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Precision Time-Domain Measurements using the Agilent E1430A

Product Note E1430A-1

Analog to digital conversion of signals followed by some form of computer based digital signal processing is now a common part of signal parameter extraction. One of the growing challenges in this parameter extraction technique is how to achieve the most precision from the ADC at hand. If you have a linear enough digitizer it is possible to increase sample precision significantly by filtering to reduce noise then correcting the sampled data to compensate for filter effects.

This paper discusses using the Agilent E1430A VXI A/D, its built in filters, and compensation to perform high precision time domain digitizing. Three applications are addressed. The first is precision waveform characterization for measuring rise time, flatness, peak occurrence, zero crossing, etc., of a signal. The second is precision pulse capture for determining pulse timing information such as pulse delay, pulse to pulse interval, and pulse arrival time. The final application addresses precision single sample amplitude capture, a special case of the pulse capture scenario. We begin with a review of some ADC basics.

Sample Rate, Noise and Sample Precision

Anyone who has selected an ADC for a particular application is aware of the trade-off between sample rate and precision. Low sample rate converters can typically achieve better signal to noise ratios, and therefore make use of more bits of resolution, than high sample rate converters. The primary reason for this is, regardless of how carefully the ADC is designed, its signal to noise performance is ultimately limited by analog mechanisms such as thermal noise and shot noise. These produce random errors with root-mean-square (RMS) levels proportional to the square root of the analog signal bandwidth. Since low sample rate ADCs are typically used to acquire signals containing only low frequencies, they are implemented using low bandwidth analog circuitry which is inherently lower noise than wide band circuitry.

The RMS noise level present in any ADC can be described in terms of sample precision by using “equivalent bits” as defined by the following.

Equivalent Bits = (SNR - 1.76) / 6.02

SNR is the ratio of the RMS noise level to a full scale signal, calculated in decibels. For example, if an ADC has a 25 micro volt RMS noise floor, on the 1 volt peak (full scale) input range, the equivalent bits will be:

Vrms	= V peak x 0.707
	= 707 mVrms
SNR	= 20 log (25 mVrms/707 µVrms)
Equivalent bits	= -89 dB
	= -89 dB - 1.76 / 6.02
	= 15.08 bits

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Caution must be used when comparing "equivalent bits" to the "bits of resolution" listed on manufacturers' data sheets. Bits of resolution merely indicate how many binary 1's and 0's are used to encode the digitized samples. The number of equivalent bits available, after taking analog noise into account, is always less than the bits of resolution. Conversely, the more equivalent bits in a sample, the lower the noise in the measurement, and the better its precision.

Normally ADCs are designed with input bandwidths comparable to the maximum signal frequency they can characterize without ambiguity (i.e. aliasing) at their specified maximum sample rate. According to the Nyquist sampling criteria this maximum signal frequency is one-half the ADC sample rate. This means the RMS noise level is roughly proportional to the square root of the specified maximum sample frequency. The challenge for an ADC user, then, is selecting a digitizer with sample rate and bandwidth sufficient for the application but limited enough to avoid the noise added by excess bandwidth.

Many ADCs have provisions to reduce their sample rate either through internal sample rate control circuits or external clock inputs. Reducing sample rate without reducing bandwidth does nothing to reduce the RMS noise level on the samples and thus does not improve sample precision.

Reducing bandwidth must be approached with some thought. Simply passing the signal through a filter before it is digitized will not increase sample precision. The filter removes noise from the signal but does not address the noise contributed by the analog mechanisms in the ADC.

Table 1.
Digital filter table for the E1430A

Bandwidth	Equivalent Bits	Sample Rate
4 MHz	11.3	10 MHz
2 MHz	11.8	5 MHz
1 MHz	12.3	2.5 MHz
500 KHz	12.8	1.25 MHz
250 KHz	13.3	625 KHz
125 KHz	13.8	312.5 KHz
62.5 KHz	14.3	156.25 KHz
31.25 KHz	14.8	78.125 KHz
15.6 KHz	15.3	39.1 KHz
7.81 KHz	15.8	19.5 KHz
3.91 KHz	16.3	9.77 KHz
1.95 KHz	16.8	4.88 KHz
977 Hz	17.3	2.44 KHz
488 Hz	17.8	1.22 KHz
244 Hz	18	610 Hz
0.238 Hz	18	5.96 Hz

However, passing the samples through a digital filter after digitizing will increase precision. Putting the filtering after the ADC reduces the noise contributed by the ADC, as well as by the signal, and has the same effect as reducing the analog input bandwidth of the ADC.

The digital filtering required to enhance sample precision can be readily designed using various filter design programs. These filters can be implemented in a PC, or workstation, or in hardware if speed is required. The E1430A provides a set of 24, pre-programmed, user selectable filters. Bandwidths start at 2 MHz and go down to 0.24 Hz in octave steps. Because these filters are implemented in hardware they execute in real-time which means they can keep up with the ADC sample rate. Each filter includes decimation which keeps the effective sample rate near the Nyquist rate of the filter upper cut-off frequency. This is important for alias free, and data efficient, sampling. Table 1 lists some of the sample rate, bandwidth, and signal-to-noise combinations provided by the E1430A. A complete listing is provided in appendix 1.

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Linearity versus Equivalent Bits

Figure 1 graphs the equivalent bits of resolution versus bandwidth data in table 1. Note that for bandwidths down to 488 Hz, sample precision increases at 0.5 bits per octave. However, for bandwidths of 244 Hz and below there is no further improvement.

The equivalent bits column in table 1 is dominated by the effects of random noise for bandwidths of 488 Hz; and higher. For bandwidths of 244 Hz and below the accuracy is limited by systematic errors such as harmonic distortion or spurious signals in the ADC. In the case of the E1430A these systematic errors are small compared to the full bandwidth noise (figure 2a). However, in most traditional ADCs the systematic errors are often comparable to the noise level of the converter before any kind of filtering (figure 2b). The prevailing philosophy is that any systematic error which is smaller than the sample-by-sample random fluctuations may be ignored, since it will be masked by the noise. To take advantage of filtering to increase precision we must abandon this philosophy.

To improve sample precision with filtering, systematic ADC errors must be small compared to the noise level of the narrowest bandwidth to which the signal is to be filtered. In the E1430A these errors (quantified as harmonics of an applied sine wave) can be as small as -110 dB relative to a full scale sinusoidal input signal (dBFS). This degree of linearity is comparable to that expected from an 18-bit ADC. Thus, the equivalent bit improvement versus bandwidth shown in table 1 can be extended all the way down to bandwidths as narrow as 488 Hz.

Figure 1
Agilent E1430A filter bandwidth versus resolution

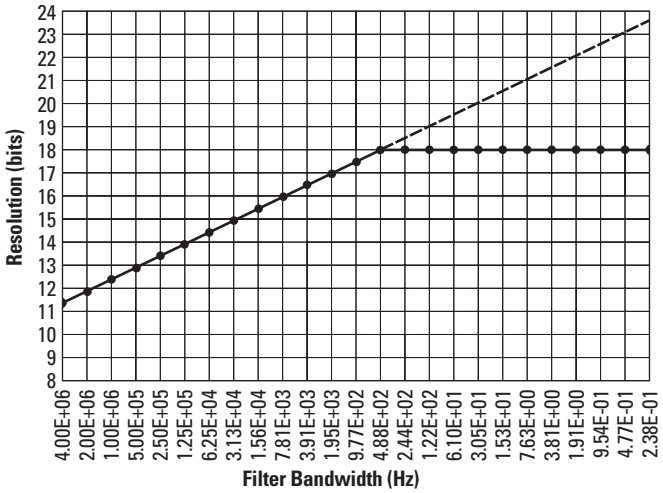


Figure 2a
Sketch of E1430A Noise and Distortion. Full bandwidth noise level = -70 dBFS (11.3 bits), Linearity ≥ -110 dBFS (18 bits)

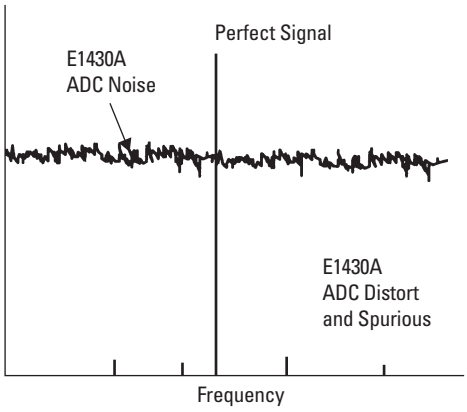
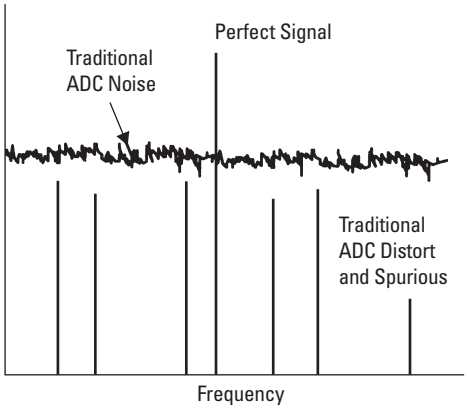


Figure 2b
Sketch of traditional ADC noise and distortion



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Other Techniques for Improving Precision

Filtering is just one digital signal processing technique that can improve sample precision if systematic errors are kept low. Examples of other techniques which improve precision by reducing random noise include: averaging, curve fitting, and Fourier transforming.

Averaging is a common processing technique for increasing measurement precision. It requires multiple measurements be made to collect multiple sets of data corresponding to the same event. The measurements are added together on a sample by sample basis and the result is divided by the number of data sets. This processing reduces the RMS deviations due to additive random noise (with a gaussian distribution) by a factor equal to the square root of the number of data sets, or averages. Thus, averaging 16 measurements increases precision by a factor of four. This is the same effect as adding two equivalent bits of resolution to the measurement.

Curve fitting does not require multiple data sets to derive its values, and it can increase the precision of a single set of data samples. In figure 3 the jagged line represents the samples collected from a single digitizing of a stepped voltage. The noise in the data simulates an ADC with an RMS noise level which is a significant fraction of the voltage being measured. A damped complex exponential curve fit was performed on the data set. The solid line shows the results of the curve fit. In this processing many samples are used to compute each point on the curve. This reduces the randomness in the data in much the same fashion as averaging.

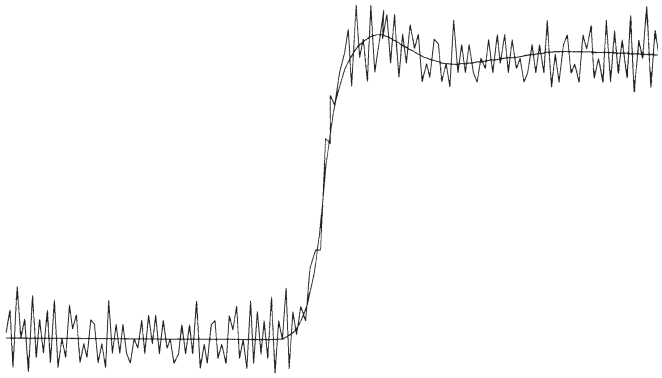


Figure 3
Curve fit of noisy data to produce response model

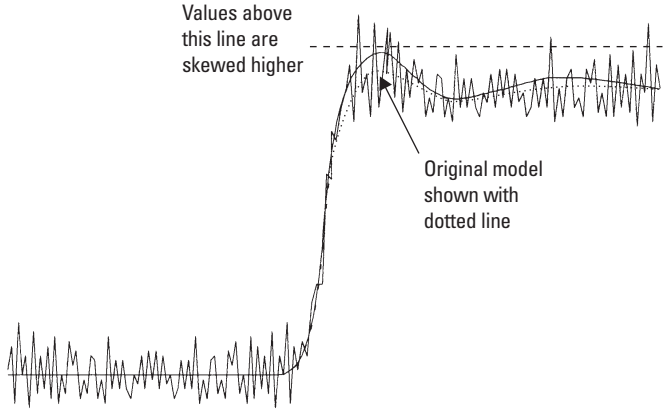


Figure 4
Curve fit of noisy data with small systematic error to produce exaggerated ringing in the response model

Performing a Fast Fourier Transform (FFT) on a set of data also increases the precision of the final result. The FFT breaks the measurement bandwidth into bins, or lines, of frequency resolution. Thus an 800 bin FFT, performed on a signal measured with a 4 MHz bandwidth, has 800 bins each 5 kHz wide. This division of the measurement bandwidth divides the noise among the bins. The noise in each bin is proportional to the width of the bin. In the example dividing the noise across 800 bins provides an improvement of 29 db (10 log (5 kHz / 4 MHz)), or 4.5 bit, in precision. If the number of samples in the data set is doubled the FFT bins of resolution will double. This halves bin width, drops the noise in each bin by 3 dB, and improves precision the equivalent of 0.5 bit.

Both curve fitting and averaging are sensitive to systematic errors in the sampled data. Figure 4 shows the impact when a systematic error in the ADC skews a few sample values upward. The dotted straight line shows the affected samples. The skew is small compared to the RMS noise level and involves only a few samples. Unfortunately systematic errors tend to reinforce. This amplifies the apparent amount of ringing on the signal, and causes the curve fitting algorithm to generate a model with exaggerated ringing.

As with filtering, to capitalize on the improvements available from signal processing, systematic errors must be kept small compared to the noise level at the ADC bandwidth used to acquire the samples.

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Precision Waveform Capture with Filtering

We have seen it is possible to dramatically increase sample precision with appropriate filtering to decrease measurement bandwidth. This filtering, however, also affects the resulting samples. The challenge for the user is to understand how the filter affects the samples and take advantage of signal processing to compensate for those effects.

Even without additional filtering every ADC has some finite analog signal bandwidth, or rise time. When you use an ADC to measure a waveform, the observed rise time will be approximately equal to the square root of the sum of the squares of the waveform rise time and the ADC rise time.

$$\text{Measured Rise Time} \cong \sqrt{(\text{Waveform rise time}) + (\text{ADC rise time})^2} \quad (3)$$

For example, if you use an ADC with a 87.5 ns rise time to measure a step with a 1 microsecond rise time, the result will be:

$$\text{Measured Rise Time} \cong \sqrt{(1\ \mu s)^2 + (0.087\ \mu s)^2} = 1.00378\ \mu s$$

The error of the measurement is less than 0.4%. Table 2 shows the relationship between the ratio of the ADC rise time to the waveform rise time, and the resulting error in the rise time measurement.

An initial examination of table 2 indicates that to achieve a rise time measurement error no worse than 0.5% the filter rise time must be at least 10 times faster than the waveform rise time. Such wide filtering will result in loss of precision due to a high RMS noise level. It is possible to achieve low rise time error with significantly narrower filters if suitably stable filtering is used and the sampled data is corrected for the effects of the filtering.

Table 2
Rise time measurement error for various ratios of waveform to ADC rise time

ADC rise time / Waveform rise time	Measurement Error Percentage
3	5
4	3
5	2
10	0.5
20	0.1

Finding the Correct Filter Bandwidth

In the frequency domain the ideal filter for maximizing precision is wide enough to pass all the frequency components contained in the waveform but no wider. Describing the characteristics of such a filter in the time domain is more difficult. The faster the waveform rise time and/or the sharper the corners at the transition points the wider the waveform's bandwidth. If the waveform has a gaussian shape the following equation can be used to estimate its bandwidth.

$$\text{Bandwidth} \cong 0.35 / \text{Rise time}$$

To select the narrowest filter bandwidth available in the E1430A for digitizing a 200 ns rise time gaussian shape waveform we calculate:

$$\text{Min. Filter Bandwidth} \cong 0.35 / \text{Rise Time} = 0.35 / 200\ ns = 1.75\ MHz$$

From table 1 we see that the E1430A does not have a 1.75 MHz filter. The closest larger filter is 2.0 MHz. The equivalent bits of resolution expected using this filter is 13.3 bits. The sample rate for this filter is 5.0 MHz. Since this is better than twice the highest frequency in the waveform it will be Nyquist sampled.

Nyquist's sampling criteria states that a signal is completely described if sampled at a rate at least twice the highest frequency of interest. The resulting samples provide a complete representation of the analog signal. Any attribute of the signal can be computed from these output samples, including: peak voltage, time of the peak occurrence, zero crossing, and rise time, using any of a number of interpolation algorithms. The same technique can be used to display or plot the signal with arbitrarily high detail. The C library of functions supplied with the E1430A contains an algorithm that will generate the detail necessary for such a display.

The 0.35 factor relating bandwidth and rise time is a good estimator of bandwidth only if the waveform has a gaussian shape in the time domain. It is not a valid estimator for waveforms with even moderately sharp corners. A more general technique for estimating the bandwidth of a waveform is to sample it with the widest bandwidth available and FFT the results. The appropriate filter bandwidth can be read directly from the FFT results. If the FFT results reveal frequency components as high as one-half the sample rate of the digitizer a wider bandwidth digitizer should be selected. Another technique for determining the bandwidth of nongaussian waveforms is to digitize the waveform using a number

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of successively narrower filters until the reconstructed waveform starts to change shape. This technique is convenient when a number of filters are readily available, as they are in the E1430A.

While filtering the waveform increases precision without inhibiting Nyquist sampling, it can cause errors in the samples. The analog and digital filters in the E1430A have relatively sharp cutoffs and a flat passbands to provide maximum alias free signal bandwidth. The nonflat group delay characteristics of these filters affect the sample values even though they do not remove any of the signal energy. These effects can be corrected by passing the samples through the compensation filters provided in the software library supplied with the E1430A.

The filters in the compensation filter library supplied with the E1430A are designed to maximize flatness in the lower 87.5% of the pass band. This means the bandwidth of the waveform must be less than 87.5% of the bandwidth selected filter. At frequencies above the waveform bandwidth the phase response of the compensated filter is unimportant since it only effects added measurement noise and has no impact on the signal. The magnitude of the compensated filter response at frequencies above the waveform bandwidth rolls off to reduce added measurement noise.

Figure 5 shows the group delay of the 4 MHz analog alias filter in the E1430A before any compensation is applied. Note that the various signal components between DC and 3.5 MHz experience delays ranging from 29 ns to 66 ns. The difference of 37 ns represents a significant phase shift for the higher frequency components and thus produces a noticeable dispersion in time of an input signal, even if it is band limited to less than 3.5 MHz. Figure 6 shows the resulting group delay after the application of the compensating digital filter. Note that all frequency components less than 3.5 MHz experience the same delay of 120 ns.

Figure 5
Delay vs. frequency for uncompensated 4 MHz alias filter. Vertical scale is 30 ns/div. Horizontal scale is 0.5 MHz/div

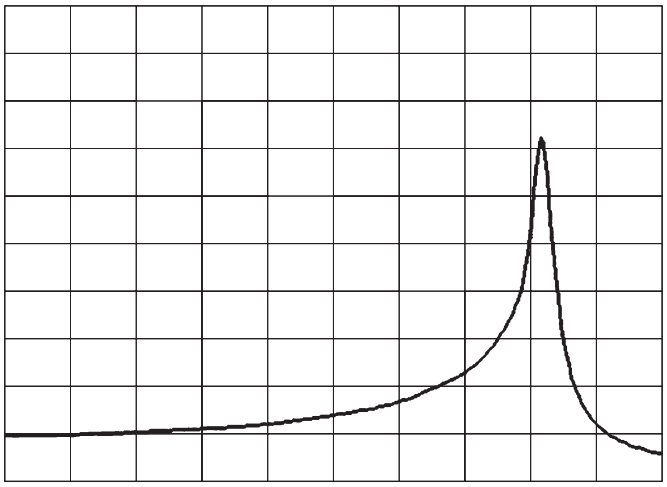


Figure 6
Delay vs. frequency for compensated 4 MHz alias filter. Vertical scale is 30 ns/div. Horizontal scale is 0.5 MHz/div

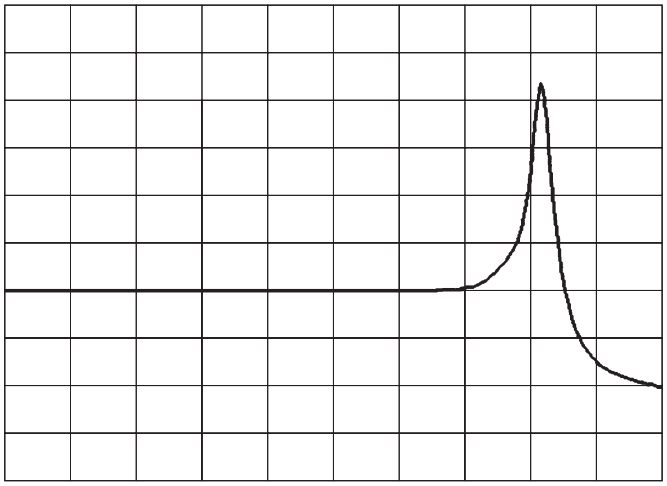
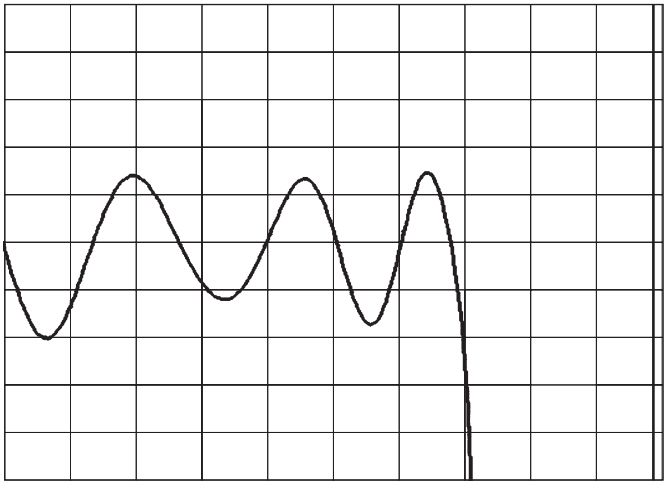


Figure 7
Deviation from linear phase for compensated 4 MHz alias filter. Vertical scale is 0.1°/div. Horizontal scale is 0.5 MHz/div



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Because a constant time delay implies a phase shift which is linear versus frequency, an alternative way to quantify the delay variations is to plot the deviation of the phase response from this ideal linear phase ramp. Figure 7 shows that for the compensated filter, the phase is within 0.2 degrees of ideal for signals up to nearly 3.5 MHz. Thus for all practical purposes a signal band limited to < 3.5 MHz (rise time < 100 ns) will not experience any filter induced dispersion effects after compensation.

Figure 8 shows the amplitude flatness of the 4 MHz analog alias filter after the data has been compensated. The compensated filter is within 0.1 dB of unity for signals up to 3.5 MHz.

Each E1430A digital alias filter bandwidth has a different delay flatness characteristic and therefore requires a different compensation filter to generate the desired flat, phase-linear response. Appendix 2 gives tap values for a FIR based compensation filter for each filter bandwidth. For the filters at 125 kHz, and below, amplitude flatness and phase linearity are dominated by the digital filters so that their responses are frequency scaled versions of each other. In this case a single set of tap values given in the last column of the table in appendix 2 can be used. Figure 9 and figure 10 show the amplitude and phase response for these narrow band filters after applying the compensation. Note that in this case the frequency scale is given in terms of the digital filter output sample rate rather than in absolute frequency. Graphs for the compensated intermediate bandwidths are not given in this paper; however, the amplitude and phase accuracy of these filters falls between the 4 MHz filter and the narrow band filters.

Figure 8
Amplitude flatness of compensated 4 MHz alias filter. Vertical scale is 0.1 dB/div. Horizontal scale is 0.5 MHz/div

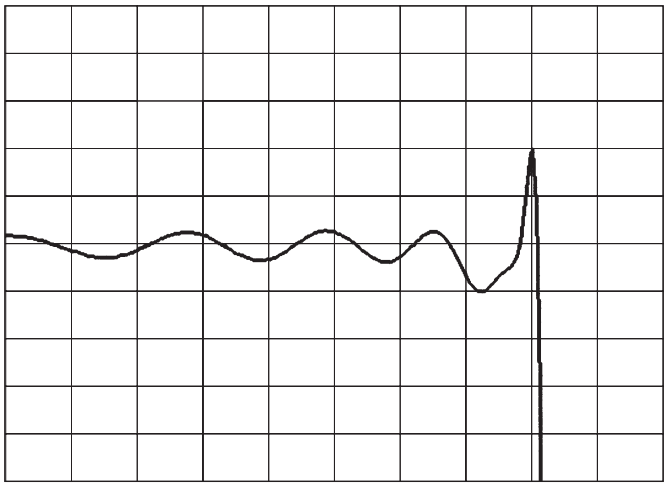


Figure 9
Amplitude flatness of compensated digital alias filter of bandwidth <<4 MHz. Vertical scale is 0.1 dB/div. Horizontal scale is .05 of digital filter output sample rate

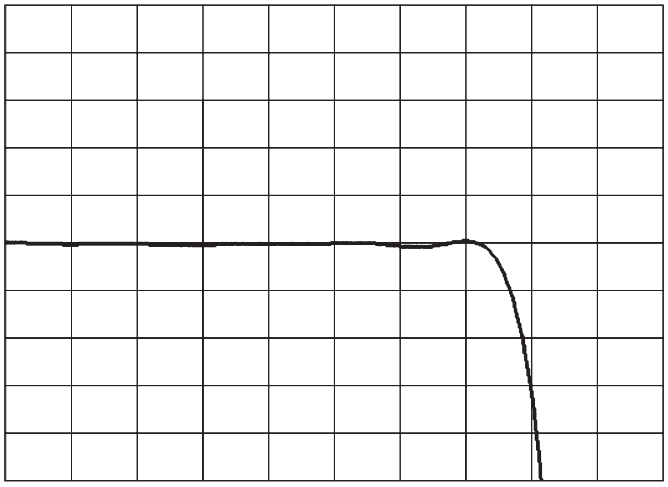
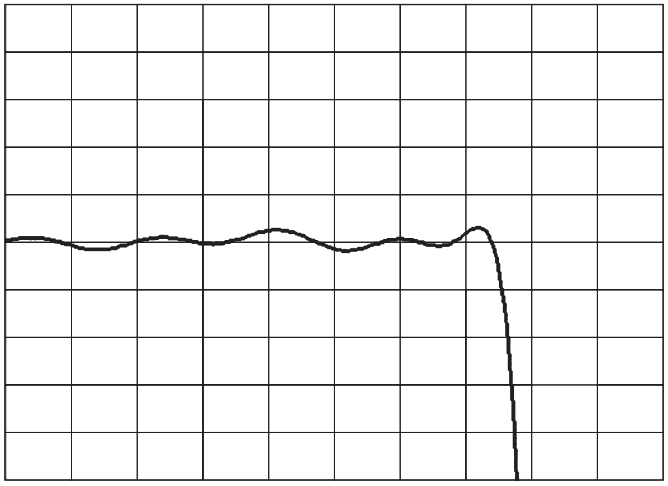


Figure 10
Deviation from linear phase for compensated digital alias filter of bandwidth <<4 MHz. Vertical scale is 0.1°/div. Horizontal scale is .05 of digital filter output sample rate



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Precision Pulse Capture with Filtering

So far our discussion has focused on Nyquist sampling of the signal of interest. In applications where event to event elapsed time is all that is of interest it is unnecessary to have a complete representation of the signal. An example is determining pulse propagation time for some radars. This requires only enough data to determine round trip propagation time of the pulses, all other data on the pulses can be filtered out without adversely effecting the answer. In such applications filtering can be used to decrease noise, and thus increase sample precision, even though the bandwidth of the filter is less than the Nyquist rate of the signal of interest.

Figure 11 shows an example of how the shape of the pulse can be changed by a filter with a bandwidth less than the Nyquist rate. This effect can not be corrected out because the filter removes too much signal information. Notice, however, the filtering does not effect pulse to pulse timing.

The response of the E1430A 4 MHz bandwidth filter to an ideal step voltage is shown in figure 12. The overshoot and ringing is primarily due to the sharp cutoff of the filter in the frequency domain. The asymmetry of the response is due to the nonlinear phase characteristics.

These responses must be compensated for to keep them from interfering with the timing data.

Figure 11
Sketch of pulse timing capture with filtering and resample

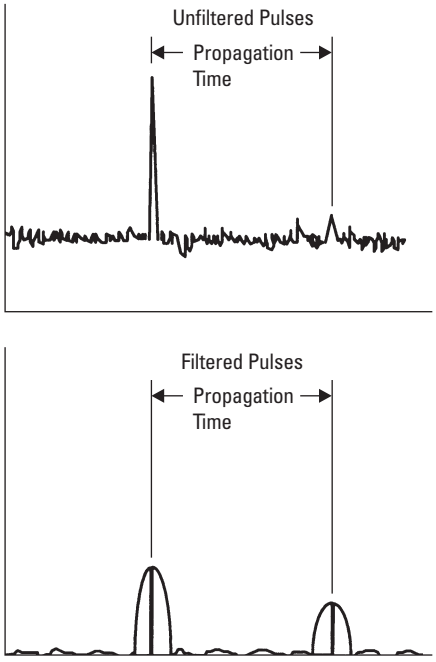
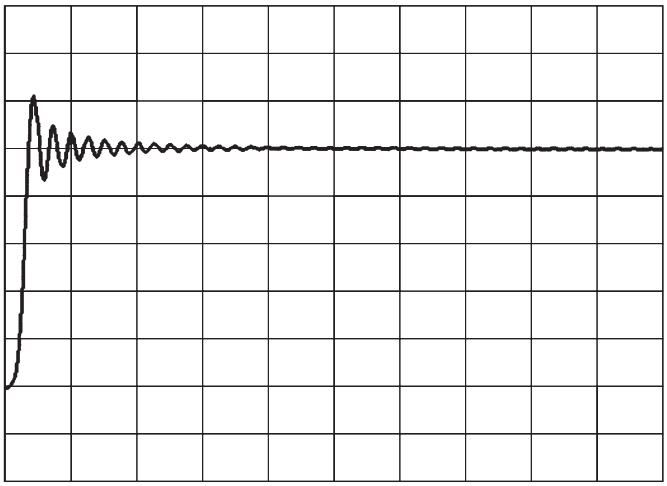


Figure 12
Step response of 4 MHz analog alias filter without compensation. Horizontal scale is 1 µs/div



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Figure 13 shows the step response of the 4 MHz alias filter after applying a 12-tap finite impulse response (FIR) digital compensation filter to the output data. This digital filter was designed so that when cascaded with the analog alias filter, the resulting response is a 1 MHz bandwidth gaussian filter. This reduction in bandwidth does not affect the location in time of the peak of the signal which is all we are trying to extract in the scenario. Notice that, even though the bandwidth has been reduced, the step settling time is improved.

To give a better appreciation for the settling time improvement, figure 14 and figure 15 show the deviation of the absolute value of step response from its final value in decibels (dB = 20 log (V_2/V_1)). Measuring from the 50% (-6 dB) point the compensator reduced the 0.1% (-60 dB) settling time from 6 µs to less than 0.8 µs.

Figure 13
Step response of 4 MHz analog alias filter with 12-tap compensation giving a Gaussian response with 1 MHz 3 dB bandwidth. Horizontal scale is 1 µs/div

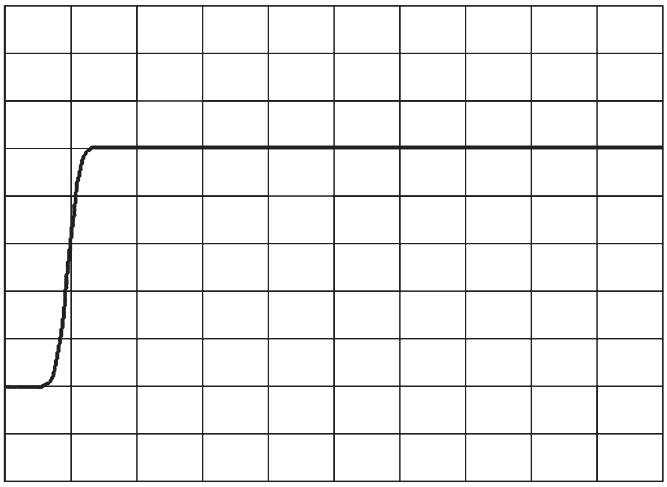


Figure 14
Log plot of step response settling for uncompensated 4 MHz analog alias filter. Vertical scale is 10 dB/div starting with 0 dB at the top of the graph. Horizontal scale is 1 µs/div

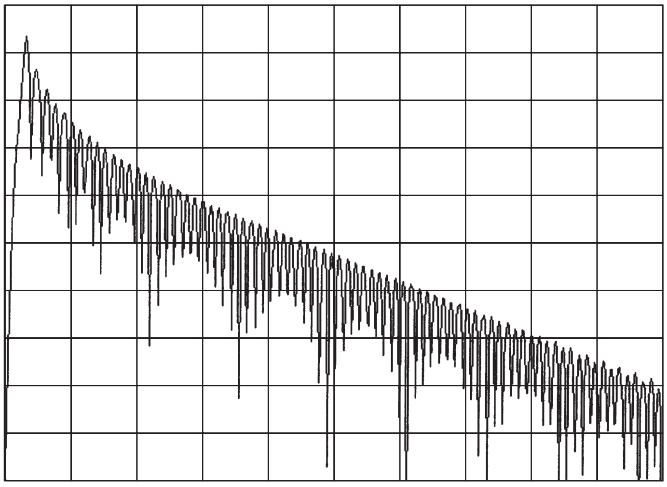
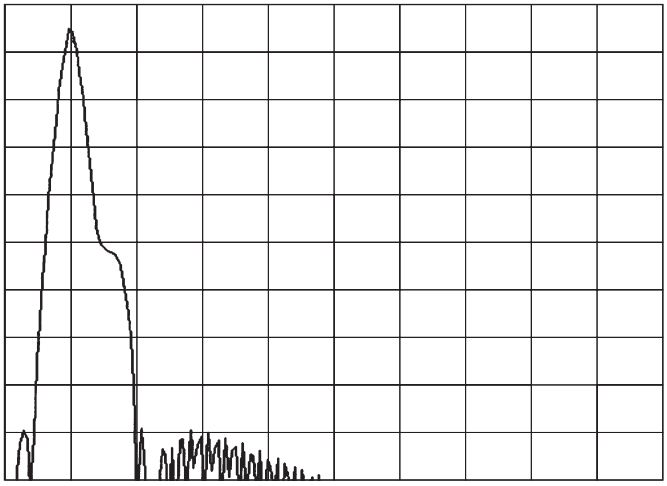


Figure 15
Log plot of step response settling for 1 MHz bandwidth compensated alias filter. Vertical scale is 10 dB/div starting with 0 dB at the top of the graph. Horizontal scale is 1 µs/div



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Figure 16 and figure 17 show the impulse response of the uncompensated 4 MHz alias filter and the compensated filter respectively. The asymmetry of the uncompensated impulse response is a result of the phase non-linearity of the frequency response (non-uniform group delay). The compensated impulse response appears symmetric because the phase non-linearities have been removed by the compensation filter.

The digital decimation filters used to reduce the bandwidth and sample rate in the E1430A provide essentially the same function as the analog alias filter. Like the analog filter, they are sharp cutoff filters with non-uniform group delay. Thus they produce ringing step responses and non-symmetric impulse responses.

Figure 16
Normalized impulse response of 4 MHz uncompensated alias filter. Horizontal scale is 1 µs/div

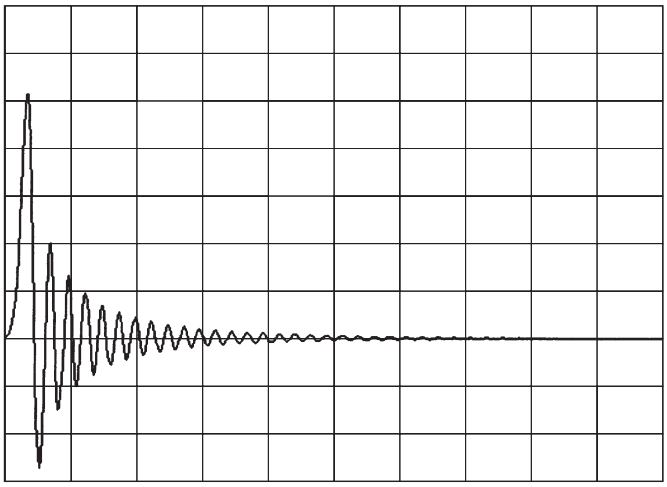
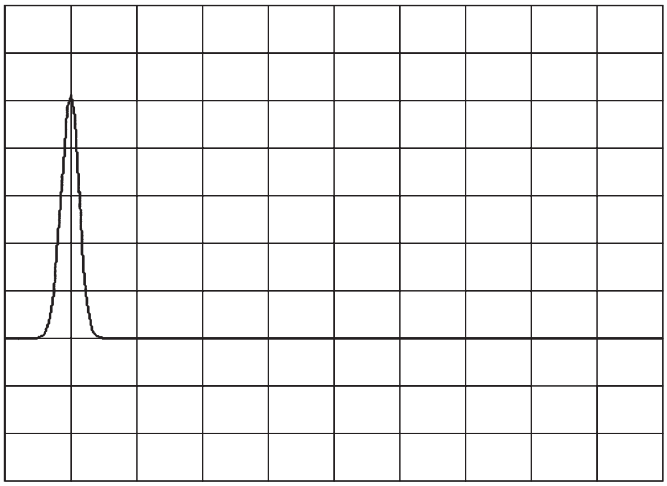


Figure 17
Normalized impulse response of 1 MHz Gaussian compensated alias filter. Horizontal scale is 1 µs/div



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Figure 18 shows a log plot of the step response error versus output sample periods for the digital filters. Note that 0.1% settling time is approximately 16 output sample periods in those digitally filtered bandwidths which are significantly narrower than the 4 MHz analog filter bandwidth. These filters can be compensated to produce 0.1% step settling within 4 output samples and a symmetric impulse response. The resulting log step settling response is shown in figure 19. This result shown on a linear vertical scale looks virtually identical it in shape to figure 15.

The tap values for the 12-tap compensation filter in this situation are different from those used to compensate the analog filter. Appendix 3 gives the tap values to be used to compensate for various E1430A bandwidths to produce a Gaussian bandwidth of Fs/10. These compensation filters are also included in the library of functions (written in C) supplied with each E1430A.

Figure 18
Log plot of step response settling for uncompensated narrow bandwidth alias filters dominated by digital decimation filters. Vertical scale is 10 dB/div starting with 0 dB at the top of the graph. Horizontal scale is 4 output sample periods/div. Digital filter bandwidth is 0.4 of the output sample frequency

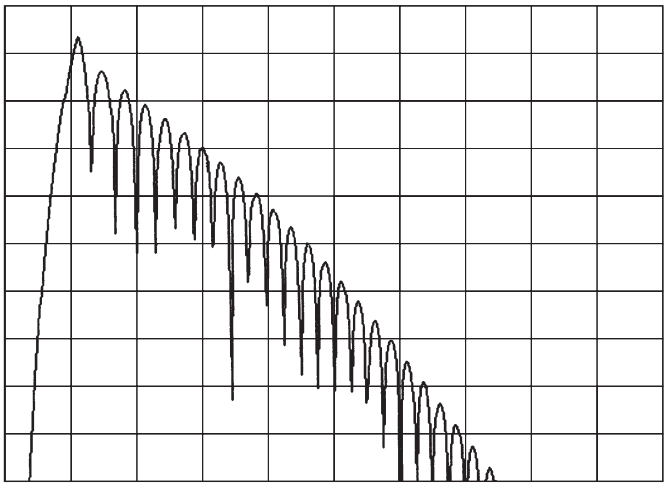
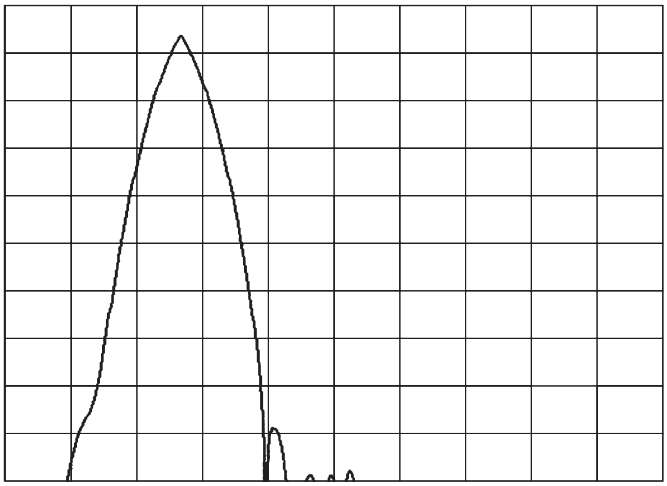


Figure 19
Log plot of step response settling for compensated narrow bandwidth digital alias filters. Vertical scale is 10 dB/div starting with 0 dB at the top of the graph. Horizontal scale is 4 output sample periods/div



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Precision Single-Sample Amplitude Capture

In some applications the information of interest is contained in a single sample of the signal. Normally, for this kind of application the samples must be synchronized with some periodic process related to the generation of the signal. For example, when measuring the output of a charge coupled imaging array it is sufficient to clock the digitizer once for each image pixel shifted out of the array. The transients associated with the clocking signal are to be ignored in this measurement, and the voltages corresponding to the pixels are to be measured after the transients have died out. Figure 20 shows a representative signal for this type of application. The sample points indicated by dots on the graph show the time points at which the digitizer is clocked to measure the input signal. In this case the transient portion of the signal between samples is to be excluded from the measurement. Any form of filtering which smears these transients into the sample points is undesirable.

One way to synchronize the E1430A with the signal source is shown in figure 21. In this case the system clock is produced by a clock generator which directly drives the device under test. A delayed version of this clock is connected to the external clock input of the E1430A. The optimum delay for the ADC clock is dependent on the particular measurement being made. The E1430A does not include a programmable delay generator, so an external one must be provided. An alternative approach (not shown) is to replace the clock generator and variable delay with a dual output clock source which has built-in variable phase between the two output channels.

Figure 20
Waveform of a typical signal where alias filtering is unwanted

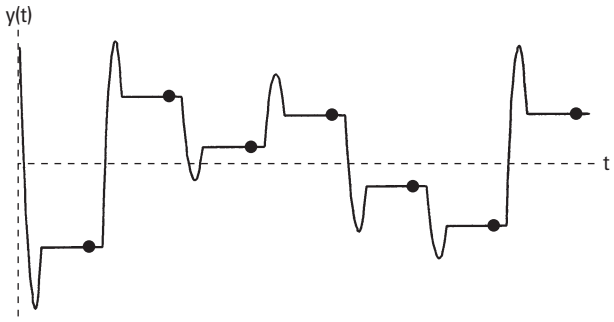
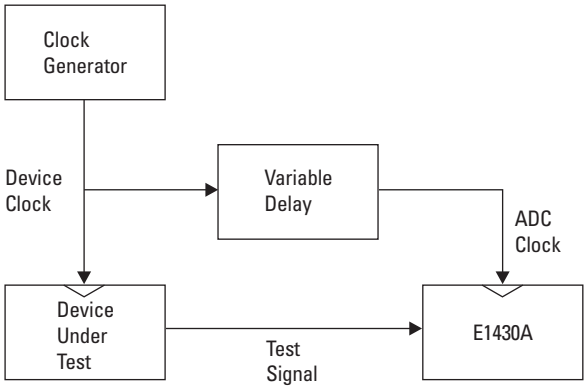


Figure 21
Connection diagram for a test system to achieve synchronous sampling



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Figure 22 shows a combined plot of impulse response and the step response of a E1430A module without any analog alias filter or digital alias filters applied. These responses result from the 20 MHz (predominantly single pole) bandwidth of the input circuitry and sample and hold circuitry in the E1430A. Figure 23 shows the step response settling on a logarithmic vertical scale to better quantify how the signal approaches it's final value. The lack of cusps in this graph (compared to figure 18) indicates that there are no zero crossings in this step settling graph. Thus, the step response is non-overshooting. From figure 23 we can see that 0.1% accurate measurements can be made within about 50 ns after the last full scale transient.

The intentional inclusion of a 20 MHz non-overshooting filter in the E1430A produces roughly the equivalent of an 8 ns sampler aperture. This bandwidth could have been made wider at the expense of an increase in the sampler noise level.

Since the input bandwidth is wide compared to the ADC sample rate, the signal is not sufficiently (Nyquist) sampled to fully characterize the filtering effects of the input bandwidth. Thus, it is not possible to "back out" the effects of the finite bandwidth by digitally filtering the sampled data. For this reason no filter coefficients are included in this section involving non-alias protected sampling. The E1430A settling graphs above are given for the purpose of establishing the appropriate ADC timing relative to the signal transients.

Figure 22 Normalized E1430A step/impulse response without alias filtering. Horizontal scale is 10 ns/div

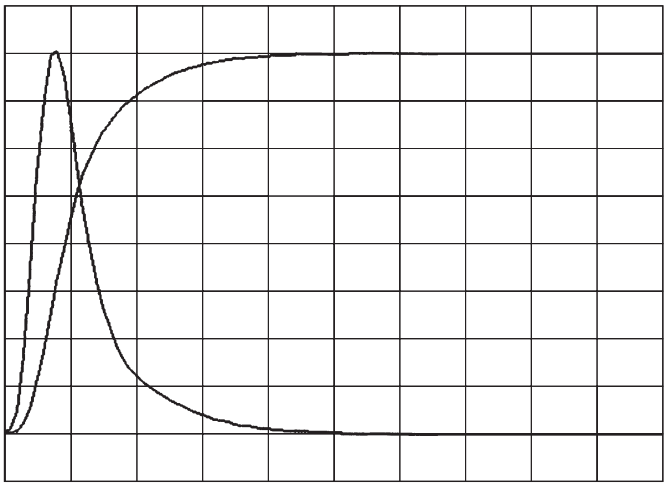
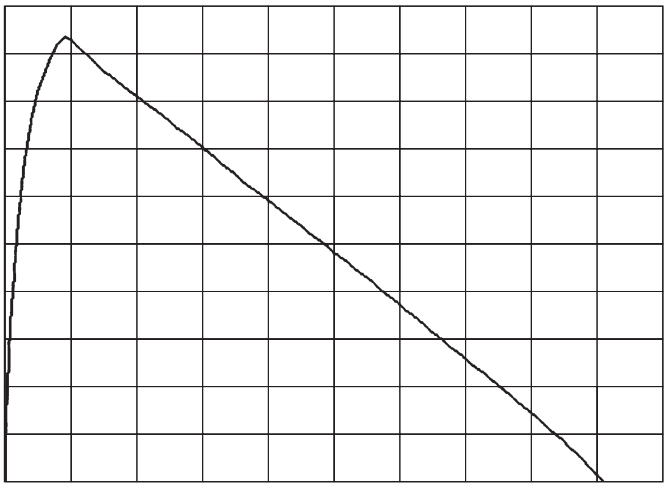


Figure 23 Log plot of step response settling for E1430A without alias filtering. Vertical scale is 10 dB/div starting with 0 dB at the top of the graph. Horizontal scale is 10 ns/div



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Appendix 1

Listing of the effective sample rate and equivalent bits for each filter in the E1430A. The data in table 1 was calculated assuming the 4 MHz analog alias filter was enabled. The data in table 2 was calculated assuming the 4 MHz analog alias filter was enabled and one stage of decimation disabled providing approximately 2.5 times Nyquist over-sampling (2 x 2.56 = 5.12). The data in table 3 was calculated assuming all filtering, including the 4 MHz analog alias filter, was disabled.

Table 1
Resolution and sample rate for E1430A filters with 4 MHz alias filter enabled.

Sample Rate (Sa/sec)	Bandwidth (Hz)	Equivalent Bits
1.00E+07	4.00E+06	11.3
5.00E+06	2.00E+06	11.8
2.50E+06	1.00E+06	12.3
1.25E+06	5.00E+05	12.8
6.25E+05	2.50E+05	13.3
3.13E+05	1.25E+05	13.8
1.56E+05	6.25E+04	14.3
7.81E+04	3.13E+04	14.8
3.91E+04	1.56E+04	15.3
1.95E+04	7.81E+03	15.8
9.77E+03	3.91E+03	16.3
4.88E+03	1.95E+03	16.8
2.44E+03	9.77E+02	17.3
1.22E+03	4.88E+02	17.8
6.10E+02	2.44E+02	18
3.05E+02	1.22E+02	18
1.53E+02	6.10E+01	18
7.63E+01	3.05E+01	18
3.81E+01	1.53E+01	18
1.91E+01	7.63E+00	18
9.54E+00	3.81E+00	18
4.77E+00	1.91E+00	18
2.38E+00	9.54E-01	18
1.19E+00	4.77E-01	18
5.96E-01	2.38E-01	18

Table 2
Resolution and sample rate for E1430A filters with the first decimation stage disabled. Effective sample rate is approximately 2.5x Nyquist.

Sample Rate (Sa/sec)	Bandwidth (Hz)	Equivalent Bits
1.00E+07	4.00E+06	11.3
1.00E+07	2.00E+06	11.8
5.00E+06	1.00E+06	12.3
2.50E+06	5.00E+05	12.8
1.25E+06	2.50E+05	13.3
6.25E+05	1.25E+05	13.8
3.13E+05	6.25E+04	14.3
1.56E+05	3.13E+04	14.8
7.81E+04	1.56E+04	15.3
3.91E+04	7.81E+03	15.8
1.95E+04	3.91E+03	16.3
9.77E+03	1.95E+03	16.8
4.88E+03	9.77E+02	17.3
2.44E+03	4.88E+02	17.8
1.22E+03	2.44E+02	18
6.10E+02	1.22E+02	18
3.05E+02	6.10E+01	18
1.53E+02	3.05E+01	18
7.63E+01	1.53E+01	18
3.81E+01	7.63E+00	18
1.91E+01	3.81E+00	18
9.54E+00	1.91E+00	18
4.77E+00	9.54E-01	18
2.38E+00	4.77E-01	18
1.19E+00	2.38E-01	18

Table 3
Resolution, and sample rate for E1430A with all filters, including 4 MHz alias filter, disabled.

Sample Rate (Sa/sec)	Bandwidth (Hz)	Equivalent Bits
1.0 E+07	2.0 E+07	10.7

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Appendix 2

FIR compensation filter taps to provide flat amplitude and delay response. For use when Nyquist sampling signals of interest. These compensation filters are provided in the C function library supplied with E1430A.

Bandwidth	4 MHz	2 MHz	1 MHz	500 kHz	250 kHz	≤ 125 kHz
Risetime	104 ns	208 ns	416 ns	832 ns	1.664 us	≥ 3.328 us
Tap 1	-0.0090632431	0.0061045697	-0.0005346914	0.0021739867	0.0021789701	0.0021762991
Tap 2	0.0403442196	-0.0266320243	0.0031899860	-0.0095495687	-0.0095636034	-0.0095510148
Tap 3	-0.1111247127	0.0780778914	-0.0111934937	0.0280763848	0.0281013384	0.0280643245
Tap 4	0.2295616464	-0.1835129300	0.0302849360	-0.0668469247	-0.0668723025	-0.0667865630
Tap 5	-0.3725473943	0.3598014077	-0.0693722530	0.1376239860	0.1376092705	0.1374409023
Tap 6	0.4584645073	-0.5583443658	0.1401288061	-0.2504943935	-0.2503416700	-0.2500554394
Tap 7	-0.3422412711	0.4784234846	-0.2527946476	0.3969269104	0.3964569140	0.3960443095
Tap 8	-0.0851389192	0.5957542630	0.3993771823	-0.4971927208	-0.4962045460	-0.4957503813
Tap 9	0.4943455069	0.2662938401	-0.5009024041	0.2788037179	0.2775298766	0.2772975152
Tap 10	0.5574054882	-0.0185992504	0.2844033769	0.6775941121	0.6772307703	0.6770235293
Tap 11	0.1287296124	0.0074204855	0.6742524217	0.3008325655	0.3013844431	0.3015128392
Tap 12	0.0112645598	-0.0047873715	0.3031607806	0.0020519443	0.0024905390	0.0025836795

Appendix 3

FIR compensation filter taps to provide fs/10 bandwidth gaussian response. For use when sampling signals with rise times faster that the filter. These filters are provided in the C function library provided with the E1430A.

Uncomp BW	4 MHz	2 MHz	1 MHz	500 kHz	250 kHz	≤ 125 kHz
Comp. BW	1 MHz	500k Hz	250 kHz	125 kHz	62.5 kHz	BW/4
RiseTime	34 ns	68 ns	136 ns	272 ns	544 ns	0.34/ (BW/4)
Tap 1	-0.0001438232	-0.0000509396	0.0000578733	0.0000671559	0.0000683615	0.0000686182
Tap 2	0.0004375388	0.0000816445	-0.0000960332	-0.0001113102	-0.0001132112	-0.0001135697
Tap 3	-0.0010889277	-0.0001024121	0.0003962592	0.0004661724	0.0004752008	0.0004774112
Tap 4	0.0031339988	0.0013022384	0.0045435212	0.0051858924	0.0052687548	0.0052918535
Tap 5	0.0090519210	0.0163253382	0.0473208419	0.0515947779	0.0521221964	0.0522574368
Tap 6	0.1054546194	0.1026347327	0.1750266951	0.1823180681	0.1831375217	0.1833301560
Tap 7	0.2426502308	0.2456293022	0.2922081784	0.2946115118	0.2947923085	0.2948214846
Tap 8	0.2979060678	0.2993970462	0.2690481450	0.2642062054	0.2636048290	0.2634562847
Tap 9	0.2093058505	0.2129508964	0.1486956912	0.1426898174	0.1420424697	0.1418929148
Tap 10	0.0988023110	0.0932374340	0.0508912601	0 0478918083	0.0475967902	0.0475301247
Tap 11	0.0277880544	0.0250132514	-0.0105864549	0.0098328276	0.0097645006	0.0097488968
Tap 12	0.0067021574	0.0035814678	0.0013211130	0.0012470729	0.0012402780	0.0012383883

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