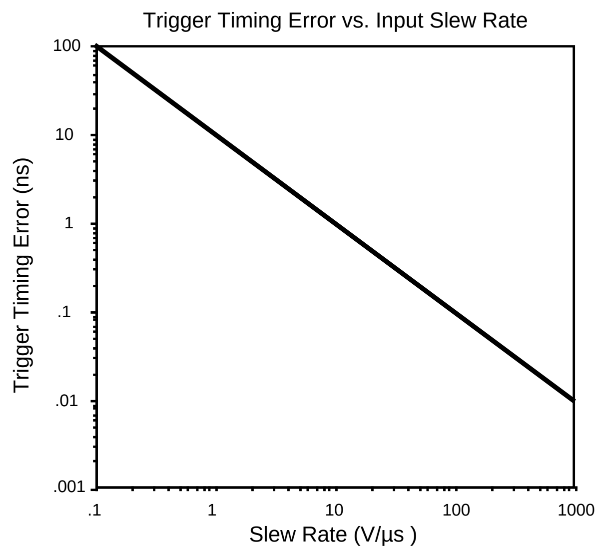


Graph 4: Effect of input noise on measeurement resolution showing reduced noise due to averaging.

Measurement Accuracy

The following equations allow one to calculate the SR620's resolution and error in the various measurement modes. The SR620's typical specifications are used in the following equations. For worst case bounds simply replace the typical with the worst case numbers.

NOTE: The quantities added to calculate the SR620's resolution are independent rms quantities and must be added in quadrature as shown in the following equation.



Graph 5: Effect of input slew rate on measurement error.

$$\text{total} = \sqrt{x_1^2 + x_2^2 + \ldots}$$

NOTE: "timebase error" refers to the sum of aging and temperature effects.

Time Interval, Width, Rise/fall Time Modes

In the time measurement modes the measurement resolution and error are given by:
where N = number of samples averaged

$$\text{Resolution} = \pm\sqrt{\frac{(25\,\text{ps})^2 + (\text{time interval} \times \text{short term stability})^2 + (\text{start trigger jitter})^2 + (\text{stop trigger jitter})^2}{N}}$$

$$\text{Error} = \pm[\text{resolution} + (\text{timebase error} \times \text{time interval}) + \text{start trigger level error} + \text{stop trigger level error} + 0.5\,\text{ns}]$$

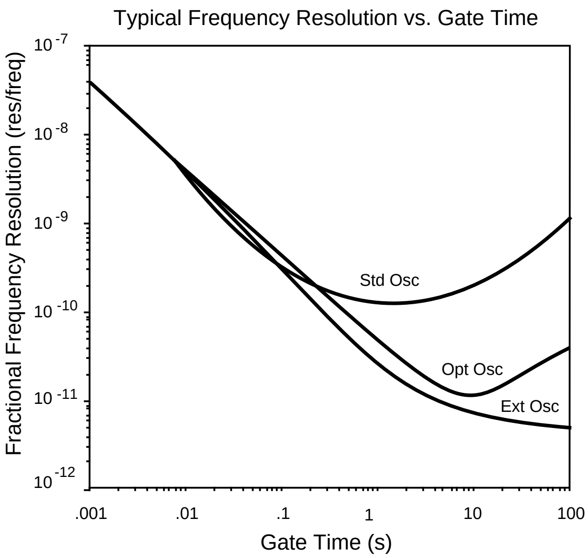
Frequency Mode

In frequency mode the measurement resolution and error are given by:

$$\text{Resolution} = \pm\frac{\text{frequency}}{\text{gate time}}\sqrt{\frac{(25\,\text{ps})^2 + (\text{short term stability} \times \text{gate time})^2 + 2 \times (\text{trigger jitter})^2}{N}}$$

$$\text{Error} = \pm\left[\text{resolution} + (\text{timebase error} \times \text{frequency}) + \frac{100\,\text{ps}}{\text{gate time}} \times \text{frequency}\right]$$

The SR620's typical single-shot frequency resolution as a function of gate time is shown in Graph 6. The curves are for the standard oscillator, the optional oven oscillator, and an external high stability reference. The input signal noise is assumed to be negligible.



Graph 6: Frequency resolution as a function of gate time for the SR620's three oscillator options.

Period Mode

In period mode the measurement resolution and error are given by:

$$\text{Resolution} = \pm \frac{\text{period}}{\text{gate time}} \sqrt{\frac{(25\,\text{ps})^2 + (\text{short term stability} \times \text{gate time})^2 + 2 \times (\text{trigger jitter})^2}{N}}$$

$$\text{Error} = \pm \left[\text{resolution} + (\text{timebase error} \times \text{period}) + \frac{100\,\text{ps}}{\text{gate time}} \times \text{period} \right]$$

where N = number of samples averaged

Phase Mode

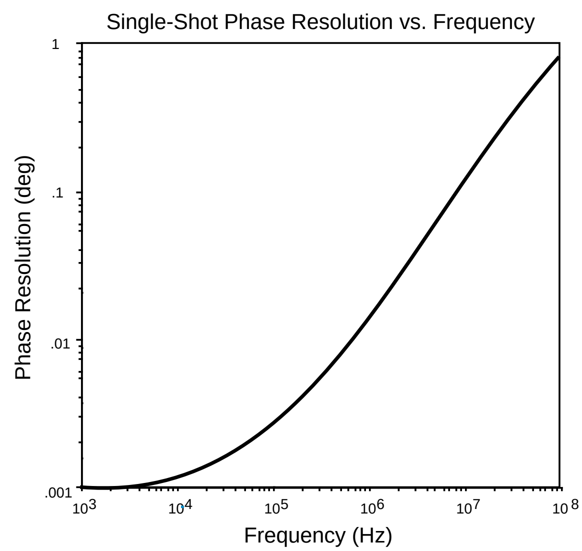
In phase mode the measurement resolution and error are given by:
where: N = number of samples averaged,

$$\text{Resolution} = \pm \left[0.001^\circ + 360 \sqrt{\frac{(25\,\text{ps})^2 + (\text{gate time} \times \text{short term stability})^2 + 2 \times (\text{trigger jitter})^2}{\text{period}^2 \times N}} \left(1 + \left(\frac{\text{phase} \times \text{period}}{360 \times \text{gate time}} \right)^2 \right) \right]$$

$$\text{Error} = \pm \left[\text{resolution} + \frac{(\text{timebase error} \times \text{time interval}) + \text{start trigger level error} + \text{stop trigger level error} + 0.5\,\text{ns}}{\text{timebase error} \times \text{period} + 1 \times 10^{-8} \times \text{period}} \times 360^\circ \right]$$

and the gate time is 10ms in internal mode

Graph 7 shows the SR620's single-shot phase resolution as a function of frequency. The resolution may be increased by averaging.



Graph 7: Single Shot Phase Resolution vs. Frequency

Count Mode

The resolution and error for count mode are:

Resolution = ±1 count

Error = ±1 count

Conclusions

The SR620 Universal Time Interval and Frequency Counter has many practical applications in the engineering and scientific environments. It can be used to measure the propagation delays of integrated circuits, the quality of a reference frequency source or any other time or frequency related quantity. Its ability to perform statistical calculations (mean, minimum, maximum, standard deviation and Allan variance) makes the SR620 applicable to almost any frequency or time related system.

Resolution, the smallest discernible difference in a measurement, is of primary interest when making comparative readings. Timebase stability, trigger noise, internal noise, etc. all contribute to limiting the resolution. The SR620 provides the user with state of the art performance with 4 ps single-shot least significant digit and 25 ps single-shot resolution. The ability to function with internal and external timebases provides the flexibility needed to balance performance and budget.

Error, the difference of a measured and actual value, is of interest when the absolute value of a measurement is important. Error consists of resolution, timebase aging, insertion delays, trigger level errors, etc.. The SR620's absolute error is typically 500 ps.

If precise timing and frequency measurements are necessary then the SR620 Universal Time Interval and Frequency Counter is the solution.

Application Note #3

About Lock-In Amplifiers

Lock-in amplifiers are used to detect and measure very small AC signals–all the way down to a few nanovolts. Accurate measurements may be made even when the small signal is obscured by noise sources many thousands of times larger. Lock-in amplifiers use a technique known as phase-sensitive detection to single out the component of the signal at a specific reference frequency and phase. Noise signals at frequencies other than the reference frequency are rejected and do not affect the measurement.

Why use a lock-in?

Let's consider an example. Suppose the signal is a 10 nV sine wave at 10 kHz. Clearly some amplification is required to bring the signal above the noise. A good low noise amplifier will have about 5 nV/√Hz of input noise. If the amplifier bandwidth is 100 kHz and the gain is 1000, then we can expect our output to be 10 µV of signal (10 nV x 1000) and 1.6 mV of broadband noise (5 nV/√Hz x √100 kHz x 1000). We won't have much luck measuring the output signal unless we single out the frequency of interest.

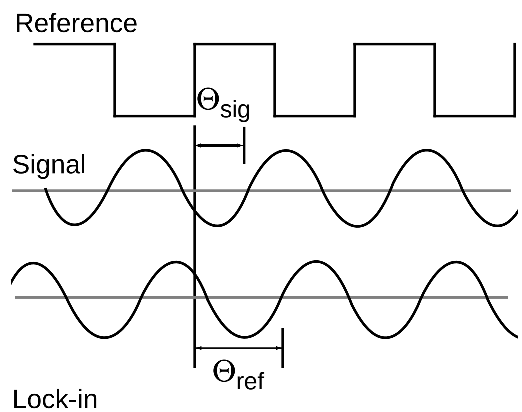
If we follow the amplifier with a bandpass filter with a Q=100 (a VERY good filter) centered at 10 kHz, any signal in a 100 Hz bandwidth will be detected (10 kHz/Q). The noise in the filter pass band will be 50 µV (5 nV/√Hz x √100 Hz x 1000) and the signal will still be 10 µV. The output noise is much greater than the signal and an accurate measurement can not be made. Further gain will not help the signal to noise problem.

Now try following the amplifier with a phase-sensitive detector (PSD). The PSD can detect the signal at 10 kHz with a bandwidth as narrow as 0.01 Hz! In this case, the noise in the detection bandwidth will be only 0.5 µV (5 nV/√Hz x √.01 Hz x 1000) while the signal is still 10 µV. The signal to noise ratio is now 20 and an accurate measurement of the signal is possible.

What is phase-sensitive detection?

Lock-in measurements require a frequency reference. Typically an experiment is excited at a fixed frequency (from an oscillator or function generator) and the lock-in detects the response from the experiment at the reference frequency. In the diagram below, the reference signal is a square wave at frequency wr. This might be the sync output from a function generator. If the sine output from the function generator is used to excite the experiment, the response might be the signal waveform shown below. The signal is Vsigsin(wrt + qsig) where Vsig is the signal amplitude, wr is the signal frequency, and qsig is the signal's phase.

Lock-in amplifiers generate their own internal reference signal usually by a phase-locked-loop locked to the external reference. In the diagram below the external reference, the lock-in's reference and the signal are all shown. The internal reference is $V_L\sin(w_L t + \theta_{ref})$.



The SR850 amplifies the signal and then multiplies it by the lock-in reference using a phase-sensitive detector or multiplier. The output of the PSD is simply the product of two sine waves.

$$V_{psd} = V_{sig}V_L\sin(\omega_r t + \theta_{sig})\sin(\omega_L t + \theta_{ref})$$

$$= 1/2\ V_{sig}V_L\cos([\omega_r - \omega_L]t + \theta_{sig} - \theta_{ref}) - 1/2\ V_{sig}V_L\cos([\omega_r + \omega_L]t + \theta_{sig} + \theta_{ref})$$

The PSD output is two AC signals, one at the difference frequency ($\omega_r - \omega_L$) and the other at the sum frequency ($\omega_r + \omega_L$).

If the PSD output is passed through a low pass filter, the AC signals are removed. What will be left? In the general case, nothing. However, if ω_r equals ω_L, the difference frequency component will be a DC signal. In this case, the filtered PSD output will be

$$V_{psd} = 1/2\ V_{sig}V_L\cos(\theta_{sig} - \theta_{ref})$$

This is a very nice signal — it is a DC signal proportional to the signal amplitude.

It's important to consider the physical nature of this multiplication and filtering process in different types of lock-ins. In traditional analog lock-ins the signal and reference are analog voltage signals. The signal and reference are multiplied in an analog multiplier, and the result is filtered with one or more stages of RC filters. In a digital lock-in such as the SR850, the signal and reference are represented by sequences of numbers. Multiplication and filtering are performed mathematically by a digital signal processing (DSP) chip. We'll discuss this in more detail later.

Narrow band detection

Let's return for a second to our generic lock-in example. Suppose that instead of being a pure sine wave the input is made up of signal plus noise. The PSD and low pass filter only detect signals whose frequencies are very close to the lock-in reference frequency. Noise signals at frequencies far from the reference are attenuated at the PSD output by the low pass filter (neither ω_{noise}-ω_{ref} nor ω_{noise}+ω_{ref} are close to DC). Noise at frequencies very close to the reference frequency will result in very low frequency AC outputs from the PSD ($|\omega_{noise}-\omega_{ref}|$ is small). Their attenuation depends upon the low pass filter bandwidth and roll-off. A narrower bandwidth will remove noise sources very close to the reference frequency, a wider bandwidth allows these signals to pass. The low pass filter bandwidth determines the bandwidth of detection. Only the signal at the reference frequency will result in a true DC output and be unaffected by the low pass filter. This is the signal we want to measure.

Where does the lock-in reference come from?

We need to make the lock-in reference the same as the signal frequency, i.e. $\omega_r = \omega_L$. Not only do the frequencies have to be the same, the phase between the signals can not change with time, otherwise cos(qsig - qref) will change and V_{psd} will not be a DC signal. In other words, the lock-in reference needs to be phase-locked to the signal reference.

Lock-in amplifiers use a phase-locked-loop (PLL) to generate the reference signal. An external reference signal (in this case, the reference square wave) is provided to the lock-in. The PLL in the lock-in locks the internal reference oscillator to this external reference, resulting in a reference sine wave at wr with a fixed phase shift of qref. Since the PLL actively tracks the external reference, changes in the external reference frequency do not affect the measurement.

Internal reference sources

In the case we've discussed, the reference is provided by the excitation source (the function generator). This is called an external reference source. In many situations the lock-in's internal oscillator may be used instead. The internal oscillator is just like a function generator (with variable sine output and a TTL sync) which is always phase-locked to the reference oscillator.

Magnitude and phase

Remember that the PSD output is proportional to V_{sig}-$\cos\theta$ where $\theta = (\theta_{sig} - \theta_{ref})$. θ is the phase difference between the signal and the lock-in reference oscillator. By adjusting θ_{ref} we can make θ equal to zero, in which case we can measure V_{sig} (cosq=1). Conversely, if θ is 90°, there will be no output at all. A lock-in with a single PSD is called a single-phase lock-in and its output is $V_{sig}\cos\theta$.

This phase dependency can be eliminated by adding a second PSD. If the second PSD multiplies the signal with the reference oscillator shifted by 90°, i.e. $V_L\sin(\omega_L t + \theta_{ref} + 90°)$, its low pass filtered output will be

$$V_{psd2} = 1/2\ V_{sig}V_{L}\sin(\theta_{sig} - \theta_{ref})$$

$$V_{psd2} \sim V_{sig}\sin\theta$$

Now we have two outputs, one proportional to $\cos\theta$ and the other proportional to $\sin\theta$. If we call the first output X and the second Y,

$$X = V_{sig}\cos\theta \qquad Y = V_{sig}\sin\theta$$

these two quantities represent the signal as a vector relative to the lock-in reference oscillator. X is called the 'in-phase' component and Y the 'quadrature' component. This is because when θ=0, X measures the signal while Y is zero.

By computing the magnitude (R) of the signal vector, the phase dependency is removed.

$$R = (X^2 + Y^2)^{1/2} = V_{sig}$$

R measures the signal amplitude and does not depend upon the phase between the signal and lock-in reference.

A dual-phase lock-in, such as the SR850 or the SR530, has two PSD's, with reference oscillators 90° apart, and can measure X, Y and R directly. In addition, the phase q between the signal and lock-in reference, can be measured according to;

$$\theta = \tan^{-1} (Y/X)$$

Digital PSD vs Analog PSD

We mentioned earlier that the implementation of a PSD is different for analog and digital lock-ins. A digital lock-in such as the SR850 multiplies the signal with the reference sine waves digitally. The amplified signal is converted to digital form using a 16 bit A/D converter sampling at 256 kHz. The A/D converter is preceded by a 102 kHz anti-aliasing filter to prevent higher frequency inputs from aliasing below 102 kHz.

This input data stream is multiplied, a point at a time, with the computed reference sine waves described previously. Every 4 µs, the input signal is sampled and the result is multiplied by both reference sine waves (90° apart).

The phase sensitive detectors (PSD's) in the SR850 act as linear multipliers, that is, they multiply the signal with a reference sine wave. Analog PSD's (both square wave and linear) have many problems associated with them. The main problems are harmonic rejection, output offsets, limited dynamic reserve and gain error.

The digital PSD multiplies the digitized signal with a digitally computed reference sine wave. Because the reference sine waves are computed to 20 bits of accuracy, they have very low harmonic content. In fact, the harmonics are at the -120 dB level! This means that the signal is multiplied by a single reference sine wave (instead of a reference and its many harmonics) and only the signal at this single reference frequency is detected. The SR850 is completely insensitive to signals at harmonics of the reference. In contrast, a square wave multiplying lock-in will detect at all of the odd harmonics of the reference (a square wave contains many large odd harmonics).

Output offset is a problem because the signal of interest is a DC output from the PSD and an output offset contributes to error and zero drift. The offset problems of analog PSD's are eliminated using the digital multiplier. There are no erroneous DC output offsets from the digital multiplication of the signal and reference. In fact, the actual multiplication is virtually error free.

The dynamic reserve of an analog PSD is limited to about 60 dB. When there is a large noise signal present, 1000 times or 60 dB greater than the full scale signal, the analog PSD measures the signal with an error. The error is caused by non-linearity in the multiplication (the error at the output depends upon the amplitude of the input). This error can be quite large (10% of full scale) and depends upon the noise amplitude, frequency, and waveform. Since noise generally varies quite a bit in these parameters, the PSD error causes quite a bit of output uncertainty.

In the digital lock-in, the dynamic reserve is limited by the quality of the A/D conversion. Once the input signal is digitized, no further errors are introduced. Certainly the accuracy of the multiplication does not depend on the size of the numbers. The A/D converter used in the SR850 is extremely linear, meaning that the presence of large noise signals does not impair its ability to correctly digitize a small signal. In fact, the dynamic reserve of the SR850 can exceed 100 dB without any problems. We'll talk more about dynamic reserve a little later.

A linear analog PSD multiplies the signal by an analog reference sine wave. Any amplitude variation in the reference amplitude shows up directly as a variation in the overall gain. Analog sine wave generators are susceptible to amplitude drift, especially as a function of temperature. The digital reference sine wave has a precise amplitude and never changes. This avoids a major source of gain error common to analog lock-ins.

The overall performance of a lock-in amplifier is largely determined by the performance of its phase sensitive detectors. In virtually all respects, the digital PSD outperforms its analog counterparts.

What does a lock-in measure?

So what exactly does the lock-in measure? Fourier's theorem basically states that any input signal can be represented as the sum of many, many sine waves of differing amplitudes, frequencies and phases. This is generally considered as representing the signal in the "frequency domain". Normal oscilloscopes display the signal in the "time domain". Except in the case of clean sine waves, the time domain representation does not convey very much information about the various frequencies which make up the signal.

A lock-in multiplies the signal by a pure sine wave at the reference frequency. All components of the input signal are multiplied by the reference simultaneously. Mathematically speaking, sine waves of differing frequencies are orthogonal, i.e. the average of the product of two sine waves is zero unless the frequencies are EXACTLY the same. The product of this multiplication yields a DC output signal proportional to the component of the signal whose frequency is exactly locked to the reference frequency. The low pass filter which follows the multiplier provides the averaging which removes the products of the reference with components at all other frequencies.

A lock-in amplifier, because it multiplies the signal with a pure sine wave, measures the single Fourier (sine) component of the signal at the reference frequency. Let's take a look at an example. Suppose the input signal is a simple square wave at frequency f. The square wave is actually composed of many sine waves at multiples of f with carefully related amplitudes and phases. A 2 Vpp square wave can be expressed as

$$S(t) = 1.273\sin(\omega t) + 0.4244\sin(3\omega t) + 0.2546\sin(5\omega t) + \ldots$$

where $\omega = 2\pi f$. The lock-in, locked to f will single out the first component. The measured signal will be $1.273\sin(\omega t)$, not the 2 Vpp that you'd measure on a scope.

In the general case, the input consists of signal plus noise. Noise is represented as varying signals at all frequencies. The ideal lock-in only responds to noise at the reference frequency. Noise at other frequencies is removed by the low pass filter following the multiplier. This "bandwidth narrowing" is the primary advantage that a lock-in amplifier provides. Only inputs with frequencies at the reference frequency result in an output.

RMS or peak?

Lock-in amplifiers as a general rule display the input signal in Volts RMS. When a lock-in displays a magnitude of 1V (rms), the component of the input signal at the reference frequency is a sine wave with an amplitude of 1 Vrms or 2.8 Vpp.

Thus, in the previous example with a 2 Vpp square wave input, the lock-in would detect the first sine component, $1.273\sin(\omega t)$. The measured and displayed magnitude would be 0.90 V (rms) ($1.273/\sqrt{2}$).

Degrees or radians?

In this discussion, frequencies have been referred to as f (Hz) and ω ($2\pi f$ radians/sec). This is because people measure frequencies in cycles per second and math works best in radians. For purposes of measurement, frequencies as measured in a lock-in amplifier are in Hz. The equations used to explain the actual calculations are sometimes written using ω to simplify the expressions.

Phase is always reported in degrees. Once again, this is more by custom than by choice. Equations written as $\sin(\omega t + \theta)$ are written as if θ is in radians mostly for simplicity. Lock-in amplifiers always manipulate and measure phase in degrees.

Dynamic Reserve

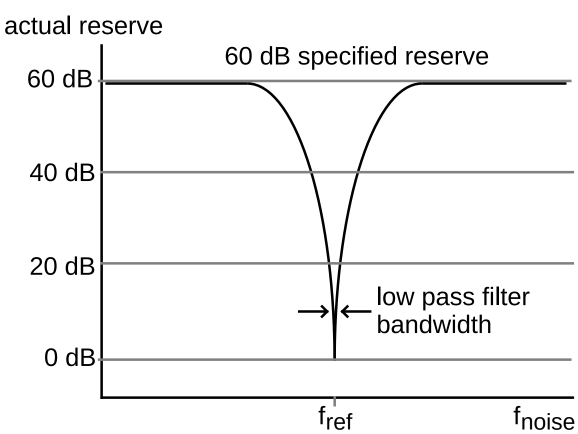
The term “Dynamic Reserve” comes up frequently in discussions about lock-in amplifiers. It’s time to discuss this term in a little more detail. Assume the lock-in input consists of a full scale signal at fref plus noise at some other frequency. The traditional definition of dynamic reserve is the ratio of the largest tolerable noise signal to the full scale signal, expressed in dB. For example, if full scale is 1 µV, then a dynamic reserve of 60 dB means noise as large as 1 mV (60 dB greater than full scale) can be tolerated at the input without overload.

The problem with this definition is the word 'tolerable'. Clearly the noise at the dynamic reserve limit should not cause an overload anywhere in the instrument — not in the input signal amplifier, PSD, lowpass filter or DC amplifier. This is accomplished by adjusting the distribution of the gain. To achieve high reserve, the input signal gain is set very low so the noise is not likely to overload. This means that the signal at the PSD is also very small. The lowpass filter then removes the large noise components from the PSD output which allows the remaining DC component to be amplified (a lot) to reach 10 V full scale. There is no problem running the input amplifier at low gain. However, as we have discussed previously, analog lock-ins have a problem with high reserve because of the linearity of the PSD and the DC offsets of the PSD and DC amplifier. In an analog lock-in, large noise signals almost always disturb the measurement in some way.

The most common problem is a DC output error caused by the noise signal. This can appear as an offset or as a gain error. Since both effects are dependent upon the noise amplitude and frequency, they can not be offset to zero in all cases and will limit the measurement accuracy. Because the errors are DC in nature, increasing the time constant does not help. Most lock-ins define tolerable noise as noise levels which do not affect the output more than a few percent of full scale. This is more severe than simply not overloading.

Another effect of high dynamic reserve is to generate noise and drift at the output. This comes about because the DC output amplifier is running at very high gain and low frequency noise and offset drift at the PSD output or the DC amplifier input will be amplified and appear large at the output. The noise is more tolerable than the DC drift errors since increasing the time constant will attenuate the noise. The DC drift in an analog lock-in is usually on the order of 1000ppm/°C when using 60 dB of dynamic reserve. This means that the zero point moves 1% of full scale over 10°C temperature change. This is generally considered the limit of tolerable.

Lastly, dynamic reserve depends on the noise frequency. Clearly noise at the reference frequency will make its way to the output without attenuation. So the dynamic reserve at fref is 0 dB. As the noise frequency moves away from the reference frequency, the dynamic reserve increases. Why? Because the low pass filter after the PSD attenuates the noise components. Remember, the PSD outputs are at a frequency of |fnoise-fref|. The rate at which the reserve increases depends upon the low pass filter time constant and roll off. The reserve increases at the rate at which the filter rolls off. This is why 24 dB/oct filters are better than 6 or 12 dB/oct filters. When the noise frequency is far away, the reserve is limited by the gain distribution and overload level of each gain element. This reserve level is the dynamic reserve referred to in the specifications.



The above graph shows the actual reserve vs the frequency of the noise. In some instruments, the signal input attenuates frequencies far outside the lock-in's operating range (f_{noise}>>100 kHz). In these cases, the reserve can be higher at these frequencies than within the operating range. While this creates a nice specification, removing noise at frequencies very far from the reference does not require a lock-in amplifier. Lock-ins are used when there is noise at frequencies near the signal. Thus, the dynamic reserve for noise within the operating range is more important.

Dynamic reserve in digital lock-ins

The SR850, with its digital phase sensitive detectors, does not suffer from DC output errors caused by large noise signals. The dynamic reserve can be increased to above 100 dB without measurement error. Large noise signals do not cause output errors from the PSD. The large DC gain does not result in increased output drift.

In fact, the only drawback to using ultra high dynamic reserves (>60 dB) is the increased output noise due to the noise of the A/D converter. This increase in output noise is only present when the dynamic reserve is increased above 60 dB AND above the minimum reserve. (If the minimum reserve is 80 dB, then increasing to 90 dB may increase the noise. As we'll discuss next, the minimum reserve does not have increased output noise no matter how large it is.)

To set a scale, the SR850's output noise at 100 dB dynamic reserve is only measurable when the signal input is grounded. Let's do a simple experiment. If the lock-in reference is at 1 kHz and a large signal is applied at 9.5 kHz, what will the lock-in output be? If the signal is increased to the dynamic reserve limit (100 dB greater than full scale), the output will reflect the noise of the signal at 1 kHz. The spectrum of any pure sine generator always has a noise floor, i.e. there is some noise at all frequencies. So even though the applied signal is at 9.5 kHz, there will be noise at all other frequencies, including the 1 kHz lock-in reference. This noise will be detected by the lock-in and appear as noise at the output. This output noise will typically be greater than the SR850's own output noise. In fact, virtually all signal sources will have a noise floor which will dominate the lock-in output noise. Of course, noise signals are generally much noisier than pure sine generators and will have much higher broadband noise floors.

If the noise does not reach the reserve limit, the SR850's own output noise may become detectable at ultra high reserves. In this case, simply lower the dynamic reserve and the DC gain will decrease and the output noise will decrease also. In general, do not run with more reserve than necessary. Certainly don't use ultra high reserve when there is virtually no noise at all.

The frequency dependence of dynamic reserve is inherent in the lock-in detection technique. The SR850, by providing more low pass filter stages, can increase the dynamic reserve close to the reference frequency. The specified reserve applies to noise signals within the operating range of the lock-in, i.e. frequencies below 100 kHz. The reserve at higher frequencies is actually higher but is generally not that useful.

Minimum dynamic reserve

The SR850 always has a minimum amount of dynamic reserve. This minimum reserve changes with the sensitivity (gain) of the instrument. At high gains (full scale sensitivity of 50 µV and below), the minimum dynamic reserve increases from 37 dB at the same rate as the sensitivity increases. For example, the minimum reserve at 5 µV sensitivity is 57 dB. In many analog lock-ins, the reserve can be lower. Why can't the SR850 run with lower reserve at this sensitivity?

The answer to this question is — Why would you want lower reserve? In an analog lock-in, lower reserve means less output error and drift. In the SR850, more reserve does not increase the output error or drift. More reserve can increase the output noise though. However, if the analog signal gain before the A/D converter is high enough, the 5 nV/√Hz noise of the signal input will be amplified to a level greater than the input noise of the A/D converter. At this point, the detected noise will reflect the actual noise at the signal input and not the A/D converter's noise. Increasing the analog gain (decreasing the reserve) will not decrease the output noise. Thus, there is no reason to decrease the reserve. At a sensitivity of 5 µV, the analog gain is sufficiently high so that A/D converter noise is not a problem. Sensitivities below 5 µV do not require any more gain since the signal to noise ratio will not be improved (the front end noise dominates). The SR850 does not increase the gain below the 5 µV sensitivity, instead, the minimum reserve increases. Of course, the input gain can be decreased and the reserve increased, in which case the A/D converter noise might be detected in the absence of any signal input.

Dynamic reserve in analog lock-ins.

Because of the limitations of their PSD's analog lock-in amplifiers must use different techniques to improve their dynamic reserve. The most common of these is the use of analog prefilters. The SR510 and SR530 have tunable bandpass filters at their inputs. The filters are designed to automatically track the reference frequency. If an intefering signal is attenuated by a filter before it reaches the lock-in input the dynamic reserve of the lock-in will be increased by that amount. For the SR510 and SR530 a dynamic reserve increase of up to 20 dB can be realized using the input bandpass filter. Of course, such filters add their own noise, and contribute to phase error, so they should only be used when necessary.

A lock-in can measure signals as small as a few nanovolts. A low noise signal amplifier is required to boost the signal to a level where the A/D converter can digitize the signal without degrading the signal to noise. The analog gain in the SR850 ranges from roughly 7 to 1000. As discussed previously, higher gains do not improve signal to noise and are not necessary.

The overall gain (AC and DC) is determined by the sensitivity. The distribution of the gain (AC versus DC) is set by the dynamic reserve.

Input noise

The input noise of the SR850 signal amplifier is about 5 nVrms/√Hz. The SR530 and SR510 lock-ins have 7 nVrms/√Hz of input noise. What does this noise figure mean? Let's set up an experiment. If an amplifier has 5 nVrms/√Hz of input noise and a gain of 1000, then the output will have 5 µVrms/√Hz of noise. Suppose the amplifier output is low pass filtered with a single RC filter (6 dB/oct roll off) with a time constant of 100 ms. What will be the noise at the filter output?

Amplifier input noise and Johnson noise of resistors are Gaussian in nature. That is, the amount of noise is proportional to the square root of the bandwidth in which the noise is measured. A single stage RC filter has an equivalent noise bandwidth (ENBW) of 1/4T where T is the time constant (RxC). This means that Gaussian noise at the filter input is filtered with an effective bandwidth equal to the ENBW. In this example, the filter sees 5 µVrms/√Hz of noise at its input. It has an ENBW of 1/(4x100ms) or 2.5 Hz. The voltage noise at the filter output will be 5 µVrms/√Hz x √2.5Hz or 7.9 µVrms. For Gaussian noise, the peak to peak noise is about 5 times the rms noise. Thus, the output will have about 40 µVpp of noise.

Input noise for a lock-in works the same way. For sensitivities below about 5 µV full scale, the input noise will determine the output noise (at minimum reserve). The amount of noise at the output is determined by the ENBW of the low pass filter. The ENBW depends upon the time constant and filter roll off. For example, suppose the lock-in is set to 5 µV full scale with a 100 ms time constant and 6 dB/oct of filter roll off. The lock-in will measure the input noise with an ENBW of 2.5 Hz. This translates to 7.9 nVrms at the input. At the output, this represents about 0.16% of full scale (7.9 nV/5 µV). The peak to peak noise will be about 0.8% of full scale.

All of this assumes that the signal input is being driven from a low impedance source. Remember resistors have Johnson noise equal to 0.13x√R nVrms/√Hz. Even a 50Ω resistor has almost 1 nVrms/√Hz of noise! A signal source impedance of 2 kΩ will have a Johnson noise greater than the SR850's input noise. To determine the overall noise of multiple noise sources, take the square root of the sum of the squares of the individual noise figures. For example, if a 2 kΩ source impedance is used, the Johnson noise will be 5.8 nVrms/√Hz. The overall noise at the SR850 input will be $[5^2 + 5.8^2]^{1/2}$ or 7.7 nVrms/√Hz.

Noise Sources

What is the origin of the noise we've been discussing? There are two types of noise we have to worry about in laboratory situations, intrinsic noise and external noise. Intrinsic noise sources like Johnson noise and shot noise are inherent to all physical processes. Though we cannot get rid of intrinsic noise sources, by being aware of their nature we can minimize their effects. External noise sources are noise sources found in the environment, such as power line noise and broadcast stations. The effect of these noises sources can be minimized by careful attention to grounding, shielding and other aspects of experimental design. We will first discuss some of the sources of intrinsic noise.

Johnson noise

Every resistor generates a noise voltage across its terminals due to thermal fluctuations in the electron density within the resistor itself. These fluctuations give rise to an open-circuit noise voltage:

$$V_{noise}(rms) = (4k\,TR\,\Delta f)^{1/2}$$

where k=Boltzmann's constant (1.38×10^{-23} J/°K), T is the temperature in °Kelvin (typically 300°K), R is the resistance in Ohms, and Δf is the bandwidth of the measurement in Hz.

Since the input signal amplifier in a lock-in typically has a bandwidth of approximately 300 kHz, the effective noise at the amplifier input is V_{noise} = 70√R nVrms or 350√R nVpp. This noise is broadband and if the source impedance of the signal is large, can determine the amount of dynamic reserve required.

The amount of noise measured by the lock-in is determined by the measurement bandwidth. Remember, the lock-in does not narrow its detection bandwidth until after the phase sensitive detectors. In a lock-in, the equivalent noise bandwidth (ENBW) of the low pass filter (time constant) sets the detection bandwidth. In this case, the measured noise of a resistor at the lock-in input, typically the source impedance of the signal, is simply:

$$V_{noise}(rms) = 0.13\sqrt{R}\,\sqrt{ENBW}\ \text{nV}$$

Shot noise

Electric current has noise due to the finite nature of the charge carriers. There is always some non-uniformity in the electron flow which generates noise in the current. This noise is called shot noise. This can appear as voltage noise when current is passed through a resistor, or as noise in a current measurement. The shot noise, or current noise, is given by:

$$I_{noise}(rms) = (2qI\,\Delta f)^{1/2}$$

where q is the electron charge (1.6×10^{-19} Coulomb), I is the RMS AC current or DC current depending upon the circuit, and Δf is the bandwidth.

When the current input of a lock-in is used to measure an AC signal current, the bandwidth is typically so small that shot noise is not important.

1/f noise

Every 10Ω resistor, no matter what it is made of, has the same Johnson noise. However, there is excess noise in addition to Johnson noise which arises from fluctuations in resistance due to the current flowing through the resistor. For carbon composition resistors, this is typically 0.1 µV-3 µV of rms noise per Volt of applied across the resistor. Metal film and wire-wound resistors have about 10 times less noise. This noise has a 1/f spectrum and makes measurements at low frequencies more difficult. Other sources of 1/f noise include noise found in vacuum tubes and semiconductors.

Total noise

All of these noise sources are incoherent. The total random noise is the square root of the sum of the squares of all the incoherent noise sources.

External noise sources

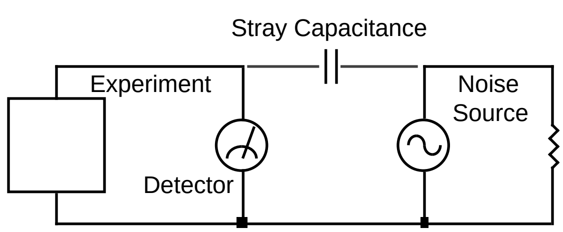
In addition to the intrinsic noise sources discussed previously, there are a variety of external noise sources within the laboratory. Most of these noise sources are asynchronous, i.e. they are not related to the reference and do not occur at the reference frequency or its harmonics. Examples include lighting fixtures, motors, cooling units, radios, computer screens, etc. These noise sources affect the measurement by increasing the required dynamic reserve or lengthening the time constant.

Some noise sources, however, are related to the reference and, if picked up in the signal, will add or subtract from the actual signal and cause errors in the measurement. Typical sources of synchronous noise are ground loops between the experiment, detector and lock-in, and electronic pick up from the reference oscillator or experimental apparatus.

Many of these noise sources can be minimized with good laboratory practice and experiment design. There are several ways in which noise sources are coupled into the signal path.

Capacitive coupling

An AC voltage from a nearby piece of apparatus can couple to a detector via a stray capacitance. Although Cstray may be very small, the coupled noise may still be larger than a weak experimental signal. This is especially damaging if the coupled noise is synchronous (at the reference frequency).



We can estimate the noise current caused by a stray capacitance by:

$$i = C_{stray}\frac{dV}{dt} = \omega C_{stray} V_{noise}$$

where ω is 2π times the noise frequency, V_{noise} is the noise amplitude, and C_{stray} is the stray capacitance.

For example, if the noise source is a power circuit, then f = 60 Hz and V_{noise} = 120 V. C_{stray} can be estimated using a parallel plate equivalent capacitor. If the capacitance is roughly an area of 1 cm^2 separated by 10 cm, then C_{stray} is 0.009 pF. The resulting noise current will be 400 pA (at 60 Hz). This small noise current can be thousands of times larger than the signal current. If the noise source is at a higher frequency, the coupled noise will be even greater.

If the noise source is at the reference frequency, then the problem is much worse. The lock-in rejects noise at other frequencies, but pick-up at the reference frequency appears as signal!

Cures for capacitive noise coupling include:

1) Removing or turning off the noise source.

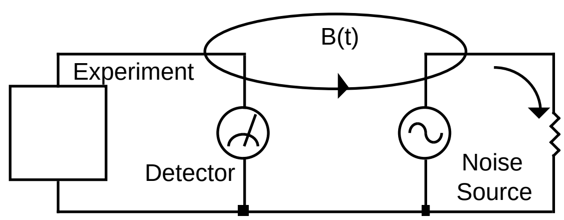
2) Keeping the noise source far from the experiment (reducing Cstray). Do not bring the signal cables close to the noise source.

3) Designing the experiment to measure voltages with low impedance (noise current generates very little voltage).

4) Installing capacitive shielding by placing both the experiment and detector in a metal box.

Inductive coupling

An AC current in a nearby piece of apparatus can couple to the experiment via a magnetic field. A changing current in a nearby circuit gives rise to a changing magnetic field which induces an emf (dØB/dt) in the loop connecting the detector to the experiment. This is like a transformer with the experiment-detector loop as the secondary winding.



Cures for inductively coupled noise include:

1) Removing or turning off the interfering noise source.

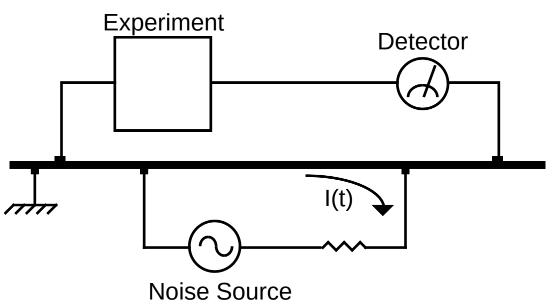
2) Reduce the area of the pick-up loop by using twisted pairs or coaxial cables, or even twisting the 2 coaxial cables used in differential connections.

3) Using magnetic shielding to prevent the magnetic field from crossing the area of the experiment.

4) Measuring currents, not voltages, from high impedance detectors.

Resistive coupling or ground loops

Currents flowing through the ground connections can give rise to noise voltages. This is especially a problem with reference frequency ground currents.



In this illustration, the detector is measuring the signal relative to a ground far from the rest of the experiment. The experiment senses the detector signal plus the voltage due to the noise source's ground return current passing through the finite resistance of the ground between the experiment and the detector. The detector and the experiment are grounded at different places which, in this case, are at different potentials.

Cures for ground loop problems include:

1) Grounding everything to the same physical point.

2) Using a heavy ground bus to reduce the resistance of ground connections.

3) Removing sources of large ground currents from the ground bus used for small signals.

Microphonics

Not all sources of noise are electrical in origin. Mechanical noise can be translated into electrical noise by microphonic effects. Physical changes in the experiment or cables (due to vibrations for example) can result in electrical noise over the entire frequency range of the lock-in.

For example, consider a coaxial cable connecting a detector to a lock-in. The capacitance of the cable is a function of its geometry. Mechanical vibrations in the cable translate into a capacitance that varies in time, typically at the vibration frequency. Since the cable is governed by Q=CV, taking the derivative, we have:

$$C\frac{dV}{dt} + V\frac{dC}{dt} = \frac{dQ}{dt} = i$$

Mechanical vibrations in the cable which cause a dC/dt will give rise to a current in the cable. This current affects the detector and the measured signal.

Some ways to minimize microphonic signals are:

1) Eliminate mechanical vibrations near the experiment.

2) Tie down cables carrying sensitive signals so they do not move.

3) Use a low noise cable that is designed to reduce microphonic effects.

Thermocouple effects

The emf created by junctions between dissimilar metals can give rise to many microvolts of slowly varying potentials. This source of noise is typically at very low frequency since the temperature of the detector and experiment generally changes slowly. This effect is large on the scale of many detector outputs and can be a problem for low frequency measurements, especially in the mHz range.Some ways to minimize thermocouple effects are:

1) Hold the temperature of the experiment or detector constant.

2) Use a compensation junction, i.e. a second junction in reverse polarity which generates an emf to cancel the thermal potential of the first junction. This second junction should be held at the same temperature as the first junction.

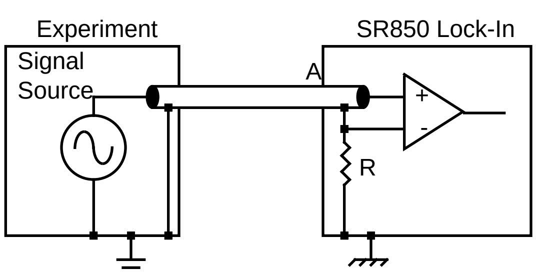
Input Connections

In order to achieve the best accuracy for a given measurement, care must be taken to minimize the various noise sources which can be found in the laboratory. With intrinsic noise (Johnson noise, 1/f noise or input noise), the experiment or detector must be designed with these noise sources in mind. These noise sources are present regardless of the input connections. The effect of noise sources in the laboratory (such as motors, signal generators, etc.) and the problem of differential grounds between the detector and the lock-in can be minimized by careful input connections.

There are two basic methods for connecting a voltage signal to the lock-in — the single-ended connection is more convenient while the differential connection eliminates spurious pick-up more effectively.

Single-Ended Voltage Connection (A)

In the first method, the lock-in uses the A input in a single-ended mode. The lock-in detects the signal as the voltage between the center and outer conductors of the A input only. The lock-in does not force the shield of the A cable to ground, rather it is internally connected to the lock-in's ground via a resistor. The value of this resistor is typically between 10Ω and 1 kΩ. The SR850 lets you choose the value of this resistor. This avoids ground loop problems between the experiment and the lock-in due to differing ground potentials. The lock-in lets the shield 'quasi-float' in order to sense the experiment ground. However, noise pickup on the shield will appear as noise to the lock-in. This is bad since the lock-in cannot reject this noise. Common mode noise, which appears on both the center and shield, is rejected by the 100 dB CMRR of the lock-in input, but noise on only the shield is not rejected at all.

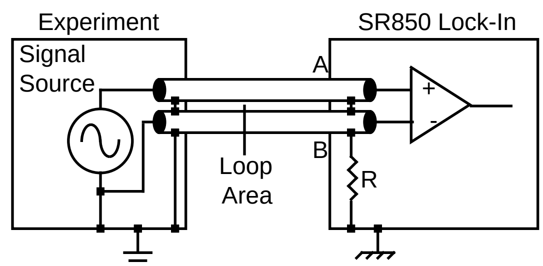


Grounds may be at different potentials

Differential Voltage Connection (A-B)

The second method of connection is the differential mode. The lock-in measures the voltage difference between the center conductors of the A and B inputs. Both of the signal connections are shielded from spurious pick-up. Noise pickup on the shields does not translate into signal noise since the shields are ignored.

When using two cables, it is important that both cables travel the same path between the experiment and the lock-in. Specifically, there should not be a large loop area enclosed by the two cables. Large loop areas are susceptible to magnetic pickup.



Grounds may be at different potentials

Common Mode Signals

Common mode signals are those signals which appear equally on both center and shield (A) or both A and B (A-B). With either connection scheme, it is important to minimize both the common mode noise and the common mode signal. Notice that the signal source is held near ground potential in both illustrations above. If the signal source floats at a nonzero potential, the signal which appears on both the A and B inputs will not be perfectly cancelled. The common mode rejection ratio (CMRR) specifies the degree of cancellation. For low frequencies, the CMRR of 100 dB indicates that the common mode signal is canceled to 1 part in 105. Even with a CMRR of 100 dB, a 100 mV common mode signal behaves like a 1 μV differential signal! This is especially bad if the common mode signal is at the reference frequency (this happens a lot due to ground loops). The CMRR decreases by about 6 dB/octave (20 dB/decade) starting at around 1 kHz.

The Lock-In as a Noise Measurement Device

Lock-in amplifiers can be used to measure noise. Noise measurements are generally used to characterize components and detectors. Remember that the lock-in detects signals close to the reference frequency. How close? Input signals within the detection bandwidth set by the low pass filter time constant and roll-off appear at the output at a frequency f=fsig-fref. Input noise near fref appears as noise at the output with a bandwidth of DC to the detection bandwidth.

The noise is simply the standard deviation (root of the mean of the squared deviations) of the measured X, Y or R. You can measure this noise exactly by recording a sequence of output values and then calculating the standard deviation directly. The noise, in Volts/√Hz, is simply the standard deviation divided by the square root of the equivalent noise bandwidth of the time constant.

For Gaussian noise, the equivalent noise bandwidth (ENBW) of a low pass filter is the bandwidth of the perfect rectangular filter which passes the same amount of noise as the real filter.

Noise Estimation

The above technique, while mathematically sound, can not provide a real time output or an analog output proportional to the measured noise. Lock-ins such as the SR510, SR530 and SR850 do provide these features, however. The quantity Xnoise is computed from the measured values of X using the following algorithm. The moving average of X is computed. This is the mean value of X over some past history. The present mean value of X is subtracted from the present value of X to find the deviation of X from the mean. Finally, the moving average of the absolute value of the deviations is calculated. This calculation is called the mean average deviation or MAD. This is not the same as an RMS calculation. However, if the noise is Gaussian in nature, then the RMS noise and the MAD noise are related by a constant factor.

The SR510, SR530, and SR850 use the MAD method to estimate the RMS noise quantities Xn, Yn and Rn. The advantage of this technique is its numerical simplicity and speed. For most applications, noise estimation and standard deviation calculations yield the same answer. Which method you use depends upon the requirements of the experiment.

Application Note #4

Signal Recovery With Photomultiplier Tubes
Photon Counting, Lock-In Detection, or Boxcar Averaging?

Which instrument is best suited for detecting signals from a photomultiplier tube? The answer is based on many factors, including the signal intensity, the signal's time and frequency distribution, the various noise sources and their time dependence and frequency distribution.

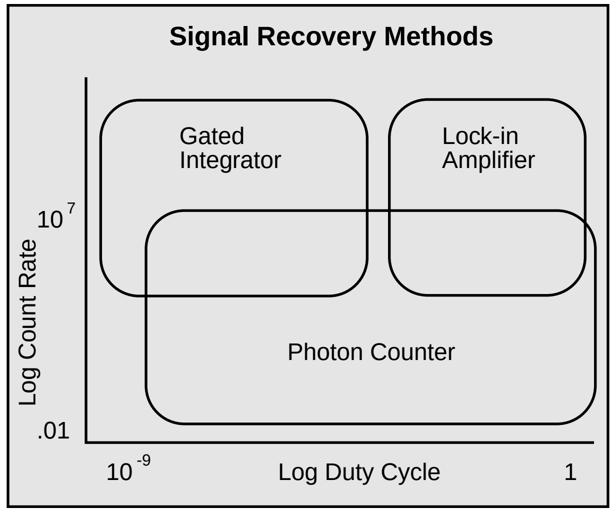
In general, the choice between boxcar averaging (gated integration) and lock-in detection (phase sensitive detection) is based on the time behavior of the signal. If the signal is fixed in frequency and has a 50% duty cycle, lock-in detection is best suited. This type of experiment commonly uses an optical chopper to modulate the signal at some low frequency. Signal photons occur at random times during the 'open' phase of the chopper. The lock-in detects the average difference between the signal during the 'open' phase and the background during the 'closed' phase.

To use a boxcar averager in the same experiment would require the use of very long, 50% duty cycle gates since the photons can arrive anywhere during the 'open' phase. Since the gated integrator is collecting noise during this entire gate, the signal is easily swamped by the noise. To correct for this, active baseline subtraction can be used where an equal gate is used to measure the background during the 'closed' phase of the chopper and subtracted from the 'open' signal. This is then identical to lock-in detection. However, lock-in amplifiers are much better suited to this, especially at low frequencies (long gates) and low signal intensities.

If the signal is confined to a very short amount of time, then gated integration is usually the best choice for signal recovery. A typical experiment might be a pulsed laser excitation where the signal lasts for only a short time (100 ps to 1 µs) at a repetition rate of 1 Hz to 10 kHz. The duty cycle of the signal is much less than 50%. By using a narrow gate to detect signal only when it is present, noise which occurs at all other times is rejected. If a longer gate is used, no more signal is measured but the detected noise will increase. Thus, a 50% duty cycle gate would not recover the signal well and lock-in detection is not suitable.

Photon counting can be used in either the lock-in or the gated mode. Using a photon counter is usually required at very low signal intensities or when the use of a pulse height discriminator to reject noise results in an increased signal to noise ratio (SNR).

As seen in the illustration, the crossover point between



analog detection and photon counting is never very distinct. At very low count rates, photon counting works well since the input discriminator virtually eliminates analog front end noise. At large count rates, analog detection works well since the analog inputs do not saturate as easily as a counter. In the middle ground, the choice should be based on SNR considerations. At best, the achievable SNR is determined by the statistical noise of the Poisson counting distribution. The analog instruments degrade the SNR due to input noise.

This applications note begins by discussing photomultiplier tubes and how to optimize their performance. The following sections discuss the SNR of the various signal recovery methods. Techniques are described which can extend the analog instruments into the 'photon counting' regime. However, these techniques have limits beyond which photon counting is preferred. Experimental data illustrating the achievable signal to noise ratios is presented.

Using Photomultiplier Tubes

Photomultiplier Tubes (PMT's) are high-gain, low noise light detectors. They can detect single photons over a spectral range of 180 to 900 nm. Windowless PMT's can be used from the near UV through the X ray region, and may also be used as particle detectors.

Photons which strike the PMT's photocathode eject an electron by the photoelectric effect. This electron is accelerated toward the first dynode by a potential of 100 to 400 Vdc. Secondary electrons are ejected when the electron strikes the first dynode, and these electrons are accelerated toward the second dynode. The process continues, typically for 8-14 dynodes, each providing an electron gain of about 4-5, to produce 106 to 107 electrons which are collected by the anode. If these electrons arrive in a 5 ns pulse into a 50 Ohm load, they will produce a 1.6 to 16 mV voltage pulse.

Geometry

There are two basic geometries for photomultiplier tubes: head-on and side-on types. The head-on type has a semitransparent photocathode, and a linear array of dynodes. The head-on types offer large photocathodes with uniform sensitivity, and lower noise. These PMT's must be operated at a higher voltage, and are usually larger and more expensive than the side-on types. Side-on types have an opaque photocathode and a circular cage of dynodes.

Spectral Response

There are a variety of materials which are used as photocathodes: the work function of the photocathode will determine the spectral response (and will influence the dark count rate) of the PMT. For photon counting, the figure of merit is the "quantum efficiency" of the PMT. A 10% quantum efficiency indicates that 1 in 10 photons which strike the photocathode will produce a photoelectron -- the rest of the incident photons will not be detected. The quantum efficiency is a function of wavelength, so select the PMT for the best quantum efficiency over the wavelength region of interest.

Gain and Risetime

When using gated detection, it is important to select a PMT with sufficient gain and short risetime. Large gains are essential to both gated integrators and photon counters. For gated integrators, the pulse risetime and width should be on the order of the gate width or less so that timing information is not lost. For photon counting, the pulse width should be smaller than the pulse-pair resolution of the counter/discriminator to avoid saturation effects. When using lock-in amplifiers, pulse risetime is unimportant while high gain extends the sensitivity of the measurement.

The criteria for a "detectable pulse" depends on the electrical noise environment of your laboratory, and the noise of your preamplifier. In laboratories with Q-switched lasers or pulsed discharges, it is difficult to reduce the noise on any coaxial cable below a few millivolts. A good, wide bandwidth preamplifier (such as the SR445) will have about 1.5 nV/√Hz, or about 25 µV rms over its 300 MHz bandwidth. Peak noise will be about 2.5 times the rms noise, so it is important that the PMT provide pulses of at least 100 µV amplitude.

Use manufacturer's specifications for the current gain and risetime to estimate the pulse amplitude from the PMT:

Amplitude (mV) = 4 x Gain (in millions)/ Risetime (in ns)

This formula assumes that the electrons will enter a 50 Ohm load in a square pulse whose duration is twice the risetime. (Since the risetime will be limited to 1.2 ns by the 300 MHz bandwidth of the preamplifier, do not use risetimes less than 1.5 ns in this formula.)

If the PMT anode is connected via a 50Ω cable to a much larger load (R>>50Ω) then the the cable termination looks like an open circuit. All of the charge in the pulse is deposited on the cable capacitance in 5 ns. The voltage on the load will be V=Q/C where C=cable capacitance. This voltage will decay exponentially with a time constant of RC where R=load resistance. In this case, the pulse height will be:

Amplitude (mV)= 160 x Gain (in millions)/ Cable C(in pF)

The current gain of a PMT is a strong function of the high voltage applied to the PMT. Very often, PMT's will be operated well above the high voltage recommended by the manufacturer, and thus substantially higher current gains (10x to 100x above specs). There are usually no detrimental affects to the PMT as long as the anode current is kept well below the rated value.

Dark Counts

PMT's are the quietest detectors available. The primary noise source is thermionic emission of electrons from the photocathode and from the first few dynodes of the electron multiplier. PMT housings which cool the PMT to about -20° C can dramatically reduce the dark counts (from a few kHz to a few Hz). The residual counts arise from radioactive decays of materials inside the PMT and from cosmic rays.

PMT's which are specifically designed for photon counting will specify their noise in terms of the rate of output pulses whose amplitudes exceed some fraction of a pulse from a single photon. More often, the noise is specified as an anode dark current. Assuming the primary source of dark current is thermionic emission from the photocathode, the dark count rate is given by :

Dark Count (kHz)= 6 x Dark Current (nA)/ Gain(millions)

PMT Base Design

PMT bases which are designed for general purpose applications are not appropriate for photon counting or fast gated integrator applications (gates < 10-20 ns). General purpose bases will not allow high count rates, and often cause problems such as double counting and poor plateau characteristics. A PMT base with the proper high voltage taper, bypassing, snubbing, and shielding is required for good time resolution and best photon counting performance.

CAUTION: Lethal High Voltages are used in PMT applications. Use extreme caution when working with these devices. Only those experienced with high voltage circuits should attempt any of these procedures. Never work alone.

Dynode Biasing

A PMT base provides bias voltages to the PMT's photocathode and dynodes from a single negative high voltage power supply. The simplest design consists of a resistive voltage divider, as shown in the figure below.

In this configuration the voltage between each dynode, and thus the current gain at each dynode, is the same. Typical current gains are three to five, so there will typically be four electrons leaving the first dynode, with a variance of about two electrons. This large relative variance (due to the small number of ejected electrons) gives rise to a large variations in the pulse height of the detected signal. Since statistical fluctuations in pulse height are caused by the low gain of the first few stages of the multiplier chain, increasing the gain of these stages will reduce pulse height variations and so improve the pulse height distribution. This is important for both photon counting and analog detection. To increase the gain of the first few stages, the resistor values in the bias chain are tapered up to increase the voltage in the front end of the multiplier chain. The resistor values are tapered slowly so that the electrostatic focusing of electrons in the multiplier chain is not adversely affected.

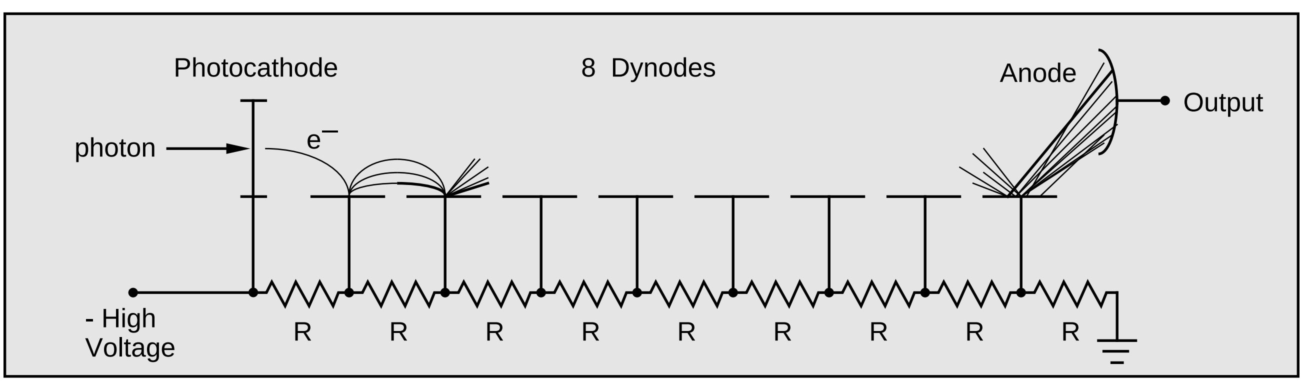
Current for the electron multiplier is provided by the bias network. Current drawn from the bias network will cause the dynode potentials to change, thus changing the tube gain. This problem is of special concern in lifetime measurements. The shape of exponential decay curves will be changed if the tube gain varies with count rate. To be certain that this is not a problem, repeat the measurement at half the original intensity.

The problem of gain variation with count rate is avoided if the current in the bias network is about 20 times the output current from the PMT's anode.

Example: If a PMT is operated so that it gives 20 mV pulses of 5 ns duration into a 50 Ohm load, then the average current at 50 MHz count rate will be 0.1 mA. If the bias resistors are chosen such that the chain current is 20 x 0.1 mA = 2 mA, then the PMT's gain will remain constant vs. count rate. If this PMT is operated at 2500 Vdc, then the power dissipated in this base is 5 Watts.

There are a few other methods to avoid this problem which do not require high bias currents. These methods depend on the fact that the majority of the output current is drawn from the last few dynodes of the multiplier:

(1) Replace the last few resistors in the bias chain with



Zener Diodes. As long as there is some reverse current through a Zener, the voltage across the diodes is nearly constant. This will prevent the voltage on these stages from dropping as the output current is increased.

(2) Use external power supplies for the last few dynodes in the multiplier chain. This approach dissipates the least amount of electrical power since the majority of the output current comes from lower voltage power supplies. However, it is the most difficult to implement.

(3) If the average count rate is low, but the peak count rate is high, then bypass capacitors on the last few stages may be used to prevent the dynode voltage from dropping (use 20x the average output current for the chain current). For a voltage drop of less than 1%, the stored charge on the last bypass capacitor should be 100x the charge output during the peak count rate. For example, the charge output during a 1 ms burst of a 100 MHz count rate, each with an amplitude of 10 mV into 50 Ohms and a pulse width of 5 ns, is 0.1 µC. If the voltage on the last dynode is 200 Vdc, then the bypass capacitor for the last dynode should have a value given by:

$$C = 100\ Q/V = 100 \times 0.1\mu C / 200V = 0.05\ \mu F$$

The current from higher dynodes is smaller so the capacitors bypassing these stages may be smaller. Only the final four or five dynodes need to be bypassed, usually with a capacitor which has half the capacitance of the following stage. To reduce the voltage requirement for these capacitors, they are usually connected in series. (See diagram below)

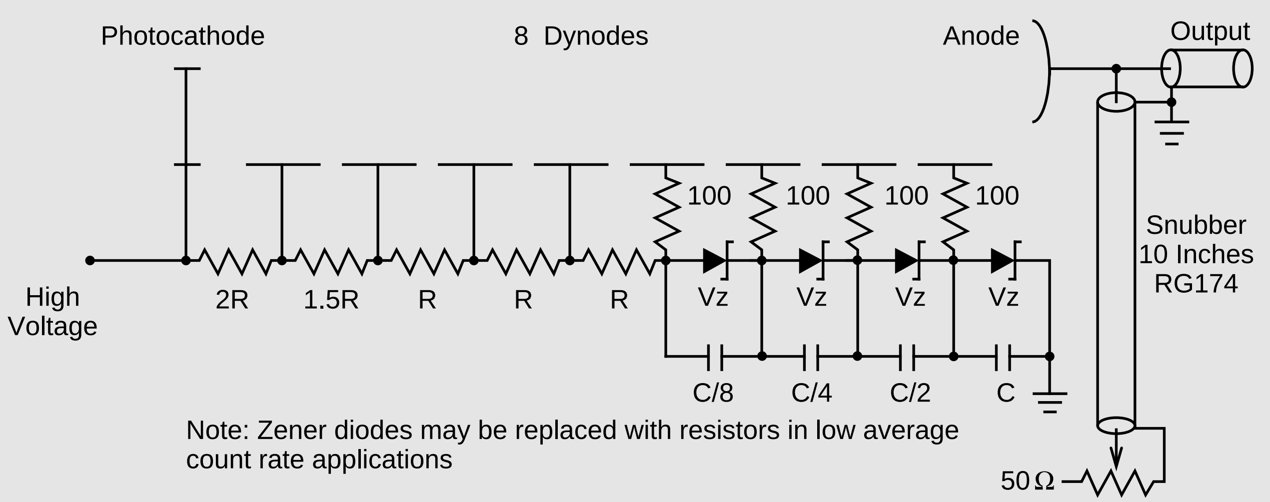
Bypassing the dynodes of a PMT may cause high frequency ringing of the anode output signal. This can cause multiple counts for a single photon or poor time resolution in a gated integrator. The problem is significantly reduced by using small resistors between the dynodes and the bypass capacitors, as shown in the diagram.

Snubbing

Snubbing refers to the practice of adding a network to the anode of the PMT to improve the shape of the output pulse for photon counting or fast gated integrator applications. This 'network' is usually a short piece of 50 Ohm coax cable which is terminated into a resistor of less than 50 Ohms. Snubbing should not be used when using a lock-in amplifier since the current conversion gain of a 50 Ohm resistor is very small.

There are four important reasons for using a snubber network:

(1) Without some dc resistive path between the anode and ground, anode dark current will charge the signal cable to a few hundred volts (last dynode potential). When the signal cable is connected to an amplifier, the stored charge on the cable may damage the front-end of the instrument. **If you decide not to use a snubber network, please install a 10-100 MΩ resistor between the anode and ground to protect your instruments.**



Note: Zener diodes may be replaced with resistors in low average count rate applications

(2) The risetime of the output current pulse is often much faster than the falltime. A snubber network may be used to sharply reduce the falltime, greatly improving the pulse pair resolution of the PMT.

(3) Ringing (with a few nanosecond period) is very common on PMT outputs (especially if the final dynode stages are bypassed with capacitors). A snubber network may be used to cancel these rings which can cause multiple counts from a single photon.

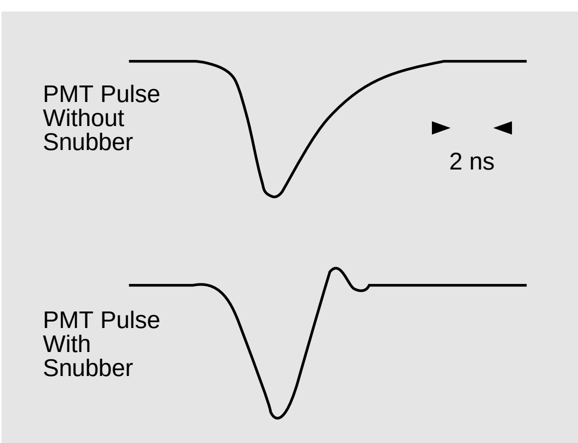
(4) The snubber network will help to terminate reflections from the input to the preamplifier.

A good starting point for a snubber network is a 10 inch piece of RG174/U coax cable with a small 50 Ohm pot connected to the end so that the terminating impedance may be adjusted from 0 to 50 Ohms. (A 10 inch cable will have a round trip time of about 5 ns -- be sure your PMT has a risetime less than this.) The other end of this cable is connected to the anode of the PMT, together with the output signal cable.

Output current pulses will split, 50% going out the signal cable, and 50% going into the snubber. If the snubber pot is adjusted to 50 Ohms there will be no reflection -- the only effect the snubber has is to attenuate the signal by a factor of two.

The reflection coefficient for a cable with a characteristic impedance R0, terminated into a resistance Rt, is given by:

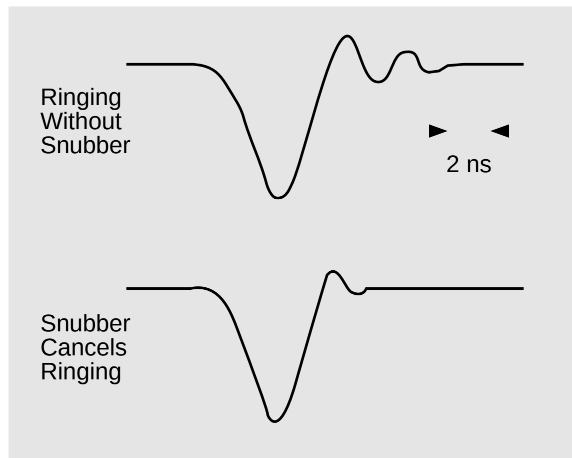
Reflection Coefficient = (Rt - R0) / (Rt + R0)



A Slow Falltime Pulse Shape
Is Improved With A Snubber

If the pot is adjusted to a value below 50 Ohms, then some portion of the signal will be inverted and reflected back toward the anode. This reflected (and inverted) signal is delayed by the round trip time in the snubber cable and sent out the signal cable. The amount of the reflection is adjusted for the best pulse shape as shown.

The round trip time in the snubber cable may be adjusted so that the reflected signal cancels anode signal ringing. This is done by using a cable length with a round trip time equal to the period of the anode ringing, as depicted below.



Anode Ringing Canceled with
Snubber Cable

Cathode Shielding

Head-on PMT's have a semitransparent photocathode which is operated at negative high voltage. **Use care so that no objects near ground potential contact the PMT near the photocathode.**

Magnetic Shielding

Electron trajectories inside the PMT will be affected by magnetic fields. A field strength of a few Gauss can dramatically reduce the gain of a PMT. A magnetic shield made of a high permeability material should be used to shield the PMT.

PMT Base Summary

(1) Taper voltage divider for higher gain in first stages.
(2) Bypass last few dynodes in pulsed applications.
(3) Use a snubber circuit to shape the output pulse for photon counting or fast gated integration.

Gated Photon Counting

Gated photon counting measures the intensity of a signal by counting the number of photons which are detected by the PMT in a given time gate. This is sometimes referred to as a 'boxcar' mode. In concept, gated photon counting is identical to gated integration except that only PMT pulses which exceed a certain threshold discriminator level are counted.

Due to the statistical nature of the secondary emission process, there is a distribution of signal pulse heights coming from the PMT. There is another distribution of noise pulse heights. Noise which results from thermionic emission from the photocathode can not be distinguished from signal, however, noise pulses from dynode thermionic emission will have a lower mean pulse height. The PMT should be operated at sufficient high voltage that the mean signal pulse height is well above the pulse height of other noise sources such as preamp noise and EMI pickup.

There are two reasons for carefully selecting the input discriminator level: 1) to improve the signal-to-noise ratio by setting the discriminator level above most of the noise pulses, but below most of the signal pulses. 2) to reduce drift: if the discriminator threshold is set to the top of the signal pulse height distribution, then small changes in the tube gain can cause a large change in the count rate.

Gain Requirement

The output of a PMT is a current pulse. This current is converted to a voltage by a load resistor. One would like to use a large resistor to get a large voltage pulse, however, in photon counting it is important to maintain a high bandwidth for the output signal. Since charge on the anode is removed by the load resistance, smaller load resistances increase the bandwidth. The bandwidth of a 10 pF anode with a 100 Ohm load is 300 MHz.

For convenience, 50 Ohm systems are usually used. The current pulse from the PMT travels down a 50 Ohm cable which is terminated by the 50 Ohm input impedance of a preamplifier. The attenuation of RG-58 coax cable at 300 MHz is about 1 dB/ 10 ft, and does not significantly degrade performance in this application.

To allow counting to 200 MHz, a preamplifier with a bandwidth which is somewhat larger than 200 MHz is required. The SR445 preamplifier has four gain of 5 amplifiers, each with 50 Ohm input impedance and a 300 MHz bandwidth. The amplifiers may be cascaded for gains of 5, 25, or 125.

The SR400 Photon Counter can detect pulses as low as 2 mV. To allow for some adjustment of the discriminator threshold and to provide better noise immunity, a more practical lower limit on pulse size is about 10 mV. The highest discriminator level which may be set is 300 mV. The preamplifier should have enough gain to amplify anode pulses to between 10 mV and 300 mV (100 mV is a good target value).

Using the result that pulse height (in mV) is about 4 times the tube gain (in millions) divided by the risetime (in ns), a PMT with a gain of 4 million and a risetime of 2 ns will provide 8 mV output pulses. Half of the pulse amplitude will be lost in the anode snubber, so a gain of 25 is required to boost the output pulses to 100 mV amplitude.

Setting the Discriminator Level

There is no exact prescription for setting the discriminator threshold: the procedure used will depend somewhat on the nature of the measurement. If dark counts are a problem then the discriminator level should be set higher than when drift is a concern. If the PMT is cooled (reducing thermionic emission) then a lower discriminator level is probably okay. If the PMT has a ring on the anode signal then the discriminator level should be set high enough so that the rings are not counted.

The 'Correct' Way

The tube should be operated at the maximum high voltage recommended by the manufacturer. Use enough preamplifier gain so that the single photon pulse height is about 100 mV. Provide enough light to the PMT for a count rate of a few megahertz. Using a 300 MHz oscilloscope, adjust the snubber termination for minimum ringing on the anode signal. Take the pulse-height spectrum of the anode signal by scanning the discriminator level and plotting count vs. discriminator level. If the PMT dark count rate is a concern, then you will also need to take the pulse height spectrum of the dark count signal. It will take much longer to take the dark count spectrum because the count rate should be much lower. The object is to find a discriminator level which is higher than the mean noise pulse height, and below the mean signal pulse height.

The 'Fast And Pretty Good' Way

This technique works very well and is particularly suited for those who do not want to make a career out of plateauing their PMT's. The PMT should be operated at (or a bit above) the recommended maximum high voltage. Provide enough illumination for a count rate of a few megahertz, and enough preamp gain to get pulse heights of about 100 mV. Using a 300 MHz oscilloscope, adjust the snubber termination impedance for the best pulse shape. Look carefully at the anode pulse shape and set the discriminator to a level which is above any ringing, but well below the mean pulse height. If there is lots of EMI or amplifier noise then increase the PMT's high voltage to increase the signal pulse height.

Signal To Noise

The probability that n photons will be detected in a time t is described by the Poisson distribution

$$P(n,t) = (Kt)n\ e\text{-}Kt / n!$$

where K is the average photon rate.

The standard deviation of any measurement is $\sqrt{N}$ where N is the number of photons detected. If many measurements of gate width T are made, the standard deviation of the data points will be $\sqrt{N_S}$ where N_S is the average number of photons detected in a single measurement. The signal to noise ratio (SNR) is simply $N_S/\sqrt{N_S} = \sqrt{N_S}$. If each data point instead consists of the sum of M measurements of gate width T, then the standard deviation will be $\sqrt{N_t}$ where $N_t = MN_S$. The SNR of this measurement is $\sqrt{Nt}=\sqrt{M} \bullet \sqrt{Ns}$ and is $\sqrt{M}$ better than the single gate measurement.

Because photon counters have a maximum count rate or pulse-pair resolution limit, they can be saturated. The probability that a photon will be counted is equal to the probability that no photon arrived in the previous time t where t is the pulse-pair resolution. If K is the total photon rate, then each photon is detected with probability e^{-Kt}. As Kt increases, eventually the number of detected photons will decrease as the PMT pulses are no longer distinct single photon pulses.

Assuming that the signal photon rate is K_S and the background or noise rate is K_b, then the detected signal is:

$$\text{SIGNAL COUNT} = K_S T\ e^{-(Ks+Kb)t}$$

where T=gate width and t=pulse-pair resolution.

The total output count is

$$\text{TOTAL COUNT} = (K_S+K_b)T\ e^{-(Ks+Kb)t}$$

The deviation of the total count is

$$(\text{TOTAL}) = [(K_S+K_b)T]^{1/2}[1 + (K_S+K_b)^2Tt]^{1/2} \times e^{-(Ks+Kb)t}$$

The SNR is just SIGNAL/ (TOTAL) or

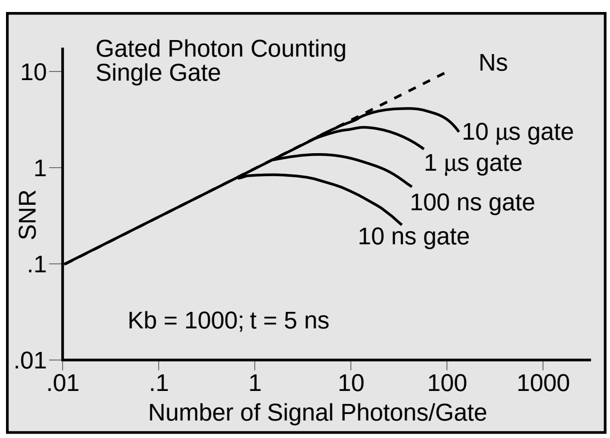
$$\text{SNR} = \frac{K_S T}{[(K_S+K_b)T]^{1/2}\ [1 + (K_S+K_b)^2Tt]^{1/2}}$$

Define:
$N_S = K_S T$ = signal photons in gate T
$N_b = K_b T$ = background photons in gate T
$n = (K_S+K_b)t$ = total photons in time t (pulse-pair resolution)

Using these definitions, the SNR is given by

$$\text{SNR} = \frac{\sqrt{N_S}}{[1 + Nb/Ns\ ]1/2\ [1 + (Ns + Nb)n]1/2}$$

The SNR is plotted below for small N_b. If N_b and n are both small, then SNR = $\sqrt{N_S}$ as discussed before. If n, the number of photons expected to occur in one pulse-



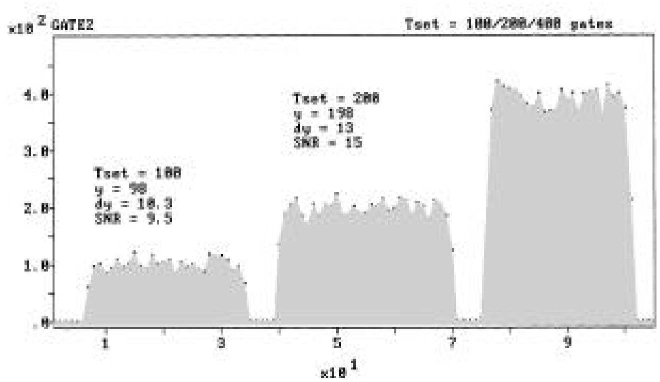
pair resolution time is large, then SNR ∝ $1/\sqrt{n}$ and decreases with increasing count rate.

If $N_b >> N_s$ then SNR ∝ $N_s/\sqrt{N_b}$. This simply means that the output noise is due to the statistics of the background counts and that the signal must exceed the deviations in the background. For example, if N_b/N_s = 10, then the SNR = 1 when the total number of signal photons counted is 10 and the total background count is 100. In this case, the counts from many gates may need to be summed together to achieve this.

Experimental data shown here demonstrates how the SNR grows as $\sqrt{N}$ where N is the total number of counts per data point. An R928 photomultiplier tube with a tapered base and snubber was used with a pulsed light source. The average pulse height from the PMT was about 15 mV into a 50Ω termination. An SR400 gated photon counter was used to collect the data. The discriminator level was set to 7 mV. The intensity of the signal was adjusted to provide an average of one signal photon per 100 ns gate. The counts from a number of gates, Tset, are summed together in each data point. Data is plotted for Tset = 100, 200, and 400 gates. In between each data set, the signal was turned off and only background data collected. For each Tset, y=average counts/data point; dy=standard deviation; SNR=y/dy.

Note that the background rate is very small and does not result in any counts when the signal is off.

A tradeoff must be made between SNR and the length of time each data point takes to accumulate. The SNR for 400 gates is twice that of 100 gates, but the 400 gate points take four times as long to acquire (SNR grows as $\sqrt{M}$ where M = # of gates). Clearly the data could be smoothed and averaged off line to improve the SNR. However, in a scanning experiment, a signal feature may only be one or two data points. Each point will be within one or two standard deviations of the average of many points and thus the signal may be characterized as having error bars of ±N.



S/N Ratio for Different Total Counts

Boxcar Averaging

Boxcar averaging, or gated integration, is an analog measurement where the signal is averaged over a short time gate and the result of many gates is averaged together. Since no discriminator is used, the variation in signal pulse height results in output deviations. Therefore, unlike the photon counter, the output noise will be greater than the √N due to counting statistics because of the pulse to pulse height fluctuations.

The signal output of a gated integrator for a signal photon rate of K_S is

SIGNAL = (1/T) K_STAeR (Volts)

where T is the gate width, A is the PMT gain, e is the electron charge, and R is the termination resistance of PMT. This assumes that the PMT output pulses are shorter in duration than T. (K_ST) is the number of photons which arrive in the gate, AeR/T is the voltage of a single pulse averaged over the gate.

The total output of the gated integrator is

OUTPUT = (1/T)(K_S+K_b)TAeR + v_n (Volts)

where Kb is the background photon rate and vn is the input voltage noise of the gated integrator.

Pulses from a PMT have an amplitude variation due to the statistics of the charge multiplication process. In addition, the number of photons detected varies as √N. Both effects are described by the Polya distribution:

$$\Delta(K_s TA = \Delta(NA) = A\sqrt{N}\left[\frac{\xi}{(\xi-1)}\right]^{\frac{1}{2}}$$

where ξ is the dynode gain per stage. As ξ becomes large, D(NA) approaches A√N which indicates that the PMT should be operated at the highest voltage (gain) possible. The pulse height variation results in output noise above the √N counting noise. For example, if A=10^7 with 14 stages, ξ = 3.16 and Δ(NA) = 1.2 A√N.

Using the above example,

$$\Delta(\text{Output}) = \left[\left(\frac{1.5}{T^2}\right)(K_s + K_b)TA^2R^2e^2 + \Delta v_n^2\right]^{\frac{1}{2}}$$

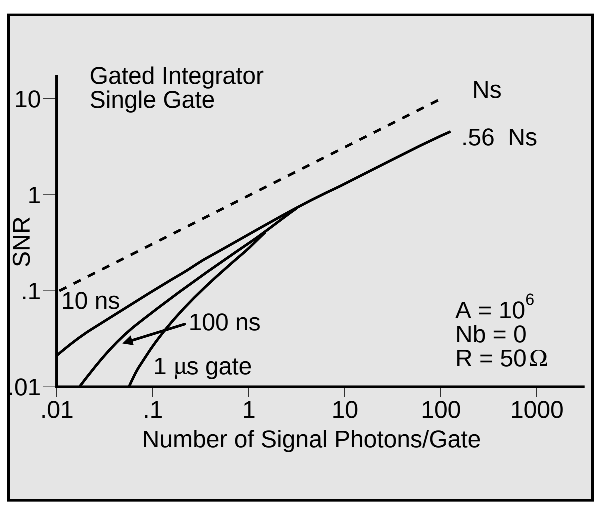
and the SNR is

$$\text{SNR} = \frac{0.8\sqrt{N_s}}{\left[1 + \frac{N_b}{N_s} + \frac{v_n^2 T}{1.5 N_s A^2 R^2 e^2}\right]^{\frac{1}{2}}}$$

where N_S=K_ST=number of signal photons which arrive in the gate, N_b=K_bT=number of background photons during the gate and v_n=input noise density (volts/√Hz). Note that the observed input voltage noise will decrease as 1/√T as the gate width is increased.

Assume a short gate such that N_b is negligible. When the statistical deviation in the number of photons observed, $\sqrt{N_S}$, which yields a noise voltage equal to $\sqrt{N_S}$(AeR/T), exceeds the averaged input noise over the gate, v_n/√T, then we are in the 'photon counting' regime. In this case, SNR ≥ 0.8√Ns. The 0.8 factor is due to the amplitude variation of the pulses. (If the tube gain is 10^6, then the factor is 0.56). However, unlike photon counting, there are no saturation effects at high count rates. In addition, since there is no discriminator, smaller photon pulses contribute to the output, increasing N_S and offsetting some of the pulse height variations.

When the input noise dominates, v_n/√T >> $\sqrt{N_S}$(AeR/T), and SNR <0.8√Ns. Note that for a constant number of photons, making the gate longer decreases the SNR. This is because the noise voltage will decrease as 1/√T(Gaussian) while the signal is decreasing by 1/T (linear averaging).

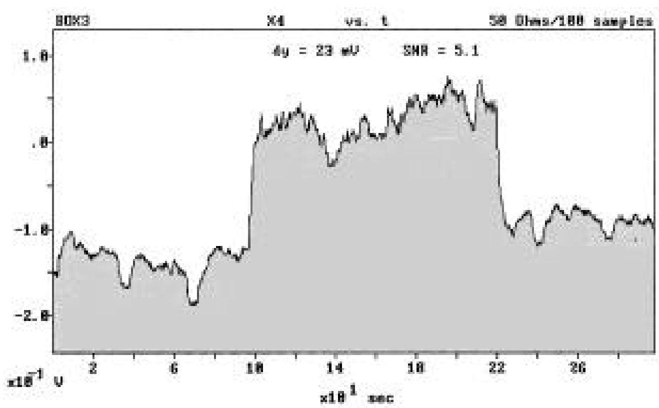


The SNR is plotted above for a single gate. In all cases, averaging over M gates increases SNR by $\sqrt{M}$. This can be seen be replacing N_S with MN_S and T with MT.

For large signals, it is clear that while the photon counter will saturate, the boxcar averager will work just fine. The fact that the achievable SNR is less than $\sqrt{N}$ is not important since the only way to improve the SNR is to attenuate the signal to a few photons per gate and count photons for many gates. The inconvenience of the longer measurement times usually far offsets the small gain in SNR which would result from counting.

For small signals, from one to much fewer than one photon per gate, and long gates, greater than 10 ns, photon counting is usually better.

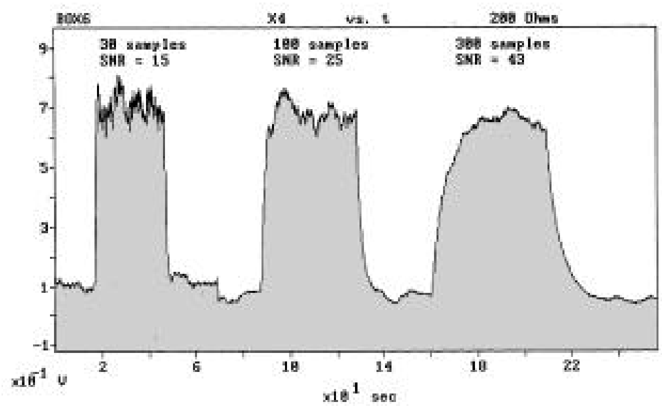
Experimental data acquired with an SR250 boxcar averager is shown below. The same signal and PMT which was used in the photon counting example was used here. The sensitivity was set to 1 V/ 5 mV and the average photon pulse had an amplitude of 15 mV. An average of 1 photon per 100 ns gate is collected and the averaging is over 100 gates.



S/N Ratio for Boxcar Integration

The SNR is only 5:1 while in the 100 gate, photon counting case it was 9.5:1. In fact, the boxcar is even worse. Since the 100 gate average is exponential, the number of gates required for the signal to go from zero to full scale is 300-500, while in the photon counting case the signal reaches full scale after 200 gates. Because of this, in a scanning experiment the photon counter could scan twice as fast and achieve twice the SNR.

Since the output noise is independent of whether there is any signal, the boxcar SNR is dominated by the input voltage noise. The SNR can be improved by increasing the size of the PMT pulses. This can be done with an SR240 preamplifier which has a lower input noise than the boxcar averager. When the gate is long, simply increasing the termination resistance can go a long way to increasing the SNR. Choose a resistance that does not widen the PMT pulses beyond about half of the gate width, otherwise, timing information will be lost. In this experiment, a 200Ω resistance increased the pulse amplitude by four and brought the SNR into the 'photon counting' regime as shown below.



S/N Ratio for 200Ω Terminating Impedance

The baseline noise is now negligible compared to the signal and the 100 gate data now has a SNR of 25. Increasing the averaging increases the SNR by $\sqrt{\text{samples}}$. As discussed before, averaging over 100 samples should be compared to photon counting for about 30 samples since the signal rises to full scale in fewer data points for photon counting.

Synchronous Photon Counting

If a signal is fixed in frequency and has a 50% duty cycle, then synchronous photon counting, or photon counting in a 'lock-in' mode, can be used. These signals usually result from the use of an optical chopper or other periodic excitation. The photon counter uses two counters and one PMT. Both counters use the one PMT as their signal source. The A counter counts pulses during the 'open' phase of the chopper and thus counts signal plus background. The B counter counts pulses only during the 'closed' cycle of the chopper and, only counts the background. The difference between the two counts, A-B, is the signal. Accumulating data over many cycles is required to measure the signal since the background rate usually far exceeds the signal rate.

Assume that the total photon rate is small (K << 100 MHz) such that saturation can be ignored. Then the signal count is

$$\text{SIGNAL COUNT} = K_S T_1$$

where T_1= the open cycle=1/2 period of the chopper frequency.

The A-B count is

$$\text{A-B COUNT} = (K_S+K_b)T_1 - K_b T_2$$

where T_2 = the closed cycle $\approx T_1$.

The noise in the A-B count is

$$\Delta(A\text{-}B) = \sqrt{(N_S + 2N_b)}$$

where $N_S = K_S T_1$ = the average number of signal photons counted during T_1, and $N_b=K_b T_1=K_b T_2$ = the average number of background photons counted during T_1 or T_2. Even though $K_b T_1=K_b T_2$, the noise in the output due to K_b is not zero since, in a given measurement, the number of background counts detected during T_1 differs from the number detected during T_2 by $\sqrt{2}\sqrt{N_b}$ ($\sqrt{N_b}$ is the uncertainty in N_b during each cycle.) The cancellation of the background improves as the number of background counts detected increases and is the dominant factor in the SNR.
The SNR is:

$$\text{SNR} = \frac{N_S}{[N_S + 2N_b]^{1/2}}$$

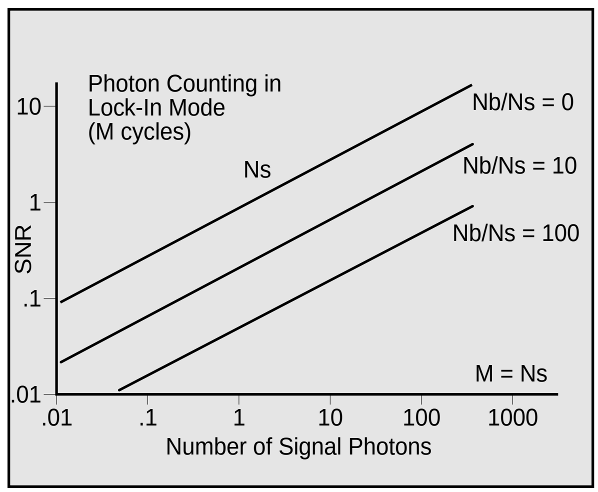
If $N_b << N_S$, then SNR = $\sqrt{N_S}$ as expected.

If Nb >> N_S, then

$$\text{SNR} = \frac{N_S}{[2Nb]^{1/2}}$$

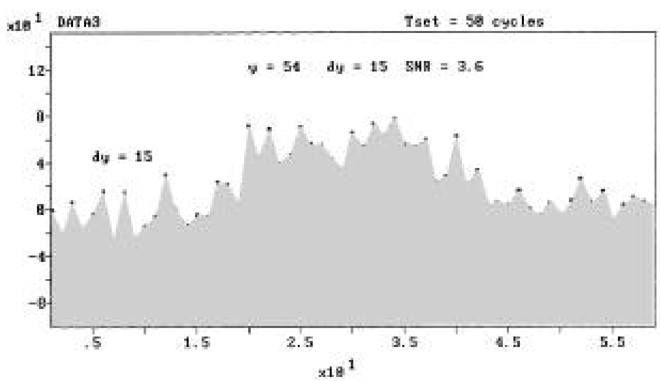
In this case, the noise is the uncertainty in the measurement of the background rate. For example, if N_b/N_S = 10, then the SNR=1 only after N_S =20 and N_b=200. This can be achieved by adding the results of many cycles together or increasing the count rate.

A plot of SNR vs total number of signal photons counted for several values of N_b/N_S is shown below. For a given N_S = average number of signal photons per cycle, the larger the background, the longer the data acquisition will take.

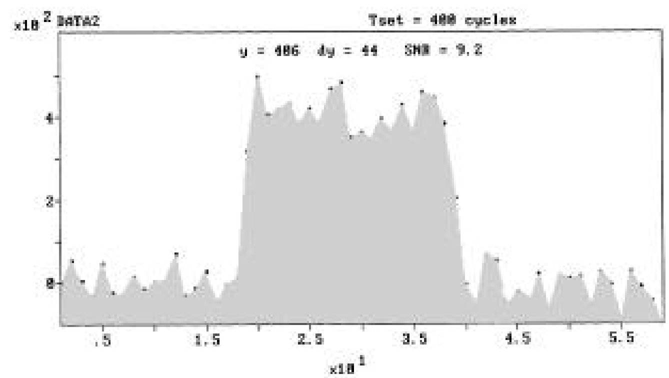


Experimental data is presented in the following figures which show that the SNR is determined by the background rate when $N_b>N_S$. The PMT and signal source from the gated experiments was used. The signal was gated on for 10 ms at a repetition rate of 50 Hz (50% duty cycle). The signal amplitude was adjusted to provide an average of 1 signal photon per cycle (50 photons/sec). The background rate of dark counts was about 100 counts/sec. The SR400 photon counter is configured for counting A-B where counter A is gated on during the signal phase and counter B is gated on during the background phase. Both counters were gated on for 9 ms.

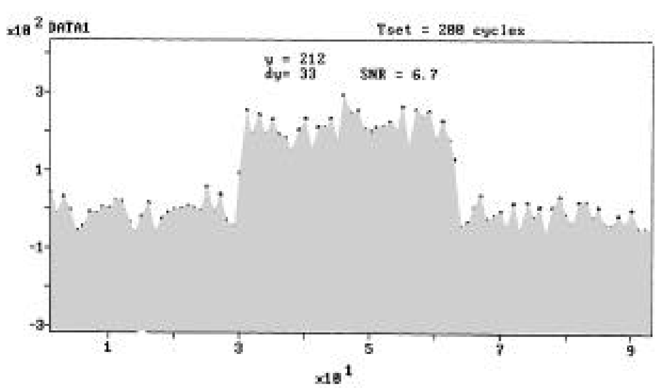
The data plots show data accumulated for 50, 200, and 400 cycles. Note that the noise is basically independent of whether the signal is on or off. This is because $2N_b > N_s$. The SNR improves as $\sqrt{M}$ where M=number of cycles as the deviations in the background count become less than the signal itself.



Synchronous Photon Counting: 50 Cycles



Synchronous Photon Counting: 400 Cycles



Synchronous Photon Counting: 200 Cycles

Lock-In Detection

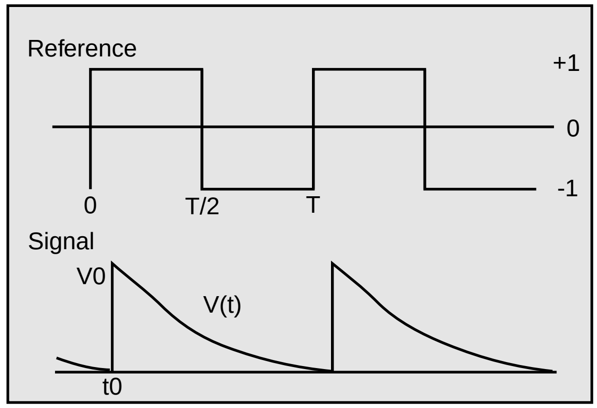
Can a lock-in amplifier detect a single photon per reference cycle? In the previous photon counting experiment, the signal was about one photon per cycle at 50 cycles/second. This is a PMT current of 50Ae where A is the PMT gain. For $A=10^7$, the current is 80 pA, which is well within the capability of an SR510/530 lock-in amplifier.

A lock-in amplifier detects current by using a large resistor or current amplifier to convert the signal current into a voltage. When detecting single photon signals from a PMT, a well chosen termination resistor can provide signal to noise ratios governed entirely by the counting statistics of the photons.

Typically, a PMT output consists of a coaxial cable terminated by a 50Ω resistor. The output voltage appears across the resistor. Since the output cable is terminated in its characteristic impedance, the output voltage pulse will be $Vo \approx AeR/\Delta t$ where Δt is the pulse width of the PMT. For $A=10^7$ and $\Delta t=5$ ns, $Vo \approx 16$ mV. Now assume that 1 photon is detected per cycle at exactly the same time during each cycle. In the time domain, the signal is a periodic series of delta functions spaced by the reference period T where $T \gg \Delta t$. In the frequency domain, the signal spectrum is a series of delta functions spaced by 1/T and extending from dc out to $1/\Delta t$. In the case where $\Delta t=5$ ns, the spectrum extends to 200 MHz. A lock-in amplifier which is locked to f=1/T is not suited to detecting this signal, not because of the amplitude, but because of the frequency spectrum.

Now suppose the output of the PMT is terminated by a high resistance R>>50Ω. Because the cable is terminated in a high impedance, the cable can be modeled solely by its capacitance C. The charge from the PMT pulse is deposited on the capacitance in time t. The voltage on the cable will be $V_o=Ae/C$. The charge then bleeds away through R over many time constants τ = RC. Thus, a photon arriving at time t=0 results in an output voltage waveform $V(t)=Voe^{-\tau/t}$. Note that the amplitude of the pulse does not vary with R. The large R serves to lengthen the pulse width and thereby change the frequency spectrum of the pulse. The frequency spectrum now has components from DC to ≈1/RC. If C=100 pF and $R=10^7\Omega$, the pulse amplitude is 1.6 mV. The frequency spectrum extends from DC to ≈1 kHz and has a larger component at 50 Hz for the lock-in to detect.

The signal output of the lock-in can be estimated by considering a square wave multiplier and a periodic photon train at the reference frequency as shown below.



The DC output of the lock-in over one cycle is

$$S(t_0) = \frac{V_o}{T}\int_0^T e^{-t/\tau}dt \qquad \text{(Volts)}$$

Since real photons arrive at a random t0 between 0 and T/2, the response for a random photon is

$$S = \frac{2}{T}\int_0^{\frac{T}{2}} S(t_0)dt_0 \qquad \text{(Volts)}$$

$$S = \frac{\tau}{T}(1-e^{-\frac{T}{\tau}}) - \frac{4\tau^2}{T^2}(1-e^{-\frac{T}{2\tau}})^2 \qquad \text{(Volts)}$$

S is the response for an average of 1 photon arriving at a random time during each reference cycle. If $\tau \gg T$, S≈0 because the RC time constant of the PMT output attenuates signals at the reference frequency. If $\tau \ll 0$, S≈0 because the signal extends to frequencies far greater than the reference frequency.

S maximizes for $\tau = T/6$ at which point S=0.065Vo. The factor 0.065 is due to the fact that the signal has many frequency components other than 1/T as well as a randomly shifting phase. Thus, the signal output of the lock-in is

$$\text{SIGNAL} = 0.065\, N_S Ae/C \qquad \text{(Volts)}$$

where N_S is the average number of signal photons per cycle and C is the cable capacitance. For T=20 ms (50 Hz), C=100 pF, R=30 MW, Ns=1, and A=10^7, the signal will be 1 mV.

The shunt resistor method is simple and easy to implement, however, phase information is lost. In many experiments, phase is not important. When phase measurements must be made, a current preamplifier is used instead. The current preamplifier eliminates the cable capacitance and the bandwidth of the amplifier is determined by the capacitance of the current gain resistor. Since this capacitance is much smaller, the time constant of the amplifier output pulse is much shorter than the case discussed above. Assuming that the reference period is much longer than this time constant, T >> τ, then the above formula applies and $S \approx Vo\tau/T$ and Vo=Ae/C where C is the capacitance of the current conversion resistor. Since τ = RC where R is the current gain, then S=AeR/T which is just the average current times the current gain resistor. The output signal will be

$$\text{SIGNAL} = N_S AeR/T \quad \text{(Volts)}$$

For the conditions stated above, the signal will be 2.4 mV. A disadvantage of this approach is that the output of the current preamplifier is a pulse of much greater amplitude and shorter duration than the simple shunt resistor. This requires the use of a higher dynamic reserve so that the high frequency components of the pulse do not overload the amplifier.

The output of the lock-in when there are background photons is

$$\text{OUTPUT}=(0.065Ae/C)(N_S+N_1-N_2) + v_n \quad \text{(Volts)}$$

where N_1 is the number of background photons detected during the open cycle and N_2 is the number detected during the closed cycle. v_n is the noise voltage of the current conversion resistor.

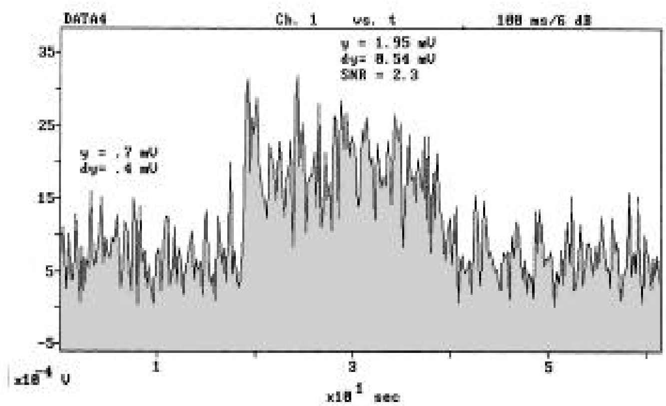
The signal to noise ratio is

$$\text{SNR} = \frac{N_s}{\left[N_s + N_b + \dfrac{v_n^2}{\Delta T(0.065Ae/C)^2}\right]^{\frac{1}{2}}}$$

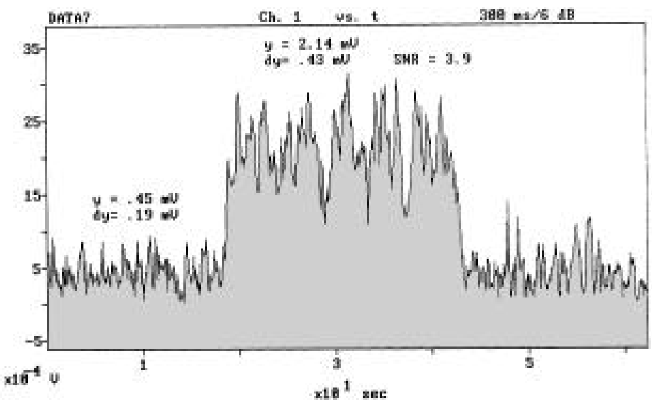
where N_S and N_b are the number of signal and background photons which occur during a lock-in output time constant ΔT and v_n=0.13$\sqrt{R}$ nV/$\sqrt{\text{Hz}}$ is the Johnson noise density of the conversion resistor. If N_S or N_b is large, then the SNR is identical to the photon counting case described in the previous section where data is accumulated for M cycles and $\Delta T \approx M$ reference cycles. If v_n dominates, then the SNR is worse than pure counting statistics. However, the Johnson noise of large resistors is very small and does not limit many measurements. For example, a 30 MΩ resistor has a noise voltage of 2 µV (for ΔT=1 s) while the signal due to 50 photons/sec is 1 mV. In fact, in this example, as long as 1 background photon is detected per second, the SNR will be dominated by counting statistics. In all cases, the SNR increases as $\sqrt{\Delta T}$ where ΔT is lock-in time constant. This is because more photons are detected and the statistical counting noise is reduced.

Experimental data is shown here. The signal source and PMT were the same as in the previous photon counting discussion. A 30 MΩ resistor was used to terminate the PMT output. An SR530 dual phase lock-in was used to measure signal magnitude. The resulting output was about 2 mV which agrees well with the calculations above for an average of 1 photon per cycle. When the PMT high voltage was off, there was no measurable output noise as expected. In all cases, the SNR is dominated by the background count rate which exceeded the signal rate by 2-3.

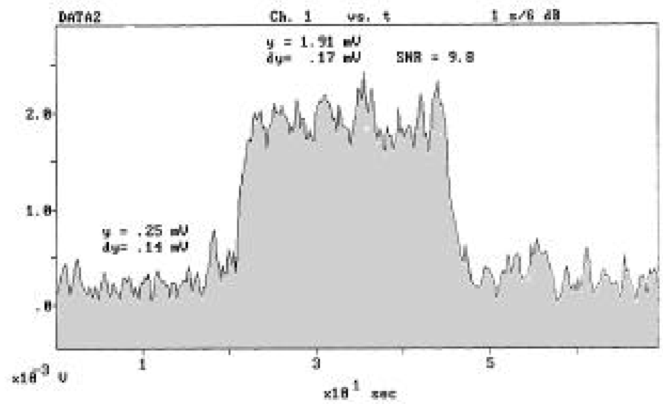
As seen from the data, a signal of 50 photons/sec at a reference of 50 Hz is easily detected with a lock-in amplifier to the same SNR as a photon counter.



Lock-In Amplifier Data: 100 ms Time Constant



Lock-In Amplifier Data: 300 ms Time Constant



Lock-In Amplifier Data: 1 s Time Constant

Conclusions

In general, large signals require the use of analog instruments such as boxcar averagers and lock-in amplifiers. While it is true that the theoretical achievable signal to noise ratio may not be as good as the counting statistics, the practical matter is that large signals take a lot less time to measure to a given SNR level.

When the signal is low, less than 1 photon per gate or cycle, the analog instruments, with the appropriate technique, can achieve photon counting signal to noise ratios. When the signal is much lower, photon counting is required.

In most experiments, the key to optimizing the measurement will lie in factors other than signal intensity. In all cases, the PMT quantum efficiency, gain, and noise are the most important factors. The initial gain from the PMT can never be replaced as well with an amplifier.

Low background or dark count rates are essential in low level measurements. External noise pickup in signal cables when lasers trigger or unstable background count rates (such as from an unstable glow discharge), can result in large fluctuations in signal amplitude far in excess of the counting statistics. These experimental factors can be the most important considerations when choosing an instrument.

For further reading;

Photomultiplier Handbook (publication PMT-62), RCA Corp., 1980.

The Art of Electronics, Horowitz and Hill, Cambridge University Press, Cambridge, 1982.

Application Note #5

Direct Digital Synthesis–Impact on Function Generator Design

Introduction

Function generators have been around for a long while. Over time, these instruments have accumulated a long list of features. Starting with just a few knobs for setting the amplitude and frequency of a sinusoidal output, function generators now provide wider frequency ranges, calibrated output levels, a variety of waveforms, modulation modes, computer interfaces, and, in some cases, arbitrary functions.

The many features added to function generators have complicated their design and increased their cost. There is an opportunity for a radical re-design of the familiar function generator using direct digital synthesis (DDS).

DDS provides remarkable frequency resolution and allows direct implementation of frequency, phase and amplitude modulation. These features which were 'tacked-on' to function generators are handled in a clean, fundamental way by DDS.

Direct Digital Synthesis

Many of the concepts of DDS are illustrated by the way in which a sine wave is generated. Figure 1 shows a block diagram of a simple DDS function generator. The sine function is stored in a RAM table. The RAM's digital sine output is converted to an analog sine wave by a DAC. The steps seen at the DAC output are filtered by a lowpass filter to provide a clean sinewave output.

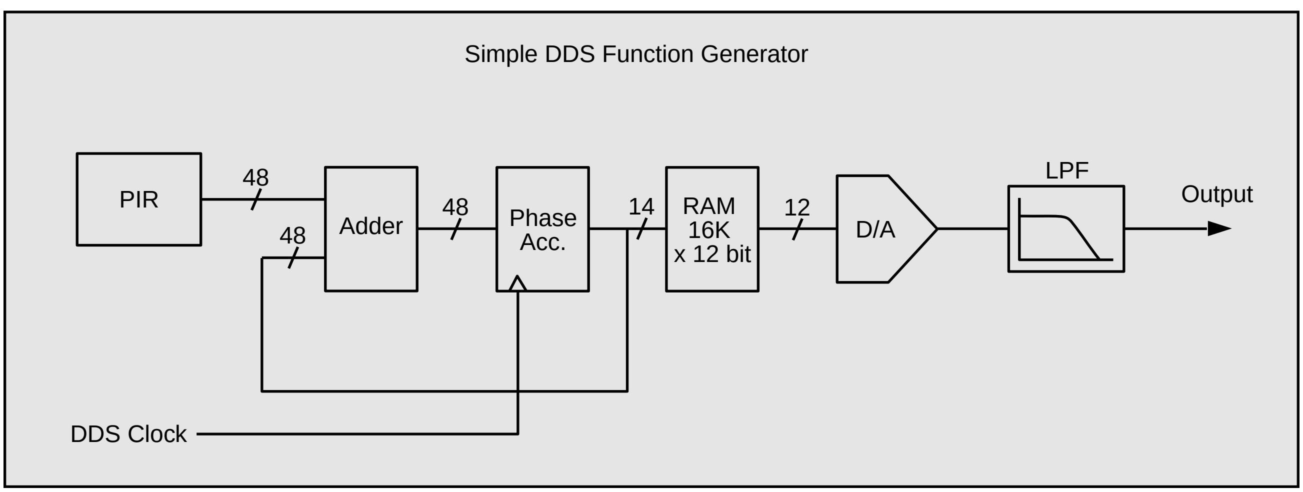
The frequency of the sine wave depends on the rate at which addresses to the RAM table are changed. Addresses are generated by adding a constant stored in the phase increment register (PIR) to the phase accumulator. Usually, the rate of additions is constant, and the frequency is changed by changing the number in the PIR.

The frequency resolution depends on the number of bits in the PIR. If the PIR, adder, and phase accumulator support 48-bit additions, then the fractional frequency resolution is one part in 247, or about one part in 1014. That means a 48-bit DDS function generator can provide better than 1 µHz resolution on a 10 MHz output.

Some Details

There are a few more details which need to be addressed in order to understand DDS in this application. Questions about sample rate, RAM size, DAC resolution, filter characteristics, and spectral purity of the output must be answered.

Figure 1. Simple DDS Block Diagram



Samples per Cycle

Our intuition might suggest that a large number of samples are required for each cycle of the sine wave to achieve good spectral purity of the output. A sketch of a sine which is approximated by a small number of samples per cycle hardly looks like a sine wave. Remarkably, only about 3 samples are required during each cycle. In fact, if we could make an arbitrarily sharp lowpass filter, we would need only two samples per cycle.

To motivate this, consider the case where we have four samples per sine cycle. This situation is shown in Figure 2. The sampled sine is reduced to a pulse train (or a square wave, if we started sampling at 45 degrees instead of at 0 degrees).

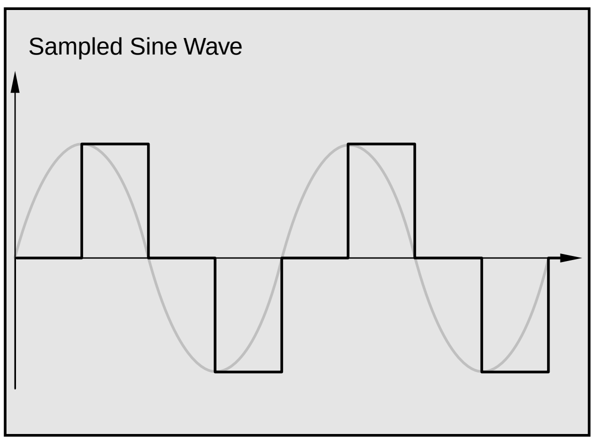


Figure 2. Sampling a Sine Four Times per Cycle.

The Fourier spectra of this pulse train has components at f, 2f, 3f.. etc. If we can arrange the lowpass filter to eliminate the harmonic components of the pulse train, then we are left with the fundamental, a pure sine wave at frequency f.

In the more general case, generating an output at f by sampling at a rate of f_s, the lowest frequency Fourier component at a frequency of f_s-f. This simple result becomes the basis of the lowpass filter specification: the filter should pass f but stop f_s-f.

Filters

Figure 3 shows a lowpass filter transfer function. As we have seen, the filter must pass the highest frequency which we wish to generate (f_{max}), but must begin their stop-band at f_s-f_{max}. Steep rolloff filters with high stop-band attenuation are hard to build. A reasonable compromise in this trade-off occurs when $f_{max}=f_s/3$. This allows the filter a one octave transition band.

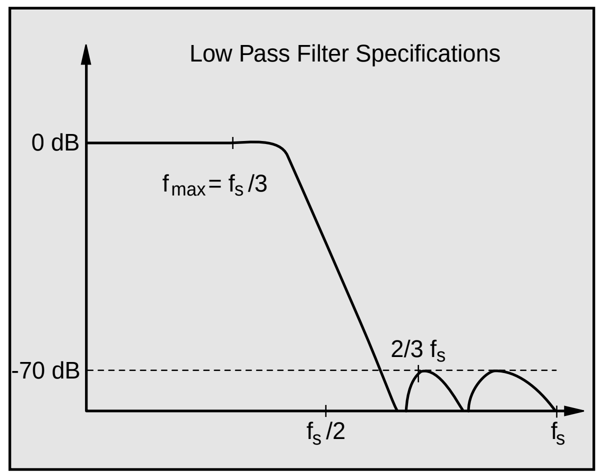


Figure 3. Lowpass Filter for DDS Outputs.

What stopband attenuation is needed? This depends on the spurious component specification of the output. A typical specification for a function generator application would be -70 dBc.

Cauer (elliptic) filters are a good choice for this application. They have fast transition bands, and may be designed with very low ripple in the pass-band. The specification for this example is met by a ninth degree Cauer filter.

Bessel Filters

While Cauer filters are the best choice for CW applications, they are unusable for arbitrary waveform generation. In the time domain, Cauer filters have a very nasty overshoot. A much better choice for arbitrary waveforms (or ramps and triangles) is the Bessel filter. The Bessel filter has a slower rolloff when compared to the Cauer filter, but it is nearly phase-linear. The lack of dispersion in a phase-linear filter will preserve the pulse shape and prevent any ringing in the time domain. A seventh degree Bessel filter with a -3 dB cutoff of $f_c = f_s/4$ is a good choice for filtering arbitrary waveforms. This filter will exhibit an output risetime of $0.35/f_c$.

DAC and RAM Requirements

Big, fast RAMs and high speed, high resolution DACs have made DDS a viable technology for function generator applications. How big, how fast, and what resolutions are required?

As we have seen, a maximum practical output frequency is $f_s/3$. So the DDS phase accumulator, RAMs, and DACs must run at three times the maximum desired output frequency.

The DAC resolution depends on the spurious component specification for the output (or the desired arbitrary waveform resolution). The DAC's quantization error and non-linearities lead to spurious outputs. To get a rough idea of the magnitude of the spurious frequency component, realize that the difference between the actual output of the DAC and the desired sine value is the source of these spurious output components. So a 12-bit DAC which is linear and monotonic to 2 LSBs will have output errors on order one part in 2048, or about –66 dB.

A short RAM table is another way to get the wrong value out of the DAC. To avoid 'phase quantization noise', there should be two more bits of address to the RAM than bits in the DAC.

Extending Frequency Range

The frequency range of the DDS output may be extended by a variety of techniques. Depending on which technique is used, some of the advantages of DDS may be lost. Just as with more conventional frequencies synthesizers, the DDS output may be doubled, mixed with other fixed sources, or used as a reference inside of a phase locked loop.

Modulation Techniques

The power and elegance of DDS are most apparent when a modulated source is required. The frequency of the output may be changed instantly to any frequency from dc to f_{max} by simply changing the number in the phase increment register. Figure 4 shows the block diagram of a DDS phase accumulator with programmable modulation capabilities.

This phase accumulator, which has been optimized for function generator applications, has two phase increment registers, PIRA and PIRB. A 48-bit wide multiplexer can switch between the PIRs in a single clock. The modulation processor can modify the PIRs at a rate of up to 10 million bytes per second, filling one PIR while the other is used as an input to the adder.

Complex modulation programs may be stored in the modulation RAM. This RAM contains op-codes and data for the modulation processor. Frequency scans illustrate the operation of this processor. When programed for a log frequency sweep, a list of up to 4000 discrete frequencies are stored in the modulation RAM by the host system. The modulation processor modifies PIRA while the adder is using PIRB and vice-versa.

More complex modulation programs may be stored, such as frequency modulation by any arbitrary function, linear or log sweeps, frequency hopping, etc. Phase modulation is easily done by programming PIRA with the nominal frequency, and using PIRB, which contains the nominal phase increment plus any desired phase shift, for a single clock cycle.

Wide frequency or phase deviations are no problem. Any phase or frequency hop may be programmed and executed in a single clock. And, since the PIRs may be modified very quickly, modulation frequencies up to several hundred kHz are possible.

In fact, arbitrary modulation programs may be stored. This feature allows the function generator to be used for modem testing, frequency agile communications, bit error rate determination, etc.

Amplitude Modulation

There are two approaches for amplitude modulation of the output waveform. Either the digital outputs from the RAM or the analog output from the DAC may be multiplied by the desired amplitude. The later approach is better for function generators so that either an internal or external source may be used for amplitude modulation.

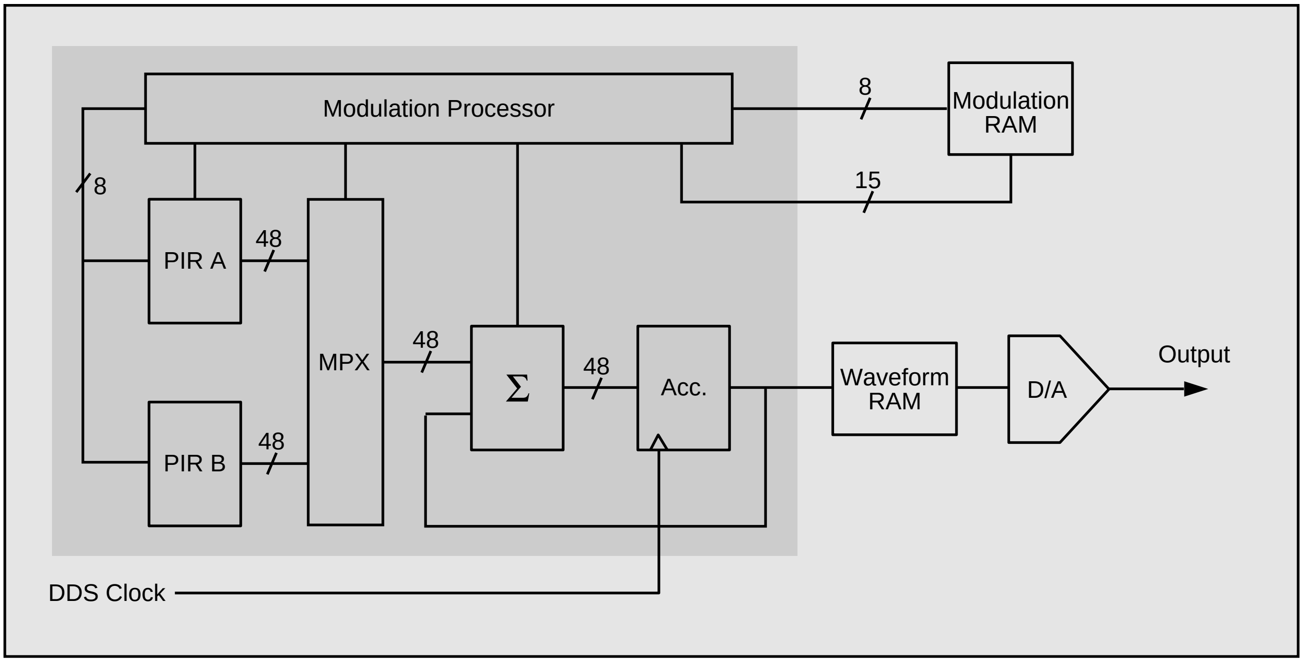


Figure 4. DDS Phase Accumulator with Modulation Processor.

Arbitrary Functions

One of the immediate benefits of the DDS architecture is that arbitrary waveform generation comes along for free. Instead of storing a sine table in the waveform RAM, a list of arbitrary values is saved. The phase accumulator is programmed to step through the stored values, one at a time, to play back the desired waveform through the output DAC.

The DDS's arbitrary waveform capability simplifies the task of generating the other 'standard' waveforms found in function generators. Ramp, sawtooth and even Gaussian white noise may be generated by changing the list of values in the waveform RAM.

The phase accumulator must be designed to support certain modes required for arbitrary waveforms. The rate at which RAM values are retrieved may be changed by simply using a different PIR value. However, variable record lengths, triggering functions, and wrap-around addressing are unique to arbitrary function generation.

As previously mentioned, a Bessel filter is required for arbitrary waveform generation. The Bessel filter will smooth the steps at the DAC output. With a -3 dB cutoff frequency, fc, of fs/4, the output will show a controlled risetime of 0.35/fc without overshoot.

Square Waves

Square waves are a special case for the DDS. One might think that a square wave could be generated by loading +1 and -1 into the waveform RAM. Indeed they can, but with the unfortunate restriction that the square-wave edges must be synchronous with the DDS sample clock. This restriction would greatly limit the resolution of available frequencies, especially at high frequencies.

A much better approach for generating squarewaves is to generate a clean sinewave, then discriminate the sine into a square wave. In this way, square waves will have the same frequency range and resolution as sine waves.

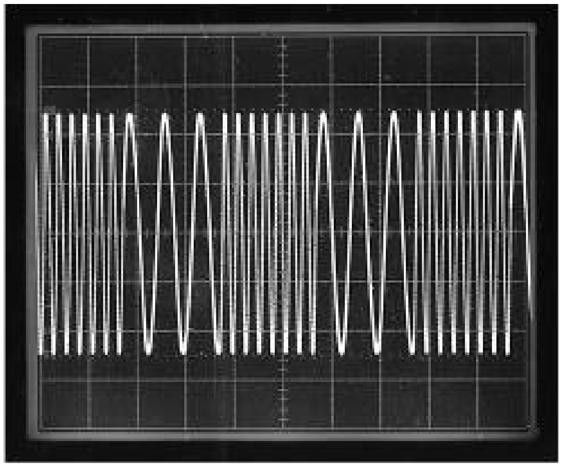


Figure 5. Frequency Shift Keying of Sinewave.

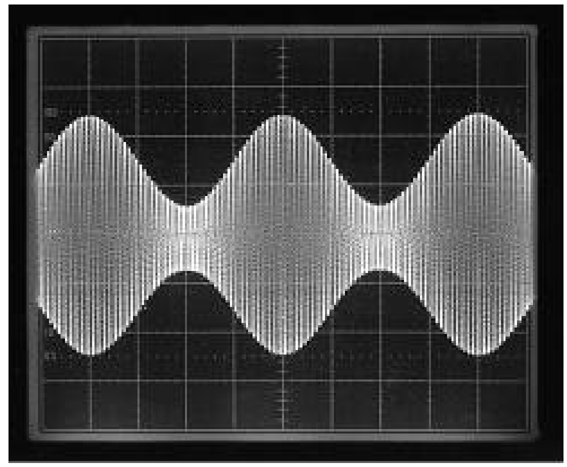


Figure 6. Amplitude modulation of Sine by Sine.

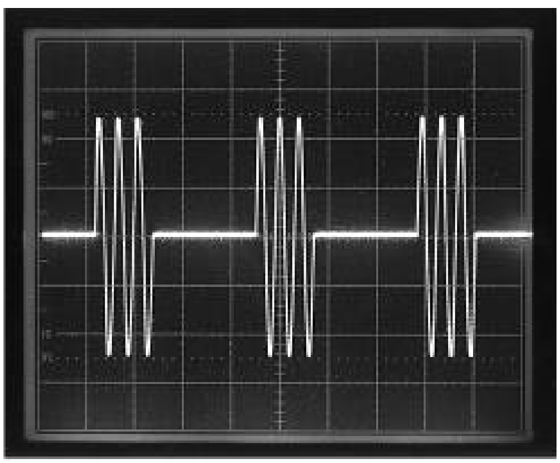


Figure 7. Three Cycle Burst of Sines

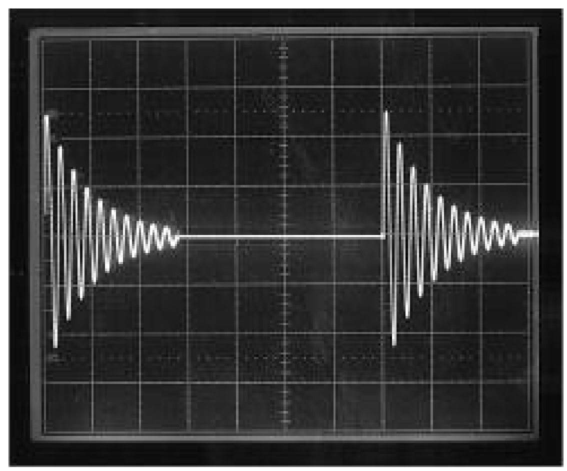


Figure 8. Arbitrary Waveforms.

Output Amplifiers

The output amplifier used in a DDS function generator must meet some stringent requirements. In order to preserve waveforms generated in the arbitrary mode, the amplifier must have a wide and flat passband, and exhibit a phase linear response well past the cutoff frequency of the Bessel filter.

The amplifier's bandwidth also determines the risetime of the squarewave output. Here again, a well behaved (phase linear) rolloff is required to prevent overshoot on the squarewave output.

Finally, the output amplifier must be able to drive 10 Vpp into a 50Ω load, meet distortion and settling specifications, and be protected against short circuits or connection to external power supplies. The output amplifier should exhibit a 50Ω output impedance regardless of output level setting.

To generate low signal levels, most function generators have output attenuators. The attenuators allow the output amplifier to work within a limited range of output levels, so that distortion and signal-to-noise ratios remain constant as the output levels are changed.

Floating Generator

Many applications require that the function generators be able to provide a signal to a load which is not ground referenced. Even if the load is nominally ground referenced, a floating generator output will provide a much cleaner signal because system ground loops are eliminated. It is important that the generator output shield is floating under all circumstances, even when the function generator is connected to a GPIB controller, or if an external frequency reference is connected to the instrument.

ASICs

DDS provides a new, clean, design approach for function generators. Much of the analog 'baggage' required for function generators is handled by digital logic circuits. Unfortunately, these logic circuits are big, complicated, and have to run fast. For example, a 15 MHz DDS requires a 48-bit adder operating at 40 MHz, with lots of glue logic. Fortunately, application specific integrated circuits (ASICs) provide a low cost solution to the problem.

A TTL prototype of the phase accumulator diagrammed in Figure 4 required about 150 ICs. The prototype was just able to work with a clock of 10 MHz. A CMOS gate array of the same design was fabricated in a 68-pin PLCC plastic package. The gate array operates at 40 MHz (worst case), uses about 1/4 watt of power, and has a recurring cost of about $10.

Conclusions

DDS based function generators are just beginning to appear in the market. These function generators offer substantial performance improvements, at reduced costs, over conventional analog function generators. As the cost of ASICs, RAMs and DACs decline, while their speed and resolution increase, expect to see DDS based function generators soon replace their analog counterparts.

Application Note #6

Signal Enhancement

Introduction

Many experimental techniques rely on the quantitative measurement of charged particles. Applications range from the very simple, such as the measurement of a dc anode current by a picoammeter, to the complex, such as the measurement of atomic state lifetimes using time-resolved particle counting.

Often, the signal of interest is obscured by noise. The noise may be fundamental to the process: discrete charges are governed by Poisson statistics which gives rise to shot noise. Or, the noise may be from more mundane sources, such as microphonics, thermal emf's, or inductive pick-up.

This article will describe methods for making useful measurements of weak signals, even in the presence of large interfering sources, emphasizing the electronic aspects of the measurement.

Figure 1 details the elements of a typical measurement situation. In this experiment the photo ionization crossection for a gas will be determined by passing a laser through the gas and by measuring the number of ions which are created.

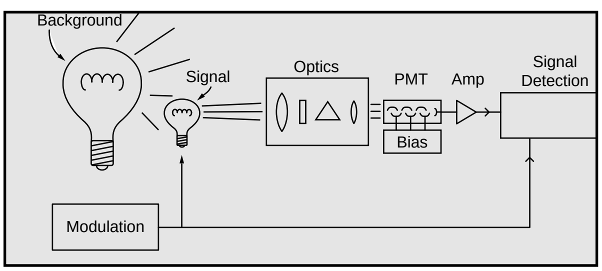


Figure 1: Prototype Experiment

In many applications, the output of the detector must be amplified or converted from a current to a voltage before the signal may be analyzed. Selection criteria for amplifiers include type (voltage or transconductance), gain, bandwidth, and noise. Often, the amplifier noise will be the limiting factor in determining the S/N ratio in a measurement, especially in situations where the charge detector has no gain.

Signal Analysis

There are two broad categories of signal analysis, depending on whether or not the source is modulated. Modulating the source allows the signal to be distinguished from the background. Often, source modulation is inherent to the measurement. For example, when a pulsed laser is used to induce a current, the signal of interest is present only after the laser fires. Other times, the modulation is "arranged", as when a CW source is chopped. Sometimes the source cannot be modulated, or the source is so dominant over the background as to make modulation unnecessary.

Noise Sources

An understanding of noise sources in a measurement is critical to achieving signal-to-noise performance near theoretical limits. The quality of a measurement may be substantially degraded by a trivial error. For example, a poor choice of termination resistance for an electron multiplier may increase current noise by several orders of magnitude.

Shot Noise

Light and electrical charge are quantized, and so the number of photons or electrons which pass a point during a period of time are subject to statistical fluctuations. If the signal mean is M photons, the standard deviation (noise) will be $\sqrt{M}$, hence the S/N $= M/\sqrt{M} = \sqrt{M}$. The mean, M, may be increased if the rate is higher or the integration time is longer. Short integration times or small signal levels will yield poor S/N values. Figure 2 shows the S/N which may be expected as a function of current level and integration time for a shot-noise limited signal.

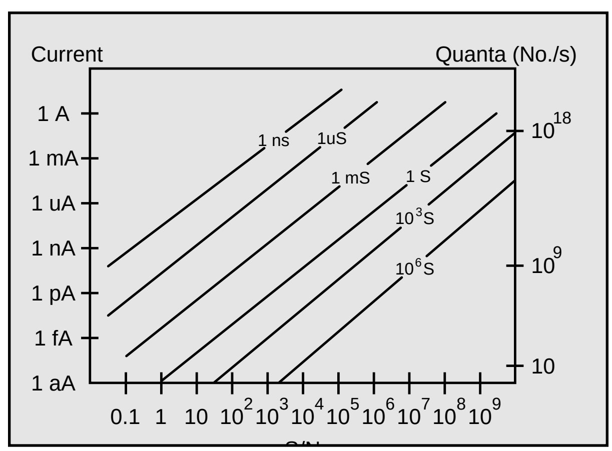


Figure 2:S/N vs. Flux and Time

"Integration Time" is a convenient parameter when using time domain signal recovery techniques. "Bandwidth" is a better choice when using frequency domain techniques. The rms noise current in the bandwidth Δf

Hz due to a "constant" current, I Amps, is given by:
$I_{shot\ noise} = \sqrt{(2 q I \Delta f)}$ where $q=1.6\times10^{-19}$ C

Johnson Noise.

The electrons which allow current conduction in a resistor are subject to random motion which increases with temperature. This fluctuation of electron density will generate a noise voltage at the terminals of the resistor. The rms value of this noise voltage for a resistor of R Ohms, at a temperature of T degrees Kelvin, in a bandwidth of Δf Hz is given by:

$$V_{johnson,rms} = \sqrt{(4kTR\Delta f)}$$

where k is Boltzman's constant. The noise voltage in a 1 Hz bandwidth is given by:

$$V_{johnson,rms} \text{ (per } \sqrt{Hz}) = 0.13nV \times \sqrt{(R(Ohms))}$$

Since the Johnson noise voltage increases with resistance, large value series resistors should be avoided in voltage amplifiers. For example, a 1 kΩ resistor has a Johnson voltage of about 4.1 nV/$\sqrt{}$Hz. If detected with a 100 MHz bandwidth, the resistor will show a noise of 41 μVrms, which has a peak-to-peak value of about 200 μV.

When a resistor is used to terminate a current source, or as a feedback element in a current-to-voltage converter, it will contribute a noise current equal to the Johnson noise voltage divided by the resistance. Here, the noise current in a 1 Hz bandwidth is given by:

$$I_{johnson\ rms} \text{ (per } \sqrt{Hz}) = 130 \text{ pA} / \sqrt{R} \text{ (Ohms)}$$

Since the Johnson noise current increases as R decreases, small value resistors should be avoided when terminating current sources. Unfortunately, small terminating resistors are required to maintain a wide frequency response. If a 1 kΩ resistor is used to terminate a current source, the resistor will contribute a noise current of about 4.1 pA/$\sqrt{}$Hz, which is about 1000x worse than the noise current of an ordinary FET input operational amplifier.

1/f Noise

The voltage across a resistor carrying a constant current will fluctuate because the resistance of the material used in the resistor varies. The magnitude of the resistance fluctuation depends on the material used: carbon composition resistors are the worst, metal film resistors are better, and wire wound resistors provide the lowest 1/f noise. The rms value of this noise source for a resistance of R Ohms, at a frequency of f Hz, in a bandwidth of Δf Hz is given by:

$$V_{1/f,rms} = IR \times \sqrt{(A\ \Delta f/f)}$$

where the dimensionless constant A has a value of about 10^{-11} for carbon. In a measurement in which the signal is the voltage across the resistor (IR), then the S/N = $3\times10^{5}\sqrt{(f/\Delta f)}$. Often, this noise source is a troublesome source of low frequency noise in voltage amplifiers.

Non-essential Noise Sources.

There are many discrete noise sources which must be avoided in order to make reliable low level light measurements. Figure 3 shows a simplified noise spectrum on log-log scales.
The key features in this noise spectra are frequencies

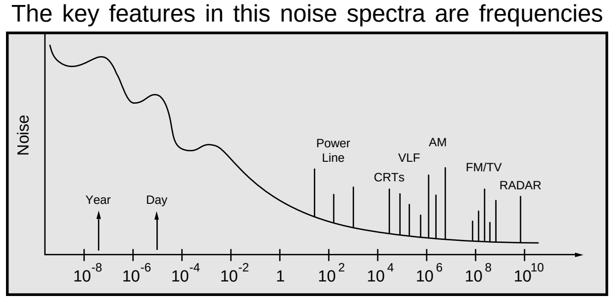


Figure 3: Simplified Noise Spectrum

worth avoiding: diurnal drifts (often seen via input offset drifts with temperature), low frequency (1/f) noise, power line frequencies and their harmonics, switching power supply and CRT display frequencies, commercial broadcast stations (AM, FM, VHF and UHF TV), special services (cellular telephones, pagers, etc.), microwave ovens and communications, to RADAR and beyond.

The best alternatives for avoiding these noise sources are:

1) Shield to reduce pick-up.
2) Use differential inputs to reject common mode noise.
3) Bandwidth limit the amplifier to match the expected signal bandwidth.
4) Choose a quiet frequency for signal modulation when using a frequency domain detection technique.
5) Trigger synchronously with interfering source when using a time domain detection technique.

Common ways for extraneous signals to interfere with a measurement are illustrated in Figure 4 (a-f).
Noise may be injected via a stray capacitance as in

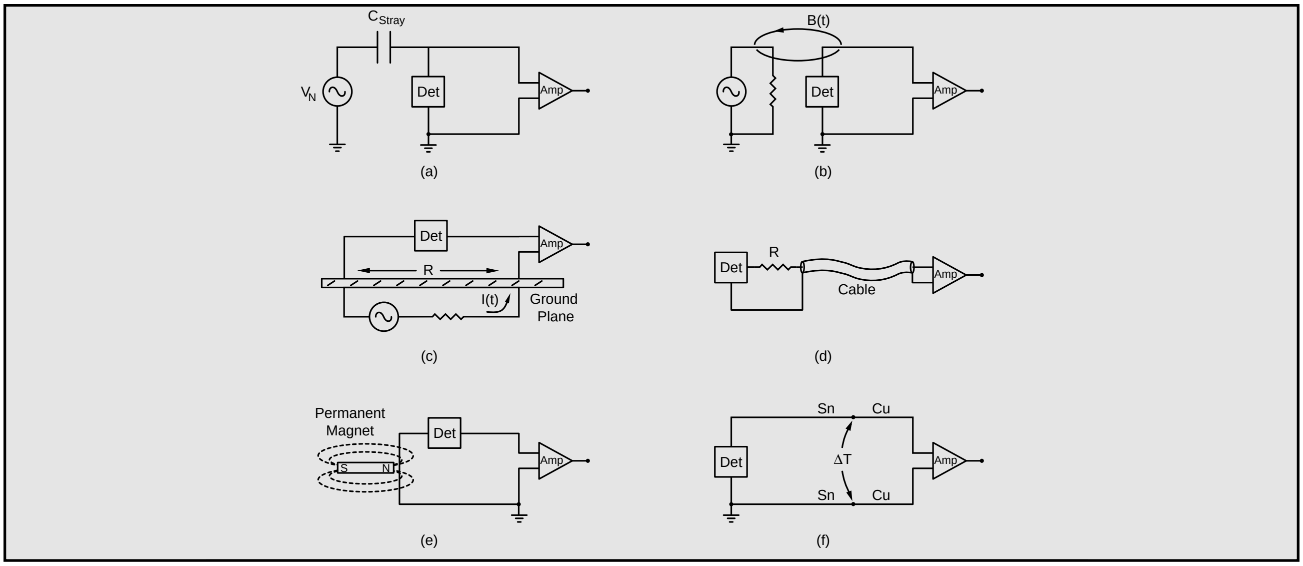


Figure 4: Coupling of Noise Sources

Figure 4a. The stray capacitance has an impedance of $1/j\omega C$. Substantial currents may be injected into low-impedance systems (such as transconductance inputs), or large voltages may appear at the input to high-impedance systems.

Inductive pick-up is illustrated in Figure 4b. The current circulating in the loop on the left will produce a magnetic field which in turn induces an emf in the loop on the right. Inductive noise pick-up may be reduced by reducing the areas of the two loops (by using twisted pairs, for example), by increasing the distance between the two loops, or by shielding. Small skin depths at high frequencies allow non-magnetic metals to be effective shields, however high-mu materials must be used to shield from low frequency magnetic fields.

Resistive coupling, or a "ground loop", is shown in Figure 4c. Here, the detector senses the output of the experiment plus the IR voltage drop from another circuit which passes current through the same ground plane. Cures for ground loop pickup include: grounding everything to the same point, using a heavier ground plane, providing separate ground return paths for large interfering currents, and using a differential connection between the signal source and amplifier.

Mechanical vibrations can create electrical signals (microphonics) as shown in Figure 4d. Here, a coaxial cable is charged by a battery through a large resistance. The voltage on the cable is V=Q/C. Any deformation of the cable will modulate the cable's capacitance. If the period of the vibration which causes the deformation is short compared to the RC time constant, then the stored charge on the cable, Q, will remain constant. In this case, a 1 ppm modulation of the cable capacitance will generate an ac signal with an amplitude of 1 ppm of the dc bias on the cable, which may be larger than the signal of interest.
The case of magnetic microphonics is illustrated in Figure 4e. Here, a dc magnetic field (the Earth's field or the field from a permanent magnet in a latching relay, for example) induces an emf in the signal path when the magnetic flux through the detection loop is modulated by mechanical motion.

Unwanted thermocouple junctions are an important source of offset and drift. As shown in Figure 4f, two thermocouple junctions are formed when a signal is connected to an amplifier. For typical interconnect materials (copper, tin) one sees about 10 µV/°C of offset. These extraneous junctions occur throughout instruments and systems: their impact may be eliminated by making ac measurements.

Amplifiers

Several considerations are involved in choosing the correct amplifier for an particular application. Often, these considerations are not independent, and compromises will be necessary. The best choice for an amplifier depends on the electrical characteristics of the detector, and on the desired gain, bandwidth, and noise performance of the system.

Charge counting and fast gated integration require amplifiers with wide bandwidth. A 350 MHz bandwidth is required to preserve a 1 ns rise time. The input impedance to these amplifiers is usually 50Ω in order to terminate coax cables into their characteristic impedance. When PMT's (which are current sources) are connected to these amplifiers, the 50Ω input impedance serves as the current to voltage converter for the PMT anode signal. Unfortunately, the small termination resistance and wide bandwidth yield lots of current noise.

It is important to choose an amplifier with a very high input impedance and low input bias current when amplifying a signal from a source with a large equivalent resistance. Commercial amplifiers designed for such applications typically have a 100 MΩ input impedance. This large input impedance will minimize attenuation of the input signal and reduce the Johnson noise current drawn through the source resistance, which can be an important noise source. Field Effect Transistors (FET's) are used in these amplifiers to reduce the input bias current to the amplifiers. Shot noise on the input bias current can be an important noise component, and temperature drift of the input bias current is a source of drift in dc measurements.

The bandwidth of a high input impedance amplifier is often determined by the RC time constant of the source, cable, and termination resistance. For example, a PMT with 1 meter of RG-58 coax (about 100 pF) terminated into a 1 MΩ resistor will have a bandwidth of about 1600 Hz. A smaller resistance would improve the bandwidth, but increase the Johnson noise current

Bipolar transistors offer an input noise voltage which may be several times smaller than the FET inputs of high input impedance amplifiers, as low as 1 nV/√Hz. Bipolar transistors have larger input bias currents, hence larger shot noise current, and so should be used only with low impedance (<1kΩ) sources.

When AC signals from very low source impedances are to be measured, transformer coupling offers very quiet inputs. The transformer is used to step-up the input voltage by its turns-ratio. The transformer's secondary is connected to the input of a bipolar transistor amplifier.

Conventional bipolar and FET input amplifiers exhibit input offset drifts on the order of 5 µV/˚C. In the case where the detector signal is a small dc voltage, such as from a bolometer, this offset drift may be the dominant noise source. A different amplifier configuration, chopper stabilized amplifiers, essentially measures the input offsets and subtracts the measured offset from the signal. A similar approach is used to "auto-zero" the offset on the input to sensitive voltmeters. Chopper stabilized amplifiers exhibit very low input offsets with virtually no input offset drift.

The use of "true-differential" or "instrumentation" amplifiers is advised to provide common mode rejection to interfering noise, or to overcome the difference in grounds between the voltage source and the amplifier. This amplifier configuration amplifies the difference between two inputs, unlike a single-ended amplifier, which amplifies the difference between the signal input and the amplifier ground. In high frequency applications, where good differential amplifiers are not available or are difficult to use, a balun or common mode choke may be used to isolate disparate grounds.

Transconductance Amplifiers.

When the detector is a current source (or has a large equivalent resistance) then a transconductance amplifier should be considered. Transconductance amplifiers (current-to-voltage converters) offer the potential of lower noise and wider bandwidth than a termination resistor and a voltage amplifier, however, some care is required in their application.

A typical transconductance amplifier configuration is shown in Figure 5. A FET input op amp would be used for its low input bias current. (Op amps with input bias currents as low as 50 fA are readily available.) The detector is a current source, Io. Assuming an ideal op amp, the transconductance gain is A = Vout/Iin = R_f, and, the input impedance of the circuit is Rin to the op amp's virtual null. (R_{in} allows negative feedback, which would have been phase shifted and attenuated by the source capacitance at high frequencies, to assure stability.) Commercial transconductance amplifiers use R's as large as 10 MΩ, with R_{in}'s which are typically R_f/1000. A low input impedance will insure that current from the source will not accumulate on the input capacitance.
This widely used configuration has several important

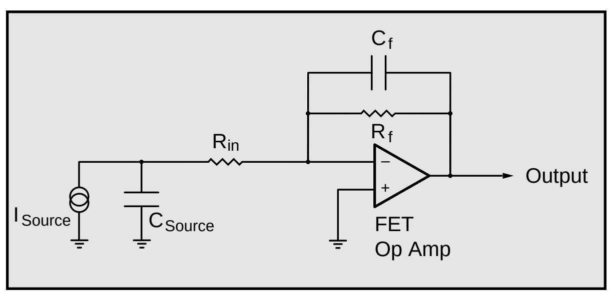


Fig. 5: Typical transconductance Amplifier

limitations which will degrade its gain, bandwidth and noise performance. The overall performance of the circuit depends critically on the source capacitance, including that of the cable connecting the source to the amplifier input. Limitations include:

1) The "virtual null" at the inverting input to the op amp is approximately R_f/Av where Av is the op amp's open loop gain at the frequency of interest. While op amps

have very high gain at frequencies below 10 Hz (typically a few million), these devices have gains of only a few hundred at 1 kHz. With an R_f of 1 GΩ, the virtual null has an impedance of 5 MΩ at 1 kHz, hardly a virtual null. If the impedance of the source capacitance is less than the input impedance, then most of the ac input current will go to charging this capacitance, thereby reducing the gain.

2) The configuration provides high gain for the voltage noise at the non-inverting input of the op-amp. At high frequencies, where the impedance of the source capacitance is small compared to R_{in}, the voltage gain for noise at the non-inverting input is R_f/R_{in}, typically about 1000. As FET input op amps with very low bias currents tend to have high input voltage noise, this term can dominate the noise performance of the design.

3) Large R_f's are desired to reduce the Johnson noise current, however large R_f's degrade the bandwidth. If low values of R_f are used, the Johnson noise current can dominate the noise performance of the design.

4) To maintain a flat frequency response, the size of the feedback capacitance must be adjusted to compensate for different source capacitances.

As many undesirable characteristics of the transconductance amplifier can be traced to the source capacitance, a system may benefit from integrating the amplifier into the detector, thereby eliminating interconnect capacitance. This approach is followed in many applications, from microphones to CCD imagers.

Signal Analysis

A variety of noise sources are avoided by AC measurement of the signal. When making DC measurements, the signal must compete with large low frequency noise sources. However, when the source is modulated, the signal may be measured at the modulation frequency, away from these large noise sources.

When the source is modulated, one may choose from gated integration, boxcar averaging, transient digitizers, lockin–amplifiers, spectrum analyzers, gated photon counters or multichannel scalers.

A measurement of the integral of a signal during a period of time can be made with a gated integrator. Commercial devices allow gates from about 100 ps to several milliseconds. A gated integrator is typically used in a pulsed laser measurement. The device can provide shot-by-shot data which is often recorded by a computer via an A/D converter. The gated integrator is recommended in situations where the signal has a very low duty cycle, low pulse repetition rate, and high instantaneous count rates.

The noise bandwidth of the gated integrator depends on the gate width: short gates will have wide bandwidths, and so will be noisy. This would suggest that longer gates would be preferred, however, the signal of interest may be very short lived, and using a gate which is much wider than the signal will not improve the S/N.

The gated integrator also behaves as a filter: the output of the gated integrator is proportional to the average of the input signal during the gate, so frequency components of the input signal which have an integral number of cycles during the gate will average to zero. This characteristic may be used to 'notch out' specific interfering signals.

It is often desirable to make gated integration measurements synchronously with an interfering source. (This is the case with time-domain signal detection techniques, and not the case with frequency domain techniques such as lockin detection.) For example, by locking the pulse repetition rate to the power-line frequency (or to any sub-multiple of this frequency) the integral of the line interference during the short gate will be the same from shot-to-shot, which will appear as a fixed offset at the output of the gated integrator.

Shot-by-shot data from a gated integrator may be averaged to improve the S/N. Commercial boxcar averagers provide linear or exponential averaging. The averaged output from the boxcar may be recorded by a computer or used to drive a strip chart recorder. Figure 6 shows a gated integrator with an exponential averaging circuit.

Lock-in Amplifiers.

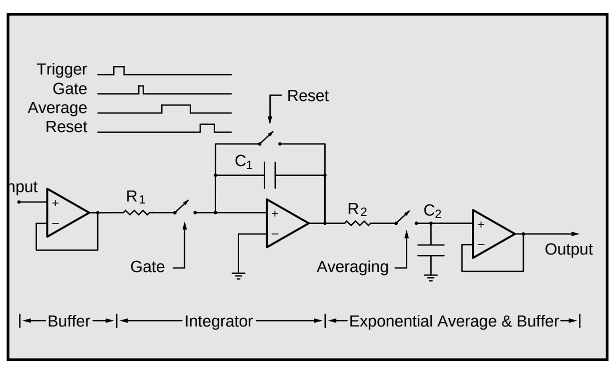


Fig. 6: Gated Integrator and Boxcar Averager

Phase sensitive synchronous detection is a powerful technique for the recovery of small signals which may be obscured by interference which is much larger than the signal of interest. In a typical application, a cw laser which induces the signal of interest will be modulated by an optical chopper. The lock-in amplifier is used to measure the amplitude and phase of the signal of interest relative to a reference output from the chopper.

Figure 8 shows a simplified block diagram for a lockin amplifier. The input signal is ac coupled to an amplifier whose output is mixed (multiplied by) the output of a phase-locked loop which is locked to the reference input. The operation of the mixer may be understood through the trigonometric identity:

$A\cos(\omega_1 t+\phi) \times B\cos(\omega_2 t) = 1/2AB[\cos((\omega_1+\omega_2)t+\phi) + \cos((\omega_1-\omega_2)t+\phi)]$

When $\omega 1=\omega 2$ there is a DC component of the mixer output, $\cos(\phi)$. The output of the mixer is passed through a lowpass filter to remove the sum frequency component. The time constant of the filter is selected to reduce the equivalent noise bandwidth: selecting longer time constants will improve the S/N at the expense of longer response times.

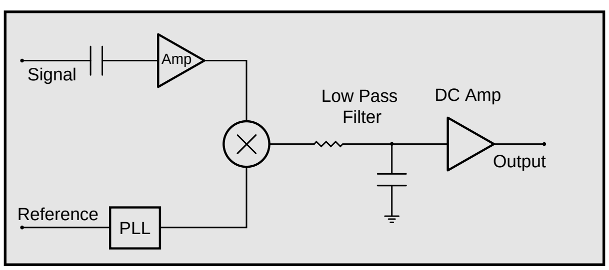


Figure 8: Lock-in Amplifier Block Diagram

The simplified block diagram shown in Figure 8 is for a "single phase" lock-in amplifier, which measures the component of the signal at a single phase with respect to the reference. A dual phase lockin has another channel which measures the component of the signal at 90 degrees relative to the first channel, which allows simultaneous measurement of the amplitude and phase of the signal.

Digital Signal Processing Techniques

Digital signal processing (DSP) techniques are rapidly replacing the older analog techniques for the synchronous detection of the signal. In these instruments, the input signal is digitized by a fast, high resolution A/D converter, and the signal's amplitude and phase are determined by high speed computations in a digital signal processor. To maintain the 100 kHz bandwidth of the analog designs, the DSP designs must complete a quarter million 16 bit A/D conversions and 20 million multiply-and-accumulate operations each second. Many of artifacts of the analog designs are eliminated by the DSP approach; for example, the output drift and dynamic range of the instruments are dramatically improved.

Photon Counting.

Photon counting techniques (which are readily applicable to particle counting measurements) offer several advantages in the measurement of light: very high sensitivity (count rates as low as 1 per minute can be a usable signal level), large dynamic range (signal levels as high as 100 MHz can be counted, allowing a 195 dB dynamic range), discrimination against low level noise (analog noise below the discriminator thresholds will not be counted), and ability to operate over widely varying duty cycles.

Key elements of a photon counting system include: a high gain PMT (or charge multiplier) operated with sufficient high voltage so that a single photoelectron (or charged particle) will generate an anode pulse of several millivolts into a 50Ω load, a fast discriminator to generate logic pulses from anode signals which exceed a set threshold, and fast gated counters to integrate the counts.

In situations where the time evolution of a light signal must be measured (LIDAR, lifetime measurements, chemical kinetics, etc.) transient photon counters allow the entire signal to be recorded for each event. In these instruments, the discriminated photon pulses are summed into different bins depending on their timing with respect to a trigger pulse. Commercial instruments offer 5 ns resolution with zero dead-time between bins. The time records from many events may be summed together in order to improve the S/N.

Which instrument is best suited for detecting signals from a photomultiplier tube? The answer is based on many factors, including the signal intensity, the signal's time and frequency distribution, the various noise sources and their time dependence and frequency distribution. In general, the choice between boxcar averaging (gated integration) and lock-in detection (phase sensitive detection) is based on the time behavior of the signal. If the signal is fixed in frequency and has a 50% duty cycle, lock-in detection is best suited. This type of experiment commonly uses an optical chopper to modulate the signal at some low frequency. Signal photons occur at random times during the 'open' phase of the chopper. The lock-in detects the average difference between the signal during the 'open' phase and the background during the 'closed' phase.

To use a boxcar averager in the same experiment would require the use of very long, 50% duty cycle gates since the photons can arrive anywhere during the 'open' phase. Since the gated integrator is collecting noise during this entire gate, the signal is easily swamped by the noise. To correct for this, baseline subtraction can be used where an equal gate is used to measure the background during the 'closed' phase of the chopper and subtracted from the 'open' signal. This is then identical to lock-in detection. However, lock-in amplifiers are much better suited to this, especially at low frequencies (long gates) and low signal intensities.

If the signal is confined to a very short amount of time, then gated integration is usually the best choice for signal recovery. A typical experiment might be a pulsed laser excitation where the signal lasts for only a short time (100 ps to 1 µs) at a repetition rate up to 10 kHz. The duty cycle of the signal is much less than 50%. By using a narrow gate to detect signal only when it is present, noise which occurs at all other times is rejected. If a longer gate is used, no more signal is measured but the detected noise will increase. Thus, a 50% duty cycle gate would not recover the signal well and lock-in detection is not suitable.

Photon counting can be used in either the lock-in or the gated mode. Using a photon counter is usually required at very low signal intensities or when the use of a pulse height discriminator to reject noise results in an improved S/N. If the evolution of a weak light signal is to be measured, a transient photon counter or multichannel scaler can greatly reduce the time required to make a measurement.

Application Note #7

Vacuum Diagnosis With a Residual Gas Analyzer

Introduction

Residual Gas Analyzer (RGA) is the term for a class of quadrupole mass spectrometers that typically cover mass ranges from 1 to 100 or 200 amu (atomic mass units) and are intended to be used for the analysis of the gasses present in high and ultra high vacuum. The RGA's resolution is sufficient to clearly distinguish peaks that are 1 amu apart. These specifications are a perfect match for the requirements of vacuum diagnosis. Not many materials with a mass greater than 200 amu will be volatile and the high resolution of research grade mass spectrometers is not necessary for the analysis of low molecular weight species. Overall, RGA's are affordable instruments that can be permanently attached to a vacuum system.

The purpose of vacuum is to remove molecules that would interfere with a process or experiment. Although the reduction of total pressure is always a concern in vacuum systems, near operating pressure the real concern becomes the presence of certain species, e.g. oxygen, water, and hydrocarbons. When operating a vacuum system with only a total pressure vacuum gauge, one must rely on the assumption that the total pressure is direct indicator of the partial pressure of these crucial impurities. This assumption presents two problems. First, a total pressure measurement cannot tell the user whether the vacuum system is filled with water, nitrogen, carbon dioxide or hydrogen. Second, total pressure measurement is not very precise— measurements better than 10% are difficult and expensive. If a system is operating at a standard pressure of 1.0×10^{-7} and the pressure rises to 1.1×10^{-7} is there a cause for concern? The additional 10% could be harmlessly inert or it could be oil vapor. The RGA is designed to address exactly this question. It immediately tells the user **what** is in their vacuum system.

The primary application of the RGA is to analyze the composition of a vacuum system. The composition can be used to detect impurities, monitor gas fills, or analyze chemistry that is occurring. The second application of the RGA is as an intrinsic leak detector. It serves this purpose very well and is in many ways superior to a portable helium leak tester. In the remainder of this application note we will illustrate the usefulness of an RGA. The data shown are real results from experiments designed to illustrate certain points.

Composition Analysis

Stanford Research Systems' RGA software allows the composition of a vacuum system to be analyzed by two methods. The most common is to measure the mass spectrum of the vacuum. This provides a "fingerprint" of the vacuum composition. A second method is to track specific species or peaks of the mass spectrum. The first method, analog scan mode, is most useful when the user does not know what is present in the chamber. Once the identities of the species have been determined, individual peaks can be tracked using either pressure vs. time, table or annunciator mode.

The Mass Spectrum

The fundamental operation of the RGA is as a mass spectrometer. Figure 1 shows a graph of partial pressure versus mass, which was measured with an RGA with an electron multiplier detector. The scan was taken of a vacuum system near its ultimate vacuum. The pressure axis is plotted on a logarithmic scale so that a

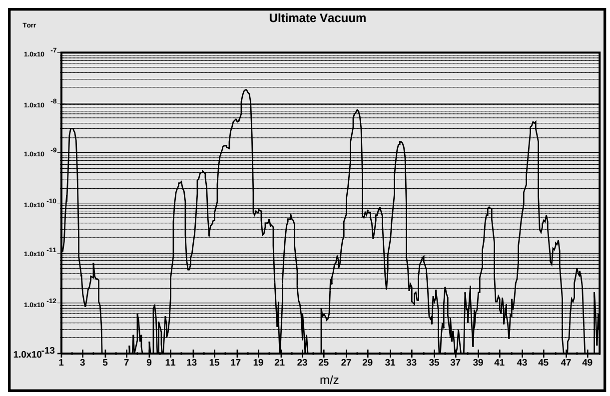


Fig.1:Partial Pressure vs. Mass

large range can be seen. The log scale makes the peaks appear wider than when plotted on a linear scale. This scan from 1 to 50 amu shows some gasses commonly present in vacuum chambers. There are many peaks, but they are caused mainly by 7 species. Hydrogen is at 2 and helium at 4. Water gives primary peaks at 16, 17, and 18 due to the species O^+, HO^+, and H_2O^+ . The smaller peaks at 19 and 20 are due to ^{18}O which is naturally present at 0.2%. Nitrogen is at 28 and also causes the peaks at 14 by atomic N^+ and the doubly ionized N_2^+. Molecular oxygen shows a peak at 32 and an isotope peak at 34. Argon shows a peak at 40. Carbon dioxide shows a peak at 44 and a peaks for CO_2^{++} and C^+ at 22 and 12. The other peaks are caused by fragments of these species and contaminants.

The presence of air components in the spectra might lead us to believe that the system is leaking, but this is untrue. The hybrid turbomolecular pump has simply reached its compression limit. The foreline of the pump was operating at a total pressure of 0.5 Torr; therefore the compression ratio is about 108 (as the pump specifications indicate). Nitrogen, oxygen and argon are all present in the same ratios as standard atmosphere. The presence of helium is interesting, because it is present in the atmosphere at about 7 ppm. Its peak might be expected 6 decades smaller than the nitrogen peak. The low compression ratio of the turbo pump for helium (10^5) explains why the peak is only three decades smaller.

The ability to detect these common species and many others is the essence of the RGA. The fragmentation of molecules in the ionizer of the RGA gives each molecule a distinct fingerprint. The fragmentation patterns for many molecules is available from the library of the SRS RGA software. Keeping a historical record of the typical spectrum of a vacuum system allows the appearance of new peaks to be instantly detected. For instance, the peak at 48 in Figure 1 is SO from SO_2. On occasion this peak is seen in our chambers. In addition to simple gasses such as SO_2, we are also interested in molecules with higher weights

Oil Contamination

Figure 2 contains a mass spectrum of a common contaminant of vacuum systems: oil. Figure 2a shows measured data while figure 2b shows a spectrum of pump oil from the library. The presence of mechanical pump oil is immediately obvious. The peaks at masses 39, 41, 43, 55, and 57 are caused by mechanical pump oil backstreaming into the vacuum chamber during a load lock sequence. The total pressure in the chamber was dominated by water and was less than $2x10^{-8}$. In this case, the total pressure might satisfy operating conditions but the spectra reveals that the system is heavily contaminated with oil. This could have been caused by improper valve sequencing or a saturated oil trap. Without an RGA, only operating procedures for valves and a maintenance schedule for traps can ensure that the cleanliness of a vacuum system is maintained. With the RGA, cleanliness can be proven **before** a process or experiment begins.

Solvent Contamination

Oil contamination is common in vacuum systems. Cleaning parts with solvents is a common approach to removing this contamination. We have observed that organic solvents such as acetone and TCE are more tenacious contaminants than the oil they are designed to remove. Figure 3 shows a spectrum indicating contamination with 1,1,1-trichloroethane as shown by the major peaks at 97 and 99, and the minor peaks at 61, 63, 117 and 119. The paired peaks are caused by the natural isotopic occurrence of ^{35}Cl and ^{37}Cl (75% and 25%). This spectrum was measured one week after the initial contact with the solvent. The TCE permeated into the o-rings in the system during a cleaning step. The TCE continued to outgas from the O-rings for two weeks and showed no signs of stopping. At that time they were removed and baked in an oven, which eventually removed the TCE.

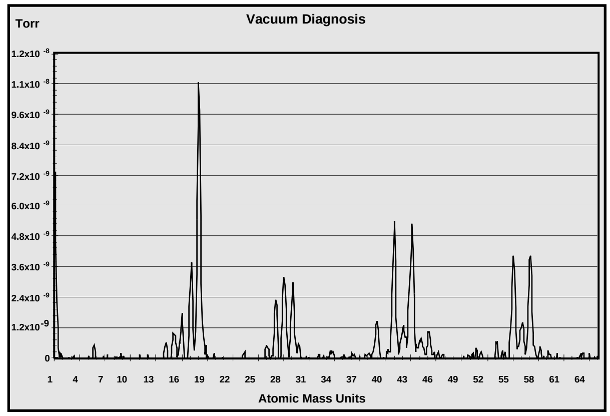


Figure 2a: Pump Oil Contamination

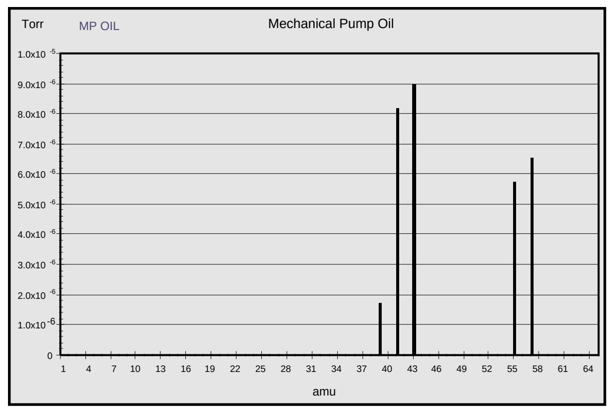


Figure 2b: Library Pump Oil Data

Data like this is invaluable to the development of cleaning procedures. While the TCE successfully removed the oil, the vacuum chamber was left more contaminated than it would have been without the cleaning. The mass spectrum provides a more accurate evaluation of cleaning procedures than pump down time and base pressure. Just because a system pumps down quickly does not guarantee that an undesirable contaminant is not present. The large dynamic range also allows evaluations to be made more quickly. The user does not have to wait several hours for the water to pump away to determine if a system is contaminated. The RGA can make measurements in the presence of a high water partial pressure.

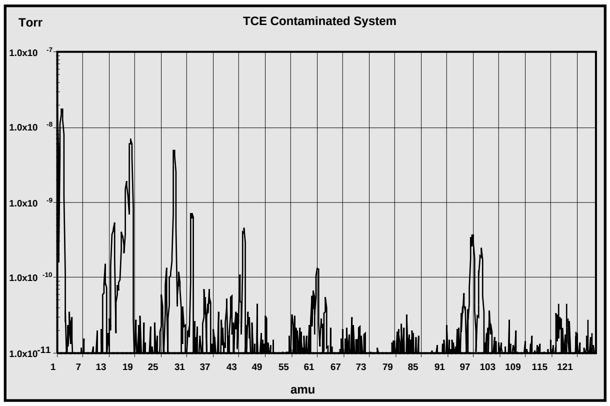


Figure 3: TCE Contaminated System

Peak Measurement

For vacuum systems that only need to be clean, the mass spectrum is the most useful measurement. During experiments and processes the partial pressure of certain species is of more interest. The RGA software provides three modes that are used to measure selected peaks. The selection of which mass is associated with which species is usually straightforward, i.e. the mass of the molecule is chosen. When two species have overlapping peaks patterns, the user chooses the strongest peak that does not interfere. For example N_2 and CO both have a mass of 28. In a system with a large CO interference, N_2 could be measured at mass 14. Because the peak at 14 is smaller than the major peak, a scaling factor is required. The set of peaks of interest are entered into the RGA software by the user. During measurements, the RGA measures only each peak. Because the whole spectrum is not recorded, data is acquired much faster. The two examples that follow show an interesting example of pressure vs. time measurements and a method of increasing the dynamic range of measurements.

Pressure vs. Time

Figure 4 shows the use of the P vs. T Mode to monitor an airlock sequence. The process opens an air lock, places the sample in it, and moves the sample into the main vacuum chamber. The air lock is pumped from atmospheric to rough vacuum using the same mechanical pump as the main vacuum chamber. This requires isolating the foreline of the turbo pump during the time the sample is in the airlock. To keep oxygen out of the main vacuum chamber, the airlock is flushed with dry nitrogen then pumped to rough vacuum. The flush is repeated two additional times before the sample is finally transferred. If this procedure is successful, the main vacuum chamber should not be disturbed. The text which follows gives a detail of the events.

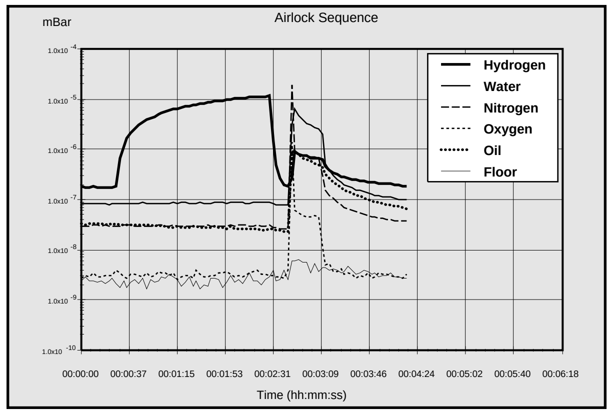


Figure 4: Airlock Sequence

To make these measurements, the electron multiplier detector has been used with a gain of 100, which allows all six channels to be recorded every three seconds. The standard Faraday cup detector is able to detect these partial pressures, but not at this rate. The “floor” channel is set to mass 21. There is rarely anything present at this mass, which allows it to be used as an indicator of the noise floor.

The sequence starts with the main vacuum chamber at its base pressure of $2x10^{-8}$ The chamber is pumped by a turbomolecular pump that is backed by a rotary vane pump. A small load lock is attached to the chamber, which can be rough pumped by the same mechanical pump and purged with nitrogen.

At 0:30, the isolation valve between the turbo pump exit and mechanical pump is shut so that the mechanical pump can be used to rough pump the load lock. During this time, the load lock is repeatedly filled with nitrogen and pumped down. Of interest in the data, is the rise in the hydrogen partial pressure during this step of the sequence. The partial pressure of hydrogen increases by a factor of 100 while the partial pressure of the other gasses barely increases. This difference is caused by the low compression ratio that turbo pumps have for light gasses. The heavier gasses are being compressed into the dead volume between the turbo pump exit and isolation valve. But, the turbo pump has insufficient compression ratio to store hydrogen in this manner, causing the partial pressure of hydrogen to rise.

At 2:30, the load lock has been roughed and the turbo pump foreline isolation valve is opened. The pressure of H_2 immediately drops back to the base pressure value. At 2:40, the load lock is opened to the main chamber causing a jump in pressure. The rise in oxygen and oil pressure indicates that the procedure is operating poorly. Even though the load lock was purged three times with 99.999% nitrogen, oxygen was still introduced into the chamber. This was either caused by a small air leak into the load lock, or permeation of oxygen out of the elastomer seals on the load lock. The rise in oil partial pressure indicates that the trap on the mechanical pump is exhausted and allowed oil to backstream into the load lock.

At 3:09, the valve between the load lock and main chamber is closed, and the pressures begin to return towards their base values. Oxygen is pumped out of the chamber. Hydrogen, water, and nitrogen recover their original values, but slowly. The oil is alarming because it persists at a higher concentration. If this sequence occurred several more time, the oil would continue to step up. The RGA allows it to be detected before reaching undesirable values.

Table Mode

The noise floor of the Faraday cup detector is about 10^{-10} mbar. Since the maximum operating pressure is 10^{-4} mbar, the dynamic range of the RGA is 6 decades or 1 ppm. The noise floor of the channel electron multiplier (CEM) is lower, but its maximum operating pressure also decreases with the noise floor. The two pressure limits change such that the dynamic range of the CEM is still 6 decades. By switching between the two detectors, measurements covering more than 6 decades can be made. The table mode of the SRS RGA software allows such a measurement to be made. The CEM status can be set independently for each mass being monitored. Figure 5 shows a configuration where the prevalent gasses are detected with the FC and the low pressure gasses are detected with the CEM. A comparison of the value for nitrogen and “floor” show that the apparent dynamic range is 8 decades or 10 ppb. The program will automatically sort the channels so that all the measurements requiring the CEM are made as a group, which minimizes the switching on and off of the detector. Without the ability to choose between FC and CEM detector for each channel, the CEM would have to be used for all channels so that the low pressure gasses could be detected. Operation like this would cause the CEM detector to saturate at the high pressure peaks. Saturation of the detector makes the value useless, and also increases the physical wear rate of the CEM.

RGA Table Scan

Ch#	Name	Mass	Value	Alarm	Speed	Cal
1	Hydrogen	2	3.8E-07	NORMAL	1	1.00
2	Water	18	7.1E-08	HIGH	1	1.00
3	Nitrogen	28	1.4E-05	HIGH	1	1.00
4	Oxygen	32	4.6E-10	NORMAL	3	1.00
5	CO2	44	3.4E-11	NORMAL	3	1.00
6	Oil	55	1.6E-12	NORMAL	3	1.00
10	Floor	21	1.5E-13	NORMAL	1	1.00

Figure 5: SRS RGA Table Display

Leak Testing

In addition to the diagnosis of vacuum systems, the RGA is invaluable as an intrinsic leak detector. It is always available and does not require perturbing the system. The user does not have roll up a large leak detector and attach it to the vacuum system. The system does not have to be brought up to atmospheric pressure. The RGA can operate in leak detection mode using any gas, so it does not require helium. For moderate leaks, argon or tetrafluoroethane (a typical gas in cans of aerosol dust remover) can be used. Only for the smallest leaks is helium necessary. Having a built - in leak detector makes working with vacuum systems much easier and faster, and the SRS RGA is far less expensive than a leak detector.

The process of leak detection with an RGA is the same as with a traditional helium leak detector. Place the software in leak detection mode, indicate the mass of the test gas and watch the partial pressure as various joints in the vacuum system are sprayed with the test gas. When the leak is sprayed with the test gas, the partial pressure will rise. The response is immediate if the leak is a direct path from the outside to the inside of the system. Figure 6 contains the result of a leak test with helium on a vacuum chamber. The tester moves the helium probe towards and then past the leak, causing the first peak. Once the location of the leak is bounded, the tester goes back to exactly locate the leak. For most situations leak testing is straightforward and no different than traditional methods. In the following sections, we discuss a few situations where traditional methods fail and how the RGA provides a better method.

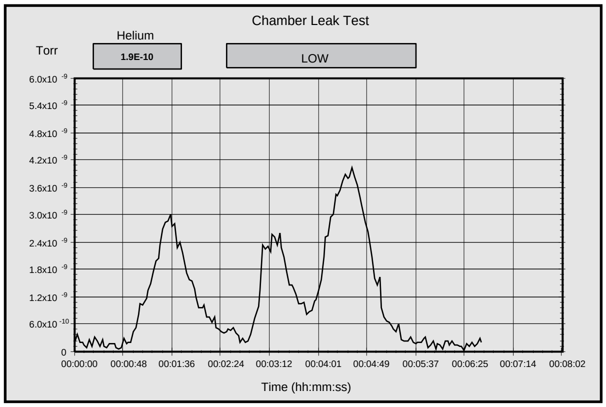


Figure 6: Helium Leak Test Data

The partial pressure of the test gas is directly related to the leak rate into the chamber. Assuming that the vacuum pump is not operating near its compression limit, the throughput of the test gas is equal to the product of the partial pressure and the effective speed of the pump at the RGA ionizer ($Q = S{\cdot}P$) In Figure 5, the partial pressure was measured in Torr. The effective speed of the turbo pump for helium was approximately 50 liter s^{-1} and the largest peak of $4x10^{-9}$ represents a leak rate of $1.5x10^{-7}$ scc/sec. From this figure we can estimate a minimum detectable leak of $1x10^{-8}$ scc/sec which is measured with the Faraday Cup detector To measure smaller leaks, the electron multiplier detector can be used, or the turbo pump can be throttled to decrease the pumping speed.

Supply Gas Valve Seats

Leaks across valves that supply gasses to a vacuum system cannot be detected with conventional helium leak testers (unless the valve supplies helium). To test a suspect valve would require removing it and attaching it to the leak tester. Because the RGA can monitor any gas, this is unnecessary; the valves on gas supply lines can be tested in situ. The procedure is simple: monitor the composition of the vacuum system with a high pressure and low pressure behind the valve seat in question. If the partial pressure of the gas in question changes, the valve seat is leaking.

Supply Gas Manifolds

Leak testing supply gas lines can be a very trying experience. The difficulty is largely because compression type fittings do not have a leak test port. Another difficulty is that supply manifolds commonly have a large number of connections in close proximity. Because the leak in a compression fitting is inside the fitting, transporting the test gas to the leak requires a large flowrate and waiting for an extended time for the gas to diffuse into the fitting. Because of the flowrate and time, it is possible that the test gas can travel to adjacent tube fittings and cause a misleading indication of a leak. On many occasions we have observed “fugitive” leaks that appear and disappear at a specific fitting. What happens is that the test gas inadvertently flows to another fitting of the manifold. Confining the test gas to the fitting under question can help, but the RGA provides a easier solution: use a gas other than helium. Helium will spread in air quickly and diffuse into many fittings. A heavy gas like argon or tetraflouroethane is far easier to confine to a specific fitting. Once the moderate leaks have been located and eliminated, a follow up with helium to check for tiny leaks is warranted.

Bellow Valves

Bellow valves can be difficult to leak test due to the large volume of gas contained between the bellow and the valve body. To perform a quick leak test it is required to change the composition of the gas in this trapped volume quickly. Unfortunately, for some valves, this volume of gas is not highly accessible. This greatly reduces the response time of a leak test. For a leak causing a base pressure of 10^{-7} mbar in a vacuum system with a 70 l/s pump, the volumetric flowrate of gas entering from the atmospheric side of the leak is 7 nl/s. For a bellow with a trapped volume of 1 ml, the response time constant would be over 40 hours. This emphasizes the importance of leak test ports on vacuum hardware.

In such situations it is common to place a bag over the body of the valve and fill the bag with helium. The RGA allows us to consider using gasses other than helium. The permeability of helium through elastomeric seals can give a false leak reading. And as in manifolds, unless the helium can be strictly confined to the valve body, it may spread to adjacent connections. Given the amount of work and lost time required to remove and repair large valves, false leak readings are expensive. A second test with another gas, such as argon, can confirm that a suspect valve is leaking before starting out on the repair.

Conclusion

An RGA is a real eye opener for users of vacuum systems. With an RGA the process of working with vacuum systems is elevated from an empirical trial and error approach to a systematic approach. It is hard to imagine anybody who has worked on a vacuum system with an RGA ever wanting to “go back” to a system without one.

Application Note #8

Using RGAs to Sample High Pressure Gasses

Introduction

The types of analysis performed by an RGA are useful in many applications other than vacuum systems. But, the RGA is intrinsically a vacuum instrument that operates best near 10^{-6} mbar. Above 10^{-5} the response becomes non-linear and above 10^{-4} the filament will be shut off by the control electronics. To sample gasses at higher pressures, a pressure reduction system is needed. These systems are basically a restriction and a vacuum pump package. Common restrictions are pinholes and capillaries, which can provide pressure reductions of more than 6 decades. The vacuum pump package consists of a turbomolecular pump and a backing pump. In addition to achieving the desired pressure reduction, the design of a system should provide for a fast response and high signal to background ratio.

At pressures common to vacuum processes, a simple aperture-based pressure reduction system is suitable. At atmospheric and higher pressures, a two stage reduction based on a capillary and aperture is used. These two systems will be used to illustrate the design of pressure reduction systems for RGA's.

Vacuum Process Sampling (10 to 10^{-5} mbar)

Figure 1 shows a schematic of a basic pressure reduction system. The system has two paths to the RGA: a high conductance path and an aperture path. The high conductance path (through Valve Hi-C) is provided so that the RGA can monitor the ultimate vacuum of systems before a process begins. The Hi-C path is also used when leak testing the vacuum system with the RGA Software's leak test mode. The aperture path provides the pressure reduction for when the vacuum process is operating at pressures up to 10 mbar.

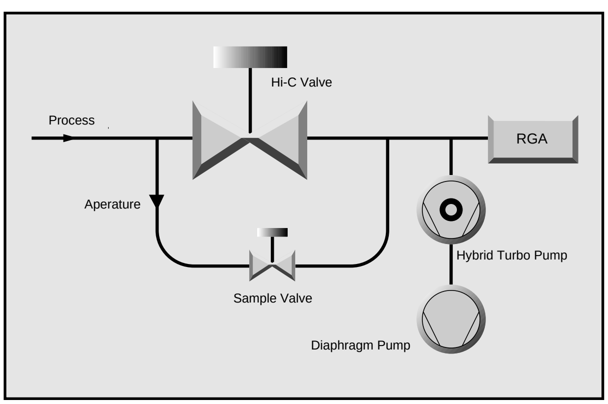


Figure 1: Schematic of a mid-vacuum pressure reduction system

Apertures can be readily designed for process pressures in the range from 10^{-3} mbar to 10 mbar. If the process always operates within a small range, the aperture can be optimized to deliver gas to the RGA at about 10^{-6} to 10^{-5} mbar. By operating the RGA at its optimum pressure the data acquisition time is kept to a minimum and the full dynamic range in partial pressure is available. For many applications, the process is operated at one pressure and the aperture can be optimized. If the process pressure varies over a range of 2 decades or more, the aperture size must be compromised to tolerate the pressure range. For example, consider a process pressure that varies from 10^{-1} to 10 mbar. The aperture would be designed to drop the pressure from 10 mbar to 10^{-5} mbar When the process pressure was at 10^{-1} mbar, the pressure at the RGA would be 10^{-7} mbar. The noise floor of the RGA does not depend on the process pressure; for a Faraday cup detector it is about 10^{-10} mbar. Therefore the dynamic range of the measurement varies from 5 decades at high process pressure to only 3 decades at the low pressures. For applications where the full dynamic range is not needed, operating the RGA at low pressure may be acceptable. If the full dynamic range is required over a variety of process pressures, a variable reduction is required. Suitable variable leak valves are available, but are significantly more expensive than a fixed aperture.

Another method of increasing the dynamic range and data acquisition rate is to use an RGA with an electron multiplier. The electron multiplier provides gains from 10^2 to 10^6 and lowers the noise floor to as low as 10^{-14} mbar. This lower noise floor allows the RGA to provide large dynamic range even at low operating pressures.

A high operating pressure (or throughput of the aperture) at the RGA also improves the signal to background ratio. In this context, signal is the gas that is drawn through the aperture and background is outgassing from the system plus backstreaming through the turbo pump. The ultimate vacuum of many turbo pump packages is about 10^{-9} mbar. The outgassing background will be mostly hydrogen, water, and nitrogen. The backstreaming background will be air. If measurements are being made near these background peaks, the operating pressure should be kept as high as possible. The background can be minimized by designing the tubing such that the effective pumping speed at the RGA ionizer is as high as possible. Figure 2 shows two layouts that both have the same "signal" level. The layout with the RGA at the end of a small tube has a small effective pumping speed and will show a larger background level.

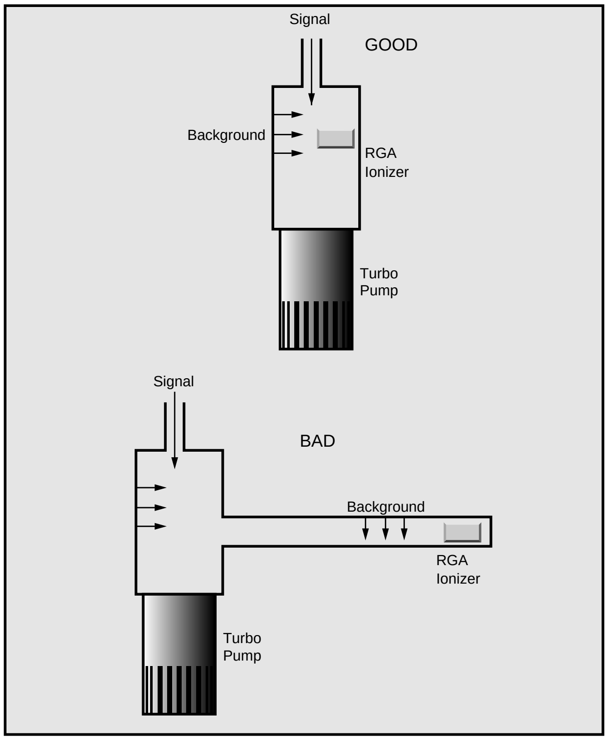


Figure 2: Two Layouts of Post-Aperature Vacuum System

The system shown in Figure 1 can be assembled as a simple package. Choosing a small (70 liter/s or less) hybrid turbo pump and a diaphragm backing pump will eliminate any concern of oil. The use of this pump pair also eliminates foreline traps and isolation valves. The operation of the system should be simple: open the Hi-C valve at low pressures, or open the sample valve at high pressures.

High Pressure Sampling (>100 mbar)

At high pressure the aperture assembly is insufficient to reduce the pressure, while maintaining response time. Consider an aperture that reduces the pressure from 10 mbar to 10^{-6} mbar when used with a 70 liter/s turbo pump. The volumetric flowrate on the high pressure side of the aperture would be 7 microliter/s. Any dead volume on the high pressure side of the aperture (Figure 3) would cause a large response time constant ($t_c = volume/flowrate$). If the aperture had a small dead volume of 1/2 inch of 0.250 OD tube (0.028 wall), the time constant would be 35 seconds. This is not an acceptable response time.

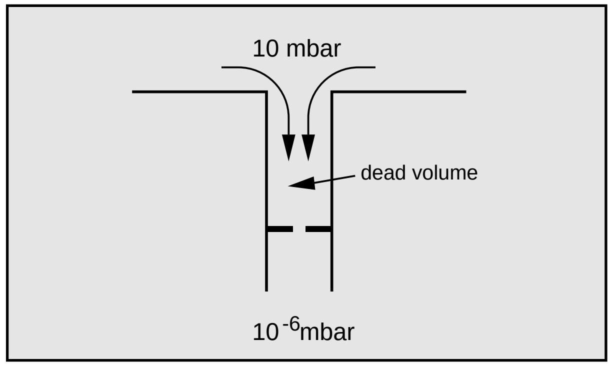


Figure 3: Small Dead Volume Slows Process Response Time

To achieve a fast response time, a capillary inlet is used with bypass pumping as shown in Figure 4. The system reduces the pressure in two stages. Most of the sampled gas is drawn through the capillary and directly to the diaphragm pump, bypassing the RGA. The pressure at the exit of the capillary is about 1 mbar. A small amount of the sampled gas is diverted to the RGA through an aperture. This configuration improves the response time in two ways. First, the pressure on the high side of the aperture is held to about 1 mbar. But even this pressure would give a time constant of 3.5 seconds in the 1/2 inch dead-volume example mentioned above. The second method to decrease the time constant is to ensure that any dead volume is well mixed. After the capillary, the gas is traveling at significant velocity (several meters per second). Proper layout of the inlet tubing will use the kinetic energy of the sampled gas to mix the dead volume (in a sense keeping the volume alive). Figure 5 shows the response to bursts of gas at the inlet of an atmospheric sampler designed with the above considerations. The sub-second response and cleanup are almost as fast as the RGA data rate.

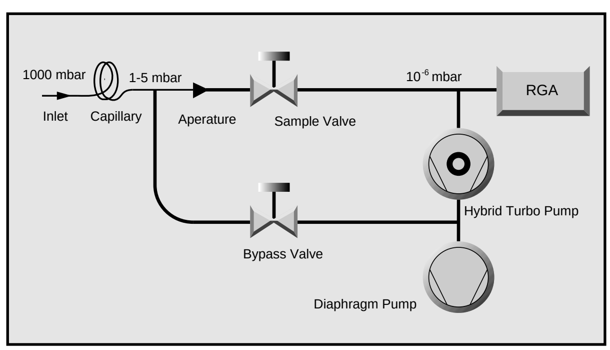


Figure 4: High Pressure Sampling Using Bypass Pumping

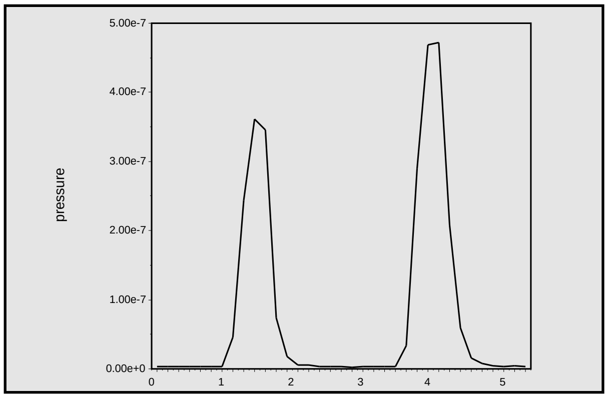


Figure 5: Response of Bypass Pumped System to Gas Bursts

Glass capillaries are available with small enough bores to reduce pressure from 1000 mbar to 10^{-6} mbar without bypass pumping. While it is possible to build an atmospheric sampling system based on a 1/4 meter 50 mm glass capillary, there are considerable reasons to use a bypass pump configuration. Bypass pumping improves the operation of a system by increasing the flowrate of gas through the capillary about 3-4 orders of magnitude. The higher flowrates and smaller pressure drop allow a wider selection of capillaries to be practical. Stainless steel and PEEK capillaries are more affordable and flexible than glass capillaries. A large flowrate means that the volumetric flowrate at the inlet of the capillary is more reasonable. For a system with 70 liter/s pumping speed, operating at 10^{-6} mbar, the volumetric flowrate at the inlet would be 70 nliter/s. Any dead volume at the inlet of the capillary would result in an unreasonable response time. With such small flowrates, inlet devices such as filters, valves, or connecting hardware cannot be used. Overall, the bypass-pumped capillary system is more flexible and only requires a minor addition of hardware (one valve and some tube).

The configuration seen in Figure 4 is made possible by the recent advances in hybrid turbomolecular/drag pumps and diaphragm pumps. Traditional designs would have relied on two rotary-vane pumps and standard turbomolecular pump. The high compression ratios of the hybrid turbo pumps allow the two stream (bypass and sample) to be combined. The low ultimate vacuum of contemporary diaphragm pumps makes them suitable as a foreline pump. The combination of these modern technologies means that an atmospheric sampling system can be constructed into very small packages (less than 8 inch high in a 19 inch rack mount chassis), which are portable and easy to operate.

Conclusion

Although the RGA is intrinsically a vacuum instrument, inlet systems are easily designed that allow it to sample gasses at any pressure. A more descriptive name for such systems would be “online quadrupole mass spectrometer”. Mass spectrometry is a well proven analytical technique, but traditionally used an expensive, large machine. Reduction in cost of quadrupoles and vacuum pumps, along with the development of easy to use software interfaces makes process analysis with mass spectrometry an attractive technique.

Application Note #9

Selecting the Right Quadrupole Gas Analyzer

Strict modern-day contamination control requirements for gas phase processes are constantly pushing the limits of performance of quadrupole gas analyzers. The quadrupole technology is rapidly evolving and adapting to lower contamination level specifications. A good understanding of the various factors affecting the detection capabilities of the different gas analysis systems currently available is an essential tool when selecting a sensor for a specific application. As is usually the case, most choices involve compromises, and a good understanding of the basic tradeoffs associated with different detector configurations will minimize mistakes and maximize productivity.

All gas phase processing setups can benefit from the addition of a quadrupole gas analyzer. The information delivered by a well-matched detector rapidly becomes an integral part of the process, and dramatically reduces the amount of guesswork that has traditionally been part of most vacuum troubleshooting procedures. As quadrupole gas analyzers become more affordable, they are rapidly becoming commonplace in all industries requiring the strict control of contamination levels in process gases. Smart software interface, lower detection limits and reduced cost of ownership are some of the features to look for in modern instruments.

The following sections of this article describe the performance specifications of open and closed ion source quadrupole mass spectrometers. The main objective of this information is to introduce the basic concepts required to choose the right analyzer for any gas phase application, and also to present some of the basic operating principles that must be kept in mind to assure the optimum performance of the instrument selected.

Residual Gas Analyzers

The prototypical residual gas analyzer (RGA) has an open ion source (OIS) and is mounted directly on a vacuum chamber so that the entire sensor is at the same pressure as the rest of the vacuum system. Small physical dimensions make it possible to attach an RGA to virtually any vacuum system, including both research and process setups. The maximum operating pressure is 10^{-4} Torr. Minimum detectable partial pressures (MDPP, typically measured for N_2 at 28 amu) are as low as 10^{-14} Torr for units equipped with an electron multiplier.

In high vacuum applications such as research chambers, surface science setups, accelerators, aerospace chambers, scanning microscopes, outgassing chambers, etc., RGAs are effectively used to monitor the quality of the vacuum and can easily detect even the most minute impurities in the low pressure gas environment. Trace impurities can be measured down to 10^{-14} Torr levels, and sub-PPM detectability is possible in the absence of background interferences. During system troubleshooting, RGAs are also used as very sensitive, in-situ, helium leak detectors.

In the semiconductor industry, RGAs are best used in evaporators, sputterers, etchers or any other high vacuum systems that are routinely pumped down to lower than 10^{-5} Torr. Their main application is to check the integrity of the vacuum seals and the quality of the vacuum before any wafers are committed to the process. Air leaks, virtual leaks and many other contaminants at very low levels can easily ruin wafers and must be

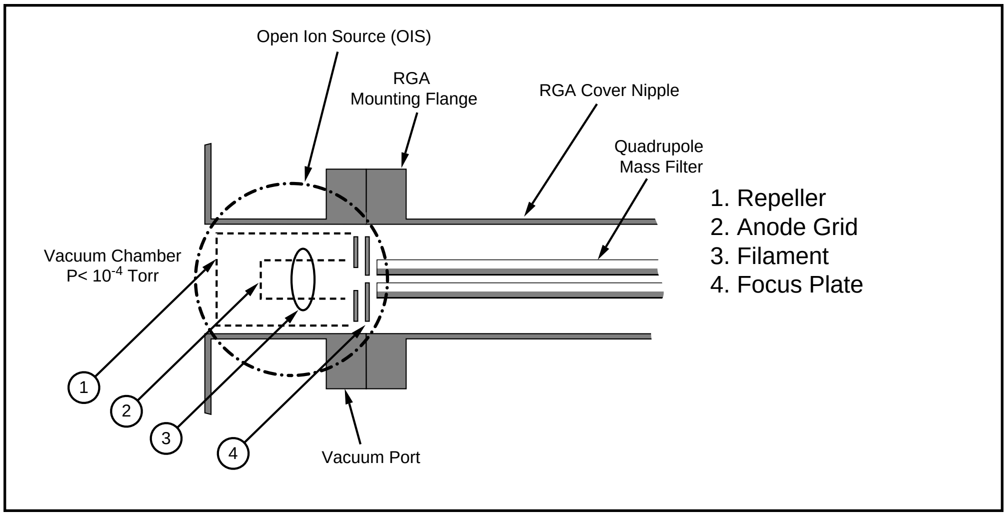


Fig.1:Schematic Diagram of OIS

detected before a process is initiated. As the semiconductor processes become more sophisticated, they also become less tolerant to contaminants. Residual gas analysis in a process chamber increases up-time and production yield and reduces cost of ownership.

The Open Ion Source (OIS)

The standard ion source used in most commercially available RGAs is the open ion source (OIS). This ionizer is considered the “do it all” source for RGAs. It has been around, in its cylindrical, axially symmetrical version since the early 1950’s. A schematic of a generic OIS design is shown in Figure 1.

The OIS penetrates into the process chamber. The filament wire and the anode wire cage are “open” to the surrounding vacuum chamber. All molecules that are present in the vacuum chamber can easily move through the ion source. The pressure in the ionizer is the same as in the rest of the surrounding vacuum and also the same as in the quadrupole mass analyzer and ion detector. The OIS is “open” to all the gaseous molecules in the vacuum chamber and can be used to monitor and detect changing gas levels as long as the overall pressure remains under 10^{-4} Torr. Higher pressures result in a decrease in sensitivity due to space charge repulsion between ions.

Performance Limitations of the OIS

OIS RGAs do an excellent job at measuring residual gas levels without affecting the gas composition of their vacuum environment. However, some potential problems must be kept in mind, particularly when the sensor is used routinely to monitor minute trace impurities (i.e. PPM and sub-PPM levels) or ultra high vacuum (UHV, $<10^{-9}$ Torr) environments.

The following is a list of the different ways in which an OIS RGA can contribute to its background signals, affecting the detection capabilities of the sensor. Methods to minimize these problems are described whenever applicable.

Outgassing

The OIS is a hot-cathode ion source. The filament wire (i.e. the cathode) must be heated to high temperatures (i.e.>1300°C) in order to establish an electron emission current. In the high vacuum, most of the energy required to heat the filament is dissipated to the surroundings through radiative processes. As a result, the entire ionizer and the adjacent walls “run hot”. The elevated temperatures result in increased outgassing from the OIS itself, and from the adjacent chamber walls. The gases emitted by outgassing can degrade the MDPP of the OIS RGA for many important species, including H_2, H_2O, N_2, CO and CO_2.

Outgassing from a hot cathode gauge is not a new problem for high vacuum users since it is also present in the Bayard-Alpert ionization gauges that have been commonplace in vacuum chambers for the last 50 years. In most cases, outgassing simply affects the composition of the gas mixture being measured. However, under some circumstances outgassing can be a serious problem and even affect the outcome of experiments or processes. Degassing the ionizer can help minimize some of the background signals; however, this usually only works as a temporary solution·

Some RGA vendors offer UHV versions of their OIS with anodes (and sometimes entire ionizer assemblies) made out of platinum clad molybdenum wire. This highly inert material exhibits decreased adsorption for many gases and provides reduced outgassing and ESD.

Water outgassing is a frequent interference, especially important because it is a serious source of contamination in many high vacuum processes. Overnight bakeouts at >200°C are the best option to minimize water outgassing from an OIS RGA.

H_2 outgassing from the OIS electrodes can be a concern for users operating in the UHV regime where residual hydrogen typically amounts to as much as 95% of the total gas mixture composition. H_2 is dissolved in most varieties of 300 series stainless steel and can readily outgas from the hot OIS electrodes. The contribution of the OIS to the H_2 background depends on its composition and can be dramatically reduced using platinum clad components. In all cases, the effect diminishes with time as the gas is depleted from the electrodes.

Electron Stimulated Desorption (ESD)

Even after an RGA has been thoroughly baked out, peaks are frequently observed at 12, 16, 19 and 35 amu, which are formed by ESD from surfaces within the OIS rather than by electron-impact ionization of gaseous species. ESD affects the RGA performance in a way similar to outgassing.

Several steps can be taken to minimize the effect:

* Degassing with high electron energies: Usually an option in commercially available instruments.
* Gold plating the ionizer: Decreases the adsorption of many gases and hence reduces the ESD effect. Using platinum clad molybdenum ionizers is also an alternative.
* Reducing the extent of the electron beam.
* Reducing the surface area of the OIS: For example, use wire mesh instead of solid perforated metal.
* Avoid exposing the ionizer to chlorinated and fluorinated compounds.

Background Interference

The quadrupole mass filter assembly has a large surface area in comparison to the ionizer and even though it does not get as hot as the ionizer during operation it can still outgas. The fact that the OIS is exposed to the same vacuum environment as the rest of the sensor makes the ionizer sensitive to the impurities outgassed by the rest of the quadrupole assembly. A serious problem for a lot of RGA users (particularly in the UHV range) is H_2O outgassing from unbaked RGAs. However, lots of other species can also affect the background readings. For example, high Ar backgrounds can be expected if the sensor was recently exposed to large levels of the gas since it tends to get adsorbed on SS surfaces and desorbs only very slowly.

The ionizer is also sensitive to impurities generated at the hot filament. Gas molecules can suffer thermal cracking and chemical reactions at the filament surface and the products of the reaction can easily find their way into the ionization region. The impurities generated in this fashion are usually an important source of contamination of the ionizer's surfaces and have a serious effect on the RGA's long term stability. For example, CO and CO_2 are emitted by most hot filaments and easily find their way into the ionizer and vacuum system.

Regular bakeouts are the most efficient way to minimize this problem. An overnight bakeout at 200°C will usually take care of most contamination problems. If the problem persists it might be necessary to clean and/or refurbish the quadrupole sensor.

Partial Pressure Reduction (PPR) systems

RGAs are not limited to the analysis of gases at pressures below 10^{-4} Torr. Higher gas pressures can be sampled with the help of a differentially-pumped pressure reducing gas inlet system (PPR), consisting of a restriction and a vacuum pump package. Common restrictions are pinholes and capillaries, which can provide pressure reductions of more than six decades of pressure. Vacuum pump packages typically consist of a turbomolecular pump backed by a foreline pump. The combined RGA, gas inlet system and pumping station constitute what is usually referred to as a Partial Pressure Reduction (PPR) System. These gas sampling systems are commonplace in gas phase processes and are available from several RGA vendors. Properly designed PPRs can monitor processes from beginning to end, providing essential information every step of the way.

The PPR system depicted in figure 2, is an example of a typical pressure reduction setup used to step process pressures down to levels acceptable to the OIS RGA. The PPR contains two inlet paths to the RGA: a high conductivity path (hi-C) for monitoring base vacuum and a low conductivity path (i.e. pressure reducing bypass loop, Lo-C) for monitoring gases at operating pressure.

The high conductivity (Hi-C) path is used when the vacuum system is at pressures below 10^{-4} Torr. At high vacuum, typical applications are leak testing and monitoring the ultimate vacuum of the chamber. For example, in a sputtering chamber, the first stage of the process is a pump-down to <10^{-6} Torr. At this point the RGA may be used to check the quality of the background for leaks and contaminants. Once the quality of the vacuum is satisfactory the sputtering chamber is backfilled with argon at a few mTorr and sputtering is started.

The low conductivity (Lo-C) path is used when the process chamber is at pressures above 10^{-4} Torr. This path contains a micro-hole orifice which reduces the pressure several decades to a level suitable for the RGA (typically around 10^{-5} Torr). Apertures are available for operating pressures as large as 10 Torr. An array of apertures or an adjustable metering valve are sometimes used to adjust the pressure reduction factor to different pressures along the process. For example, during a sputtering process the Lo-C path may be used to monitor water vapor and hydrocarbon levels to assure they do not exceed certain critical levels that degrade the quality of the sputtered films.

A pair of pumps draws the gas through the aperture to the RGA establishing the pressure drop. The pumps used in these systems are usually very compact, oil-free and low maintenance.

For pressures higher than 10 Torr, the gas flow rate into the sample inlet side of a single-stage PPR such as shown in Figure 2 becomes extremely small and the time response is too slow for any practical measurements. In those cases, a dual-stage bypass pumped gas-sampling system, with a much larger gas flow rate and faster response, is a much better choice than a single-stage PPR. Bypass pumped gas-sampling systems, with medium pressure intermediate stages, capable of analyzing gas mixtures up to several atmospheres, are available from several RGA vendors.

Performance limitations of PPR systems

PPRs do an excellent job sampling gases at pressures below 10 Torr. The information they provide is routinely used to diagnose and control gas phase processes in a large variety of industries. As prices drop and the technology evolves, the instruments are continuously finding new fields of application.

A large number of PPR systems are dedicated to the detection of trace impurities in gas mixtures. OIS RGAs have adequate sensitivity and dynamic range to detect part-per-million (PPM) level contaminants in principle; however, interferences from process gases and background interferences from the sensor itself can make the detection of PPM levels of impurities with a PPR difficult in practice.

Background interferences

The background gases present in the analyzer chamber can obscure the MDPPs of some important gases (particularly: H_2, H_2O, N2, CO, and CO_2). Background gases are due to outgassing, electron stimulated desorption, and the finite compression ratio of the pumping system.

In order to best illustrate this point, lets consider, as an example, the analysis of water in a 10^{-2} Torr Ar sputtering process. During process monitoring, the mass spectrometer typically runs at about 10^{-5} Torr, corresponding to a three-decade reduction factor across the Lo-C path of the PPR. The pressure drop, brings 1 PPM of water in the process chamber to a partial pressure in the mass spectrometer of about 10^{-11} Torr (well within the detection limit of a typical RGA). However, with the mass spectrometer isolated from the process gases, the residual pressure in the PPR chamber is, at best, in the order of 10^{-9} Torr with most of that being water. This water level is one hundred times larger than the 10^{-11} Torr corresponding to a PPM of water in the process chamber, meaning that the water vapor concentration cannot be reliably detected or measured to better than 100 PPM under these "common" operating conditions.

The MDPP limit could be improved to 20 PPM increasing the operating pressure in the RGA chamber to 5×10^{-5} Torr during analysis. However, even a 20 PPM MDPP limit for water might not be low enough in some cases. The addition of a cryopump, with a large pumping speed for water, has been proved to dramatically minimize the water background in the PPR's quadrupole chamber; however, this is rarely done in practice because of the high cost of the pump. The same limita-

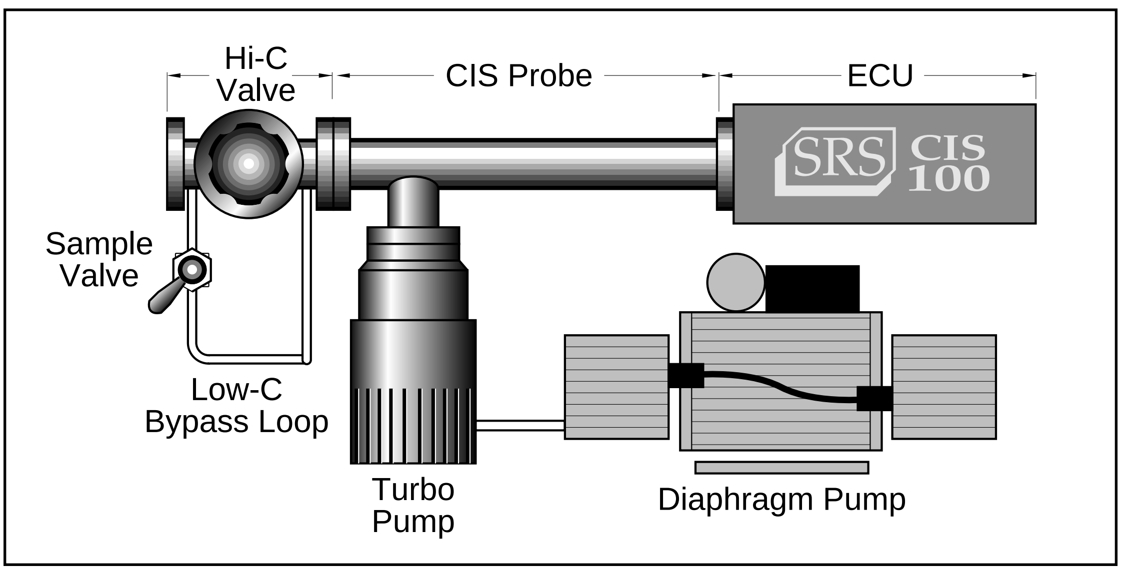


Fig.2:PPR Inlet System Components

tions must be kept in mind for other potential interfering gases. In order for any species to be detectable at the PPM level (i.e. 10^{-8} Torr in a 10mTorr process) the residual mass spectrum for the PPR must show pressure readings of less than 10^{-11} Torr at the mass values corresponding to the peaks of that species. Such levels are not easily achieved repeatedly in most vacuum system unless the necessary precautions are taken to minimize all sources of contamination. The problem is usually more serious for masses under 50 amu where there are always background peaks in the residual mass spectrum.

Even though the RGA is intrinsically capable of performing sub-PPM measurements, it is not always easy to find places in the residual mass spectrum of the RGA where the background is at PPM levels.

A common source of background interference in PPRs is contamination from pump oil backstreaming into the PPR chamber from conventional, oil-based, roughing pumps. Switching to a completely oil-free pumping station eliminates this problem.

The MDPP limit for air is usually limited by the compression ratio of the pumping station. In most PPR systems, the N_2 levels are usually under 10^{-9} Torr, with oxygen levels approximately five times lower. This corresponds to MDPP levels of better than 20 PPM for N_2 @ 28 amu and 4 PPM for O2 @ 32 amu in a 10 mTorr process.

Hydrogen is generally impossible to detect at PPM levels because it outgasses readily from the analyzer and it is not effectively pumped by most turbo pumps. Some of the tricks that are used to minimize the H_2 background signals include using a Pt Clad Molybdenum OIS and the addition of a special pumping station with increased pumping speed for hydrogen.

Process Gas Interference

The other limitation to PPM detection levels in a typical OIS RGA based PPR system is caused by interference from the same process gases that are being analyzed. The best way to illustrate this point is to go back to the example of water analysis in the 10 mTorr Ar sputtering process. We saw that detecting water at better than 20 PPM levels is very difficult unless the PPR chamber is very carefully baked out and protected from water contamination. However, as we will see, this is only part of the problem, there is also a serious interference at m/e 18 from the same Ar used in the sputtering system. The isotope ^{36}Ar is present at 0.34%. In the electron ionization process, doubly charged argon is formed leading to peaks at m/e 20 ($^{40}Ar^{++}$) and m/e 18 ($^{36}Ar^{++}$). For 70 eV electron impact energy, a typical level of $^{36}Ar^{++}$ is 350 PPM. Therefore, if you want to detect PPM levels of water in an Ar based sputtering system, you must solve two problems:

1. Background contribution of water outgassing from the sensor
2. Interference at m/e 18 from $^{36}Ar^{++}$.

A thorough bakeout can reduce the background water contribution to the low tens-of-PPM levels, but eliminating the 36Ar^{++} interference requires the use of several tricks. Some manufacturers simply choose to monitor the m/e 17 peak due to the $[OH]^+$ water fragment. For 70 eV ionizing electrons, this peak is four times smaller than the main one at 18 amu. This results in a significant reduction in sensitivity for water detection and also adds the problem of abundance sensitivity while trying to measure the mass 17 intensity next to a large $^{36}Ar^{++}$ peak at 18 amu.

A better option (and the one recommended for RGAs with programmable ionizer voltages) is to operate the ionizer with the electron impact energy reduced to <40eV. This ionization energy is below the appearance potential (i.e. 43.5eV) of Ar^{++}. For example, the peaks at masses 18, 19 and 20 due to Ar^{++} disappear while operating an RGA with 35 eV electrons, and this is achieved with minimal reduction in the sensitivity of detection of Ar^+ at 36, 38 and 40 amu.

Different electron ionization energies are routinely used to selectively ionize species in a gas mixture. Tables with ionization potentials for many different gases are readily available from the general mass spectrometery literature. Reduction of the electron energy usually imposes an extra load of work on the filament and can reduce its lifetime. However, the reduced interference effects offset the extra costs of filament replacement.

The Closed Ion Source (CIS)

In applications requiring the measurement of pressures between 10^{-4} and 10^{-2} Torr, the problem of background and process gas interferences to the mass spectra can be significantly reduced by replacing the traditional OIS PPR configuration described above with a closed ion source (CIS) sampling system. A cross section of a generic CIS setup is shown in figure 3.

The CIS Ionizer sits on top of the quadrupole mass filter replacing the more traditional OIS used in conventional RGAs. It consists of a short, gas-tight tube with two very small openings for the entrance of electrons and the exit of ions. Electrons enter the ionizing region through an entrance slit of small dimensions. The ions are formed close to, and attracted by, a single extraction plate and exit the ionizer through a circular aperture of small diameter. Alumina rings seal the tube from the rest of the quadrupole mass assembly and provide electrical insulation for the biased electrodes. Ions are produced by electron impact directly at the process pressure.

A pumping system, similar to the one used in PPR systems, keeps the filament and the rest of the quadrupole assembly at pressures below 10^{-5} Torr through differential pumping (i.e. two decades of pressure reduction). The design is very simple and was successfully applied for many years to gas chromatography-mass spectrometry instruments before it was adopted by quadrupole gas analyzers. Most commercially available CIS systems are designed to operate between 10^{-2} and 10^{-11} Torr, and offer PPM level detectability over the entire mass range for process pressures between 10^{-4} and 10^{-2} Torr.

Performance Differences Between the PPR and CIS Systems

An understanding of the performance differences between the CIS setup and the more traditional OIS RGA based PPR is indispensable when choosing the sensor setup that is best suited for a particular process application. Process engineers should carefully weight all differences before selecting an analyzer configuration for their application.

Direct Sampling

The CIS Anode can be viewed as a high conductance tube connected directly to the process chamber. The pressure in the ionization area is virtually the same as that in the process chamber. The CIS Ionizer produces ions by electron impact directly at process pressure while the rest of the mass analyzer and the filament are kept under high vacuum. Direct sampling provides good sensitivity due to the large ion densities available and also fast response times. The “memory effects”, typically associated to pressure reduction and conductance orifices, are significantly reduced. Also, the fractionation effects due to the molecular weight dependent diffusivities of the different gas molecules through the PPR apertures are absent.

Signal-to-Background Ratios

Because the sampling pressure in the CIS is typically two decades higher than that of the rest of the sensor's vacuum system, the signal-to-background ratio is significantly increased relative to the OIS PPR systems. This is particularly important when measuring common residual gases, such as water. In order to illustrate this point, we go back to the water measurement example

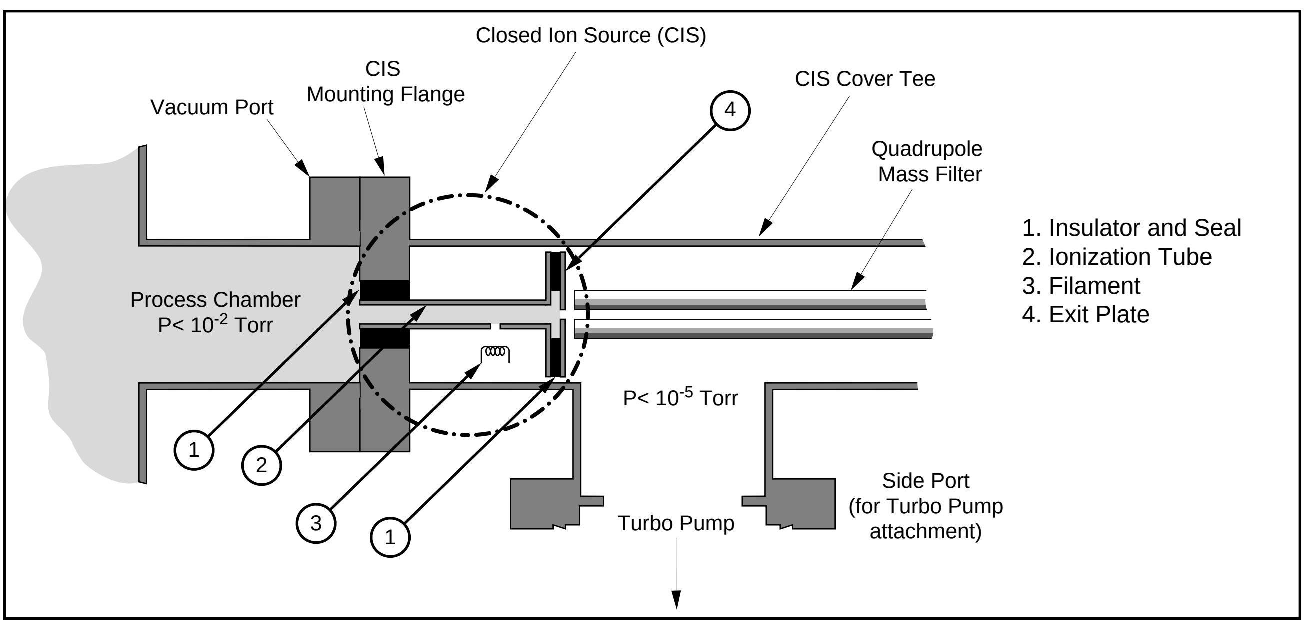


Fig.2:Schematic Diagram of an OIS

in a 10^{-2} Torr Ar sputtering process. The Ar gas is ionized directly at 10^{-2} Torr (i.e. three orders of magnitude higher than in the OIS PPR!) but in the same background (10^{-9} Torr) of residual water. This residual water signal now corresponds to a 100 PPB MDPP level for water in the CIS system. This is quite an improvement over the OIS PPR performance!

The combination of direct sampling and differential pumping provides the potential for PPM and sub-PPM detection limits for even the most pervasive residual gases. For other common interferences, such as organic contaminants or reaction by-products of the filament, the gas tight design of the source reduces the visibility of the ionization region to those gases providing a very clean residual gas spectrum, free of many of the spectral overlaps that are common in OIS PPR setups.

Interference from contaminants generated by ESD is also reduced in the CIS because a much smaller electron beam penetrates the ionizing volume. In addition, the inside walls of most commercially available CISs are coated with highly inert materials such as gold, platinum clad and pure molybdenum which adsorb less impurities than stainless steel.

The ability of the CIS to sample gases directly in the mTorr range and to provide PPM level detectability across its entire mass range has made CIS systems the instrument of choice in semiconductor processing applications such as PVD, CVD and etching.

Ionizer Contamination

In an OIS PPR system, sample molecules that have suffered thermal cracking or chemical reaction at the filament, are free to drift into the ionization region. This is a very significant source of surface contaminants for electron impact ionizers. In contrast, the gas tight design of the CIS reduces the visibility of the source to those contaminant gases, providing reduced contamination and better long term stability. Most CIS manufacturers utilize exclusively Tungsten filaments in their systems. W resists many corrosive gases (such as WF6) and reactive gases (such as Silane) minimizing reactions at the filament that contribute to the background, also resulting in extended filament lifetime.

Versatility

When properly matched to a process, both OIS PPR and CIS systems are very versatile instruments that provide crucial information throughout an entire gas phase process.

A PPR system fitted with a dual path gas inlet can switch effortlessly from a highly sensitive RGA mode of operation to a Process Monitoring mode by simply switching from the Hi-C to the Lo-C sample paths.

Different modes of operation can also be easily achieved in a CIS by simply changing some of the sensor's ionization parameters. A CIS Gas Analyzer, even though not as sensitive as an RGA, can tackle most residual gas analysis and leak checking tests that are required in process chambers. The sensitivity of the CIS is reduced over the OIS because of the very small holes for electron entrance and ion exit. However, in most cases running the electron multiplier at higher gain levels than the RGA makes up the reduction in sensitivity. Typical MDPP values for CIS systems, fitted with an optional electron multiplier and operated in the RGA Mode, are in the order of 10^{-11} Torr. This is about two decades higher than the MDPP values that can be achieved with PPRs operated in the RGA mode with the Hi-C sampling path open.

The CIS ionizer can also be reconfigured for on-line process monitoring and control and verification of process gas purity at the point of use. The electron emission current is raised during residual gas analysis to increase sensitivity, and reduced during process monitoring to avoid space charge saturation effects in the ionizing volume at the higher pressures.

The tight design of the CIS makes it possible to operate the ionizer at lower electron ionization energies than are possible with OISs. Most of the commercially available CIS systems offer at least two electron energy settings of 70 and 35 eV. The 70eV setting is mostly used for leak testing and routine gas analysis. The spectra collected are virtually identical to those obtained with standard RGAs. The 35eV setting is used during process monitoring to eliminate process gas interference peaks. A common application of the low energy mode is to the elimination of the doubly ionized $^{36}Ar^{++}$ peak that interferes with water detection at 18 amu in sputtering processes. CIS systems with user programmable ionizer voltages offer the highest versatility, since they can be configured to selectively ionize species in a gas mixture by carefully adjusting the electron impact energy.

High Pressure Sampling with a CIS Gas Analyzer

CIS Analyzers can sample gases directly up to about 10^{-2} Torr pressure levels. The upper pressure limit is set by the reduction in mean free path for ion-neutral collisions, which takes place at higher pressures, and results in significant scattering of ions and reduced sensitivity. However, operation is not limited to the analysis of gases at pressures below 10^{-2} Torr. Higher gas pressures can be sampled with the help of a differentially pumped pressure reducing gas inlet system (PPR) just as it is done with conventional RGAs. A pressure reducing gas inlet system matched to the conductance of the CIS analyzer will allow the sensor to sample gas pressures as large as 10 Torr. As in the case of the PPR systems, the penalty paid is reduced sampling speed, fractionation of the gas mixture at the sample inlet and possible memory effects at the ionizer.

For pressures higher than 10 Torr, the gas flow rate into the closed ionizer becomes extremely small and the time response is too slow for any practical measurements. In those cases, a bypass pumped gas-sampling system, with a much larger glass flow rate and faster response, is a much better choice than a single restriction into the CIS ionizer.

Conclusions

Any vacuum processing setup can benefit from the addition of a quadrupole gas analyzer. A good understanding of the different factors affecting the performance of the different quadrupole gas analysis systems currently available, is an essential tool when selecting the best sensor configuration for any application. Quadrupole gas sampling systems are available from several different manufacturers and it is often difficult to decide which one constitutes the best match for a process. In most cases, there is more than one way to set up the measurements, and each choice involves compromises. A good understanding of the basic differences between the available options minimizes problems and maximizes productivity.

As quadrupole gas analyzers become more affordable, they will become commonplace in all industries requiring the strict control of contamination levels in process gasses. It is the responsibility of the manufacturers of quadrupole gas analyzers to keep vacuum users and process engineers informed about the different sensor options available and the new developments in the technology.

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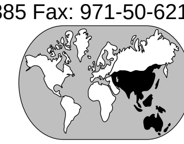

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