ANALOG DEVICES

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RMS-to-DC Converters Ease Measurement Tasks

RMS-to-DC converters compute the true rms value of a signal without regard to waveform. You can use them as building blocks in a variety of measurement circuits.

by Bob Clarke, Mark Fazio, and Dave Scott

INTRODUCTION

RMS-to-DC converters provide a dc output equal to the rms value of an ac or fluctuating dc input. Analog Devices provides a selection of five such converters: the AD536A, AD636, AD637, AD736, and AD737. The theory and applications of the AD536A, AD636, and AD637 are covered in great detail in the *RMS-to-DC Conversion Application Guide, Second Edition.*

The newer AD736 and AD737 rms-to-dc converters, however, are not covered in the guide. This application note supplements the guide and discusses the theory and applications of the AD736 and AD737. It also contains information on how to enhance the accuracy and reduce the settling time of the AD637.

There are five main sections: How RMS-to-DC Converters Work, How to Select an RMS-to-DC Converter, Theory of the AD736 and AD737, Applications of the AD736 and AD737, and Applications of the AD637. Other sources of information on rms-to-dc converters include the *RMS-to-DC Conversion Application Guide, Second Edition*, the *Nonlinear Circuits Handbook, Second Edition,* and the data sheets for the AD536A, AD636, AD637, AD736, and AD737; all are available from Analog Devices.

WHY USE AN RMS-TO-DC CONVERTER?

Early multimeters used a simple rectifier and averaging circuit for ac measurements. These meters were then calibrated to read the rms value, but this was correct only for one waveform, invariably, a sine wave. In contrast to averaging circuits, true rms-to-dc converters measure the rms value of an input signal without regard to waveform.

Waveforms differ in their crest factor, which is defined as the ratio of the peak signal amplitude to the rms amplitude, that is, Crest Factor = V_{PEAK}/V_{RMS}. Many common waveforms, such as sine and triangle waves, have relatively low (≤2) crest factors. Such other waveforms as low-duty-cycle pulse trains and SCR waveforms have high crest factors.

To obtain accurate results using an averaging circuit, the user would have to know the waveform in advance and apply a correction factor. RMS-to-DC converters provide an accurate answer for a variety of crest factors. The AD637 handles crest factors as large as 10 with no more than 1% additional error; the AD736 and AD737 handle crest factors as large as 5. Table I contrasts true rms values and the measurement errors introduced by average-responding circuits for various waveforms.

HOW RMS-TO-DC CONVERTERS WORK

The rms-to-dc converters discussed here solve an implicit equation for the rms value of a voltage. The following discussion will show the transformations that lead from the definition of rms voltage to the implicit equation. It will then explain the implementation of the implicit equation in a monolithic rms-to-dc converter.

The definition of the rms value of a voltage is

$$V_{RMS} = \sqrt{\frac{1}{T}\int_0^T [V(t)^2]\,dt} \qquad (1)$$

17

where V_{RMS} is the rms value, T is the duration of the measurement, and V(t) is the instantaneous voltage, a function of time, but not necessarily periodic.

Squaring both sides of this equation yields

$$V_{RMS}^2 = \frac{1}{T}\int_0^T [V(t)^2]\,dt\,. \qquad (2)$$

The integral can be approximated as a running average:

$$Avg\,[V(t)]^2 = \frac{1}{T}\int_0^T [V(t)^2]\,dt \qquad (3)$$

then Equation 2 simplifies to

$$V_{RMS}^2 = Avg\,[V(t)^2]. \qquad (4)$$

Table I. Error Introduced by an Average Responding Circuit When Measuring Common Waveforms

Waveform Type 1 Volt Peak Amplitude	Crest Factor (V_{PEAK}/V_{RMS})	True RMS Value	Average Responding Circuit Calibrated to Read RMS Value of Sine Waves Will Read	% of Reading Error* Average Responding Circuit
Undistorted Sine Wave	1.414 ($\sqrt{2}$)	0.707 V	0.707 V	0%
Amplitude-Symmetrical Square Wave	1.00 (Exact)	1.00 V	1.11 V	+11.0%
Triangle Wave	1.732 ($\sqrt{3}$)	0.577 V	0.555 V	−4%
Gaussian Noise (98% of Peaks <1 V)	3	0.333	0.266	−20.2%
Examples of Unipolar Pulse Trains	2 10	0.5 V 0.1 V	0.25 V 0.01 V	−50% −90%
SCR Waveforms 50% Duty Cycle 25% Duty Cycle	2 4.7	0.495 V 0.212 V	0.354 V 0.150 V	−28% −30%

$$\% \text{ of Reading Error} = \frac{\text{Average Responding Value} - \text{True RMS Value}}{\text{True RMS Value}} \times 100\%$$

Dividing both sides by V_{RMS} yields

$$V_{RMS} = \frac{Avg\,[V(t)^2]}{V_{RMS}}. \tag{5}$$

This expression provides the basis for the implicit solution for V_{RMS}, and is the technique used in Analog Devices' line of monolithic rms-to-dc converters.

Note that taking the square root of both sides of Equation 4 yields

$$V_{RMS} = \sqrt{Avg\,[V(t)^2]} \tag{6}$$

which is an alternate way of expressing the rms (root of mean of square) value of the function.

The implicit method of rms computation is preferable to the explicit method (successively squaring, averaging, and taking the square root of the input signal) for practical reasons that result in a superior dynamic range. Using the explicit method, the output of the squarer will vary over a 10,000:1 dynamic range (1 mV to 10 V) for a 100:1 (0.1 V to 10 V) instantaneous input. Since the input squarer used in the explicit method will have errors greater than 1 mV, the error will strongly depend on signal level, resulting in an overall dynamic range of less than 100:1.

Figure 1 shows the implicit method of rms-to-dc conversion. The circuit is essentially an analog computer that solves Equation 5. The Analog Devices AD536A, AD636, AD637, AD736, and AD737 all use variations on this theme.

The input stage is a unity-gain buffer, which is uncommitted in the AD536A, AD636, and AD637 and committed in the AD736 and AD737. "Uncommitted" in this context means that both inputs and the output connection are accessible; the user has the option of using this buffer as a high impedance input for the converter, using it to build an active filter to follow the rms-to-dc converter's own averaging filter, or simply leaving it unconnected.

An absolute-value circuit (that is, a precision full-wave rectifier) follows the input buffer. The output of the absolute-value circuit drives a squarer/divider. The squarer/divider squares the input signal and divides it by the output signal, which is the averaged output of the squaring circuit. By closing the loop around the divider, Equation 5 is solved continuously.

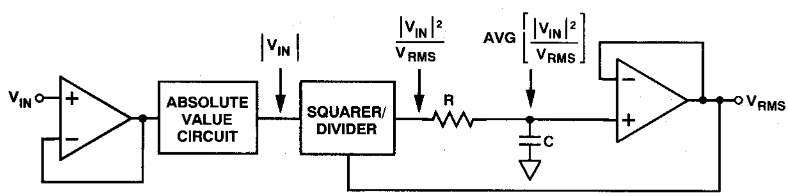


Figure 1. Implicit Method of RMS-to-DC Conversion Used in AD536A, AD636, AD637, AD736, and AD737

HOW TO SELECT AN RMS-TO-DC CONVERTER

Selecting an rms-to-dc converter means picking the product whose attributes best match the requirements of the application. Unfortunately, no one converter fits every situation, so trade-offs must be made between accuracy, bandwidth, power consumption, input signal level, crest factor, and settling time.

The AD637 accepts input voltages as high as 7 V rms and is Analog Devices' most accurate *and* widest-bandwidth rms-to-dc converter. Its −3 dB bandwidth is 8 MHz for a 1 V rms input. It has an auxiliary dB output that is proportional to the logarithm of the input signal over a 60 dB range and a power-down feature that reduces its quiescent current from 3 mA to 450 µA.

Both the AD736 and AD737 are optimized for use in portable instruments; they consume less than 200 µA of quiescent current and accept signal levels from 0 to 200 mV rms; as we'll see later, external attenuators can be added for other signal ranges. The AD737 also has a power-down input, which allows the user to reduce its quiescent current from 160 µA to 40 µA in portable applications. Table II summarizes these specifications for the AD637, AD736, and AD737. The AD637 is the best all-around performer, with its superior combination of accuracy, dynamic range, crest factor, and settling time. It also has the widest bandwidth, as shown in Table III.

The AD637 should also be chosen if the application requires high accuracy and a quick response for large, abrupt changes in signal level. The AD637's settling time is independent of signal level, while, for a given value of averaging capacitor, the settling time of the AD736 and AD737 depends on signal level, being longer for low level signals and shorter for high level signals.

Although they have less bandwidth, the AD736 and AD737 perform better than the AD637 for low level signals (<10 mV) and consume less power. (Later in this application note, we will see how to improve the AD637's performance for low level (<20 mV) signals by using an external preamplifier.) They can also be used as general purpose parts, replacing such op amp circuits as averaging converters and precision rectifiers. Both the AD737 and AD637 also have a power-down feature.

Table II. RMS-to-DC Converter Selection Guide

Model	Conversion Accuracy ±mV±% Reading	Maximum Power Consumption	Continuous Input (V_{RMS})	Crest Factor for ≤ 1% Additional Error	Relative Settling Time	Comments
AD637J	±1 mV ± 0.5%	3 mA @ ±15 V	7 @ V_S = ±15	≤10	Fast	Highest Accuracy
AD637K	±0.5 mV ± 0.2%	3 mA @ ±15 V	7 @ V_S = ±15	≤10		Highest Bandwidth Precision Applications
AD736A/J	±0.5 mV ± 0.5%	0.2 mA @ ±5 V	1 @ V_S = ±5	≤3	Slow	Low Cost
AD736B/K	±0.3 mV ± 0.3%	0.2 mA @ ±5 V	1 @ V_S = ±5	≤3		Low Power Output Buffer
AD737A/J	±0.4 mV ± 0.5%	0.16 mA @ ±5 V	1 @ V_S = ±5	≤3	Slow	Low Cost
AD737B/K	±0.2 mV ± 0.3%	0.16 mA @ ±5 V	1 @ V_S = ±5	≤3		Lowest Power No Output Buffer

17

Table III. RMS-to-DC Converter Bandwidth vs. Accuracy

Bandwidth (kHz) for 1% Additional Error for	AD637	AD736		AD737	
		Pin 1	Pin 2	Pin 1	Pin 2
V_{IN} = 1 mV	NA	1 kHz	1 kHz	1 kHz	1 kHz
V_{IN} = 10 mV	NA	6 kHz	6 kHz	6 kHz	6 kHz
V_{IN} = 20 mV	11 kHz	NA	NA	NA	NA
V_{IN} = 200 mV	66 kHz	90 kHz	33 kHz	90 kHz	33 kHz
3 dB Bandwidth (kHz) for					
V_{IN} = 1 mV	NA	5 kHz	5 kHz	5 kHz	5 kHz
V_{IN} = 10 mV	NA	55 kHz	55 kHz	55 kHz	55 kHz
V_{IN} = 20 mV	150 kHz	NA	NA	NA	NA
V_{IN} = 200 mV	1000 kHz	460 kHz	190 kHz	460 kHz	190 kHz

THEORY OF THE AD736 AND AD737

To better understand how the AD736 works, consider the simplified block diagram, first as drawn on the AD736 data sheet (Figure 2), and then redrawn with the averaging (C_{AV}) and filter (C_F) capacitors in Figure 3 to better show signal flow.

The input to the AD736 and AD737 is through a FET-input op amp connected as a unity-gain buffer. This amplifier allows both a high impedance, buffered input (Pin 2) or a low impedance input (Pin 1) that provides a wider dynamic range. The high impedance input, with its low input bias current, is well suited for use with high impedance input attenuators.

The output of the buffer drives a full-wave rectifier or absolute value circuit, which in turn drives a 2-quadrant squarer/divider. The output of the squarer/divider drives the summing node of an inverting op amp connected as a current-to-voltage converter. Pin 3 gives access to this node to connect a filter capacitor in parallel with the 8 kΩ feedback resistor to form a 1-pole low-pass filter.

The AD737 (Figures 4 and 5) is similar in design and function to the AD736 except that the AD737 lacks an output buffer, which is omitted in order to reduce power consumption, and has a power-down feature that further reduces power consumption. Its output stage is a simple open-collector NPN transistor with an 8 kΩ load resistor.

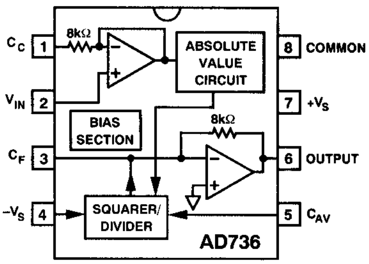


Figure 2. Simplified Block Diagram of the AD736

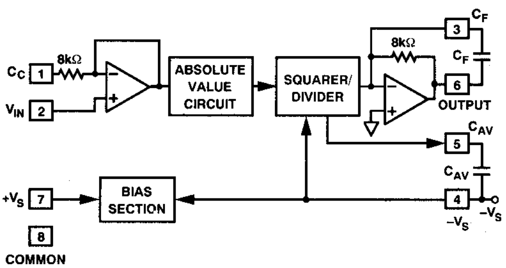


Figure 3. Redrawn Simplified Block Diagram of the AD736

It develops its output voltage by sinking current through this resistor. The external averaging capacitor (C_{AV}) for the AD736 and AD737 is connected between Pins 4 ($-V_S$) and 5 (C_{AV}), which places it across a transistor's base-emitter junction in the rms core. This means that the resistance in parallel with the averaging capacitor is that of a diode, and thus signal-level dependent. The resulting time constant is inversely proportional to the rms value.

Because the external averaging capacitor, C_{AV}, "holds" the rectified input signal during rms computation, its value directly affects the accuracy of the measurement —especially at low frequencies. (The larger the value of C_{AV}, the lower the error.) Also, because the averaging capacitor appears across a base-emitter junction in the squarer/divider whose resistance varies with signal level, the averaging time constant will increase linearly as the input signal is reduced.

Consequently, as the input level decreases, errors due to nonideal averaging will *decrease* while the time it takes for the circuit to settle to the new rms level will *increase*. Therefore, lower input levels allow the circuit to perform better (due to increased averaging) but increase the waiting time between measurements because the capacitor takes longer to discharge. Thus, a trade-off between computational accuracy and settling time is required. This topic is discussed in detail in the *RMS-to-DC Conversion Application Guide, Second Edition.*

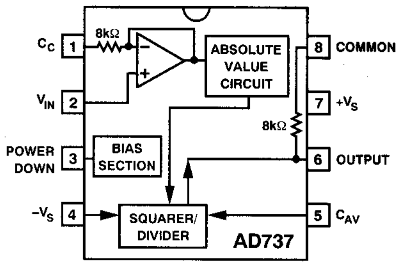


Figure 4. Simplified Block Diagram of the AD737

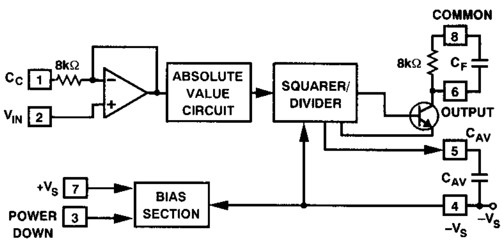


Figure 5. Redrawn Simplified Block Diagram of the AD737

DC Error, Output Ripple, and Averaging Error

Figure 6 shows the typical output waveform of the AD736 and AD737 with a sine wave input applied. The ideal output of V_{OUT} = rms (V_{IN}) is never achieved; instead, the output contains both a dc error and an ac ripple component.

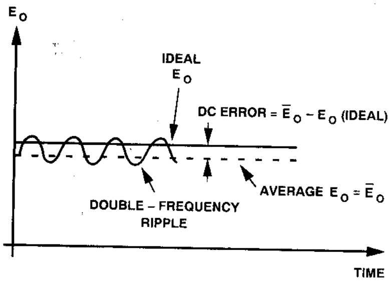


Figure 6. Output Waveform of the AD736 and AD737 for a Sine Wave Input Voltage

The dc error is the difference between the average of the output signal (when the ripple in the output has been removed by filtering) and the ideal dc output. The dc error component is therefore set solely by the value of averaging capacitor used—no amount of post filtering (i.e., using a very large C_F) will reduce this error, although the ripple may be removed by using a large value of C_F.

In most cases, the combined magnitudes of both the dc and ac error components need to be considered when selecting values of capacitors C_{AV} and C_F. This combined error, representing the maximum uncertainty of the measurement is termed the "averaging error" and is equal to the peak value of the output ripple plus the dc error.

As the input frequency increases, both the dc and ac error components decrease rapidly: if the input frequency doubles, the dc error and ripple reduce to 1/2 and 1/4 their original values, respectively, and rapidly become insignificant. Table IV provides practical values of C_{AV} and C_F for several common applications. Figure 7 shows the additional error versus crest factor of the AD736 and AD737 for various values of C_{AV}.

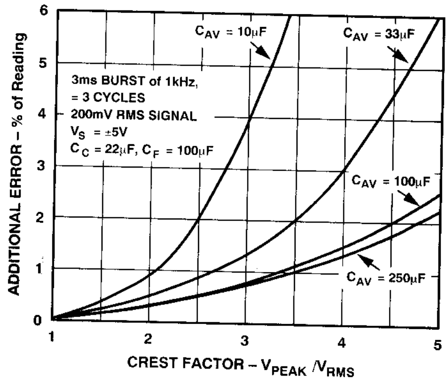


Figure 7. Additional Error vs. Crest Factor of the AD736 and AD737 for Various Values of C_{AV}

Table IV. Practical Values for C_{AV} and C_F for AD736 and AD737

Application	RMS Input Level	Low Frequency Cutoff (−3 dB)	Max Crest Factor	C_{AV}	C_F	Settling Time* to 1%
General Purpose RMS Computation	0 V–1 V	20 Hz	5	150 µF	10 µF	360 ms
		200 Hz	5	15 µF	1 µF	36 ms
	0–200 mV	20 Hz	5	33 µF	10 µF	360 ms
		200 Hz	5	3.3 µF	1 µF	36 ms
General Purpose Average Responding	0 V–1 V	20 Hz		NONE	33 µF	1.2 sec
		200 Hz		NONE	3.3 µF	120 ms
	0 mV–200 mV	20 Hz		NONE	33 µF	1.2 sec
		200 Hz		NONE	3.3 µF	120 ms
SCR Waveform Measurement	0 mV–200 mV	50 Hz	5	100 µF	33 µF	1.2 sec
		60 Hz	5	82 µF	27 µF	1.0 sec
	0 mV–100 mV	50 Hz	5	50 µF	33 µF	1.2 sec
		60 Hz	5	47 µF	27 µF	1.0 sec
Audio Applications						
Speech	0 mV–200 mV	300 Hz	3	1.5 µF	0.5 µF	18 ms
Music	0 mV–100 mV	20 Hz	10	100 µF	68 µF	2.4 sec

*Settling time is specified over the stated rms input level with the input signal increasing from zero. Settling times will be greater for decreasing amplitude input.

Calculating AD737 Settling Time

The graph of Figure 8 may be used to approximate the time required for the AD736 or AD737 to settle when its input level is reduced in amplitude. The total time required for the rms converter to settle will be the difference between two settling times extracted from the graph—the initial settling time minus the final settling time.

As an example, consider the following conditions: a 33 μF averaging capacitor, an initial rms input level of 100 mV, and a final (reduced) input level of 1 mV. From Figure 8, the initial settling time (where the 100 mV line intersects the 33 μF line) is around 80 ms. The settling time corresponding to the new or final input level of 1 mV is about 8 seconds. Therefore the net time for the circuit to settle to its new value will be dominated by the final settling time.

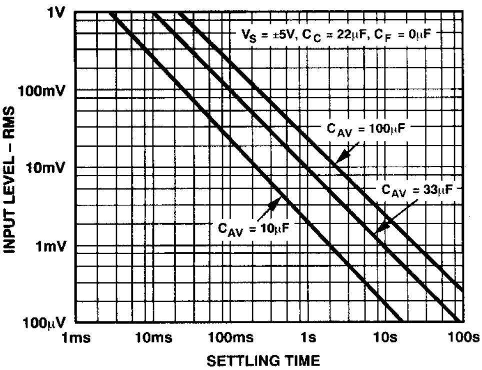


Figure 8. Settling Time vs. RMS Input Level of the AD736 and AD737 for Various Values of C_{AV}

APPLICATIONS OF THE AD736 AND AD737

AD736 as Precision Rectifier

Building a precision rectifier requires two op amps, two diodes, and a handful of matched resistors. An easy way to replace all these parts and save some board space is to use an rms-to-dc converter. Just omit the averaging capacitor and disconnect the feedback; this uses only the converter's internal precision rectifier (Figure 9), which, being monolithic, has inherently matched diodes.

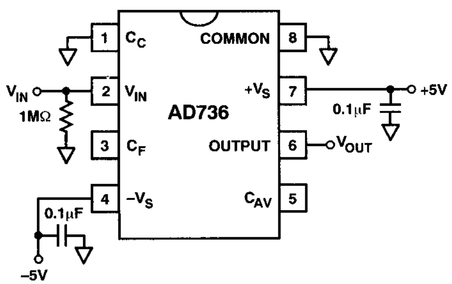
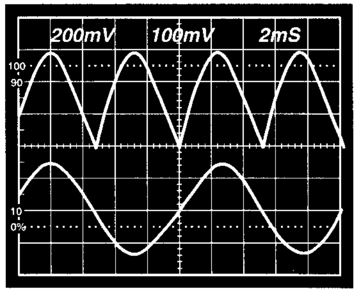


Figure 9. AD736 Connected as a Precision Rectifier

One note about precision rectifiers: as the input waveform crosses through zero, the op amp must instantaneously switch one diode on and the other off. For this reason, the bandwidth of precision rectifiers is always much less than one might otherwise expect based on



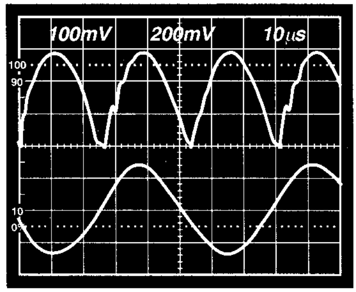


Figure 10. Performance of AD736 Precision Rectifier at 1 kHz (Top Photo) and 19 kHz (Bottom)

the gain-bandwidth product, open-loop gain, and slew rate of the op amp. The monolithic precision rectifiers used in rms-to-dc converters fare much better than discrete precision rectifiers in this regard, as the photo of the AD736's performance as a precision rectifier in Figure 10 shows.

Extending the AD736 and AD737 Full-Scale Input Ranges

The high impedance input (Pin 2) of the AD736 and AD737 allows simple resistive attenuators (Figure 11) to be used to extend their input range. Without input attenuation, both the AD736 and AD737 can accurately measure input signals as large as 200 mV rms with crest factors of 1 to 3.

The external attenuator simply reduces the full-scale input to the 200 mV rms input range of the AD736 or AD737. For a maximum 7 V rms input (10 V peak) input, for example, the attenuator should be a 35:1 (7/.2) voltage divider. The reading of the converter should be scaled by the factor of attenuation used. An external attenuator can also be used with the converter's low impedance input (Pin 1), as will be shown later in Figure 13.

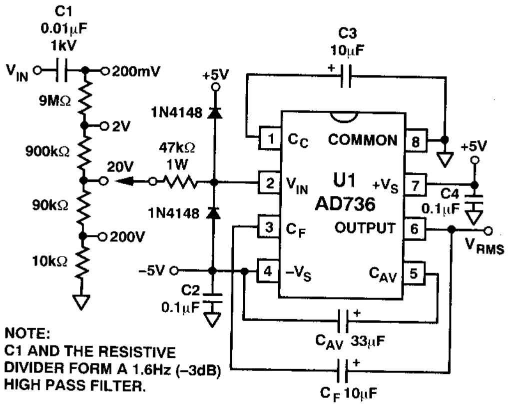


Figure 11. By using an external input attenuator, you can extend the measurement range of the AD736 and AD737. Although the AD736 is shown, this technique also works with the AD737.

Single-Supply Operation of the AD736

In dual-supply operation, the output (Pin 6) of the AD736 is at 0 V, halfway between the supply rails. But in single-supply operation, the output is at 1/2 V_{CC}. By adding a single-supply op amp as a differential amplifier, however, you can build a true "0 V out for 0 V" single-supply circuit with a ground-referenced output (Figure 12). For this circuit, $V_{RMS} = 0$ when $V_{IN} = 0$ and $V_{RMS} = 200$ mV dc when $V_{IN} = 200$ mV rms.

In this circuit, a single 9 V positive supply powers the AD736. Resistors R7 and R8 form a voltage divider across the 9 V battery that establishes a local "ground" rail at 1/2 V_{CC}, or 4.5 V. The AD736's "COMMON" pin, its 22 MΩ input bias resistor, and the inverting input of U2 (via R4 and R5) are all connected to this rail. The quiescent output voltage of the AD736, which is referenced to its COMMON pin, is 4.5 V.

A single-supply op amp, U2, is connected as a unity-gain, differential amplifier. Large-valued feedback resistors (R2 through R5) are used to minimize loading of the 4.5 V rail. U2 amplifies the difference between local ground at 4.5 V and the output of the AD736, which is also at 4.5 V for 0 V rms input. As the rms input to the AD736 increases from 0 mV to 200 mV, the AD736's output increases from 4.5 V to 4.7 V. U2's output is the difference between the AD736's output and 4.5 V, or 0 mV to 200 mV dc.

The remainder of the circuit works as follows. The AD736's input is ac coupled; R1 provides a path for the BiFET op amp's input bias current (typically 1 pA) to flow. The offset voltage due to bias current flowing through R1's 22 MΩ resistance is negligible. C3 (10 μF), connected between U1's Pins 1 and 8, provides a low frequency cutoff of 2 Hz; you can select other cutoff values using

$$f = \frac{1}{2\pi RC} \qquad (7)$$

where f is the −3 dB frequency in Hz, C is in farads, π is 3.1416, and R is fixed at 8 kΩ ±20% by the AD736.

The averaging capacitor, C_{AV}, is 33 μF and is connected between pins 4 and 5 of U1. An optional 10 μF filter capacitor, C_F, in parallel with an 8 kΩ feedback resistor across the output buffer forms a 1-pole low-pass filter with a 2 Hz cutoff frequency.

You can calculate the value of C_F using the expression

$$f = \frac{1}{2\pi RC_F} \qquad (8)$$

where f is the −3 dB frequency in Hz, C_F is in Farads, π is 3.1416, and R is fixed at 8 kΩ ±20% by the AD736. Or, since R is fixed,

$$f = \frac{20\ Hz}{C_F\ (\mu F)}. \qquad (9)$$

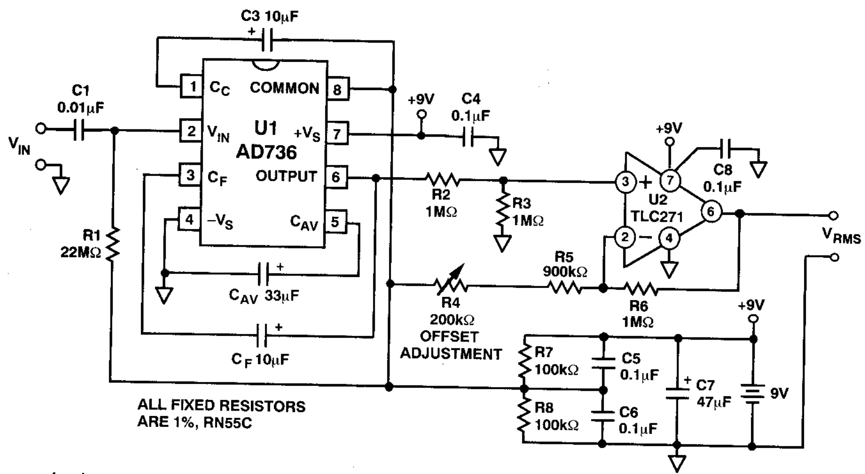


Figure 12. By using a single supply op amp to level shift the AD736's output, you can build a true single supply circuit that supplies 0 V output for 0 V input.

Single-Supply Operation of the AD737

You can also build a "true" single-supply circuit—0 V out for 0 V in—using the AD737. Note that the circuit in Figure 13 shows three design techniques: how to operate the AD737 from a single supply, how to use a resistive attenuator in series with the low impedance input of the AD737, and how to use a single-supply op amp to convert the AD737's output current to a voltage.

The combination of the input attenuator (R1 and R2) and the output op amp (U2) allows the circuit to provide 0 V dc to 2 V dc output for 0 V rms to 2 V_{RMS} full-scale input. You can also use this resistive attenuator with the AD736. The circuit consumes just 192 µA from a 9 V supply at 10 mV rms input and 240 µA from a 9 V supply at 2 V rms input.

At the input of the circuit, an additional resistance (formed here by the sum of R1 and R2) in series with Pin 1 serves as an attenuator. You can calculate the value of this resistance using the expression

$$R_{IN} = \frac{8\ k\Omega \times V_{FS}}{0.2\ V} - 8\ k\Omega \qquad (10)$$

where R_{IN} is the value of the series input resistance and V_{FS} is the desired full-scale input voltage. For the 2 V full-scale input voltage used here,

$$R_{IN} = \frac{8\ k\Omega \times 2\ V}{0.2\ V} - 8\ k\Omega \qquad (11)$$

which yields R_{IN} = 72 kΩ. For a 10 V full-scale input voltage, R_{IN} would equal 392 kΩ. Due to the thin film resistors used in the manufacturing process for the AD737 (and AD736), the tolerance of R_{IN} is 20%. Thus, the external resistance (R1 plus R2) must be 72 kΩ ±20% to compensate for the tolerance of the internal resistance.

As in the AD736 single supply circuit, a single 9 V battery powers the AD737, and two 100 kΩ resistors (R7 and R8) across the 9 V battery form a voltage divider that establishes a local "ground" rail at 1/2 V_{CC}, or 4.5 V. A single-supply op amp, U2, serves as an I/V converter that produces a 2 V full-scale output.

The output of the AD737 is an open-collector NPN transistor that normally sinks current through the 8 kΩ resistor connected through the COMMON pin to ground. In this circuit, the COMMON pin is left unconnected, and the AD737 sinks current from the node at the inverting input of U2. In order to maintain the same potential at its inverting and noninverting nodes, U2 increases its output voltage (V_{RMS}), which increases the current through feedback resistor R6. This circuit is balanced when the current through the R6 equals the current absorbed by the AD737.

The gain of the output stage can be increased by simply increasing the value of the feedback resistor, R6. The combination of C4 and R6 form the post filter in this circuit and, because R6 is an order of magnitude larger than the 8 kΩ internal resistor in the AD736 and AD737, this circuit allows a ×10 smaller value of C4 for the same amount of filtering (i.e., T = C4R6 = C_F × 8 kΩ).

A 3-Chip Digital Panel Meter for Differential Current or Voltage Measurement

By using an AD22050 difference amplifier, an AD737 rms-to-dc converter, and an ICL7136 single-chip DMM, you can build a complete, 3-chip digital panel meter that measures the rms value of an alternating current (or voltage). The circuit uses very little power: the AD22050 draws less than 300 µA quiescent current and the AD737 draws less than 160 µA quiescent current.

Figure 14 shows the circuit. The transfer function of the circuit for a 200 mV full-scale reading is

$$20 \times I_{IN} R_{SENSE} = 200\ mV \qquad (12)$$

where 20 is U1's gain, I_{IN} is the input current in Amperes, R_{SENSE} is the value of the sense resistor in Ohms,

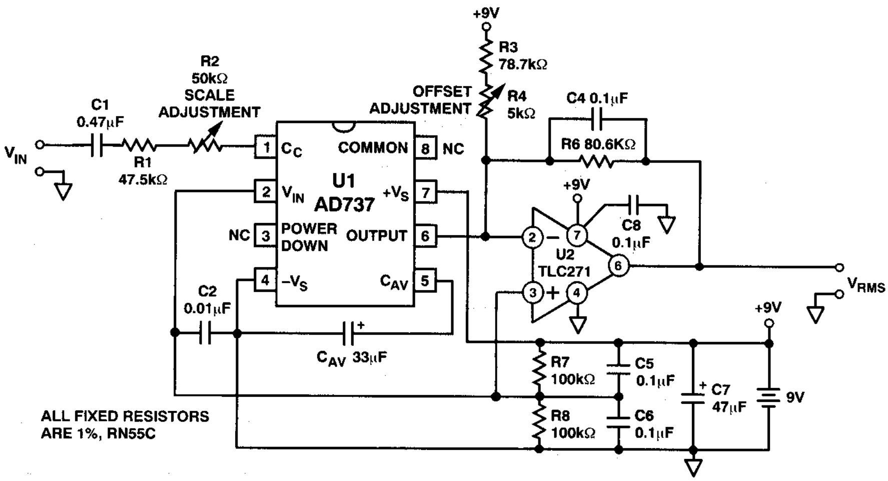


Figure 13. You can also build a true single-supply circuit using the AD737. Note that this circuit also shows how to use an input attenuator with the AD737's low impedance input.

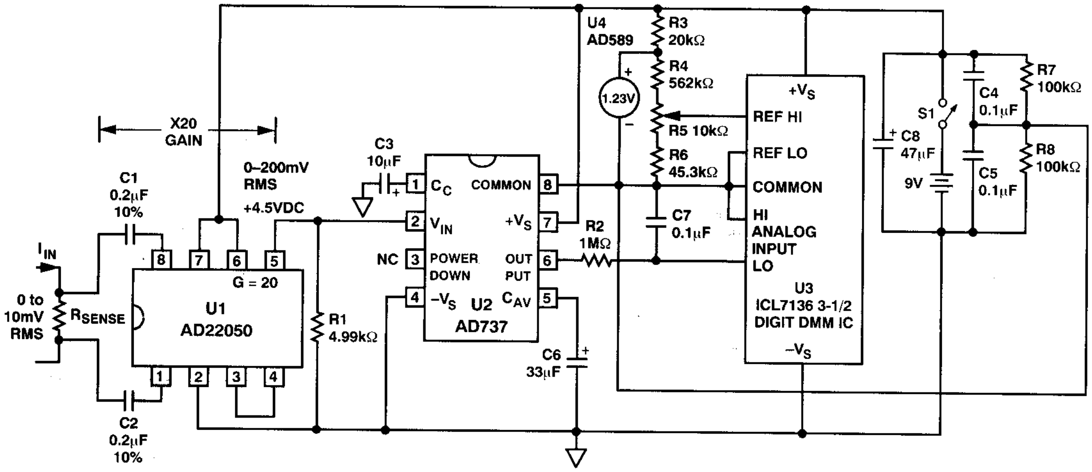


Figure 14. By using an AD737 rms-to-dc converter, an AD22050 difference amplifier, and an ICL7136 single-chip DMM, you can build a true rms low power digital panel meter.

and 200 mV is the full-scale reading of the ICL7136. A more convenient expression that provides a 100 mV full-scale reading for 10 mV across R_{SENSE} is

$$R_{SENSE} = \frac{5\ mV}{I_{IN}}. \qquad (13)$$

For example, to measure a full-scale current of 100 mA rms with a full-scale reading of 100 mV:

$$R_{SENSE} = \frac{5\ mV}{100\ mA} = 50\ m\Omega. \qquad (14)$$

Here's how the circuit works. The input current is converted to a voltage by R_{SENSE}. The input amplifier, U1, an AD22050, is a single-supply difference amplifier that amplifies the input signal by a factor of ×20; the gain of the AD22050 can be changed by adding external resistors (as shown on the next page). Its −3 dB bandwidth is 100 kHz and its slew rate is 0.1 V/μs. Note that a mismatch between C1 and C2 will degrade the CMRR of this circuit.

Connecting Pin 7 of the AD22050 to the positive supply pulls its zero-signal output at Pin 5 to 1/2 of the 9 V supply, or 4.5 V. Resistors R7 and R8 split the supplies for the AD737. The AD737 supplies a differential output between its Pins 6 and 8 to the COMMON and LO inputs of a 3-1/2-digit DMM IC, an ICL7136. (The 3-1/2-digit display and its connections are omitted for simplicity.) R2 and C7 form a simple RC filter. The small value of C7 is made possible by U3's high input impedance. U4, an AD589, provides an external 1.23 V reference to calibrate U3. To calibrate this circuit, adjust R5 to provide a 100 mV reference voltage between the REF HI and REF LO inputs of the ICL7136.

Changing the AD22050's Input Gain

The AD22050 consists of two stages: an input preamplifier and an output buffer. The gain of the AD22050 preamplifier, from the input Pins 1 and 8 to its output at Pin 3, is ×10, and that of the output buffer is ×2, making the overall gain ×20. Many applications will call for gains greater than or less than ×20. Both of these situations are readily accommodated by the addition of one external resistor. Note that this is possible because the output resistance of the input buffer (at Pin A1) is deliberately raised to 100 kΩ ±1%.

The gain may be raised by connecting a resistor from the output of the buffer amplifier (Pin 5) to its noninverting input (Pin 4) as shown in Figure 15. The gain is now multiplied by the factor R/(R−100), where R is in kΩ; for example, the gain is doubled for R = 200 kΩ. Overall gains as high as 160 are readily achievable in this way. Note that the gain becomes increasingly dependent on the accuracy of the resistor value at high gains.

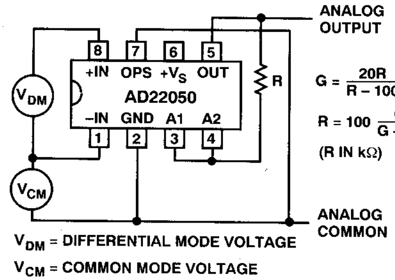


Figure 15. AD22050 Configured for Gains Greater than 20

Because the output of the AD22050 preamplifier has an output resistance of 100 kΩ (±1%), an external resistor connected from Pin 4 to ground (Figure 16) lowers the gain by a factor R/(100+R), where again R is in kΩ. When configuring the AD22050 for low gains, however, care should be taken not to exceed the output capabilities of the preamplifier, because the preamplifier with its gain of ×10 may saturate before the AD22050's output stage does.

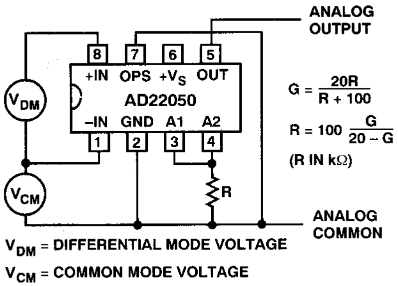


Figure 16. AD22050 Configured for Gains Less than 20

AD736 Single-Supply Circuit With Output Scaling

You can also use the AD22050 difference amplifier with the AD736 to build a 0 V in, 0 V out circuit. Figure 17 shows the circuit. The AD22050 amplifies the difference between the COMMON and V_{OUT} with a fixed gain of ×20, transforming the 0 mV to 200 mV input range to 0 V to 4 V. The output of the AD22050 can go within about 20 mV of ground, so the useful range of this circuit for 1% of reading accuracy is 10 mV to 200 mV ac rms input for 100 mV to 4 V dc output. The bandwidth of this circuit for less than 1 of reading error is 40 Hz to 6 kHz at 10 mV rms input, extending to 36 kHz at 200 mV rms input.

You can add an output low-pass filter by placing a capacitor from the junction of Pins 3 and 4 of the AD22050 to ground. The −3 dB cutoff frequency of this filter is

$$f = \frac{1}{2\pi C \times 100\ k\Omega} \tag{15}$$

where C is in farads. Alternately, f = 1.59 Hz per µF.

Transmit Current Measurements Using 4-to-20 mA Transmitter

You can also measure alternating current and transmit the results on a 4-to-20 mA current loop. Figure 18 shows the AD22050, the AD736, and the AD694, a 4-to-20 mA transmitter, configured as a 0-to-10 mV ac rms in, 4-to-20 mA out current measuring subsystem for use in process-control loops.

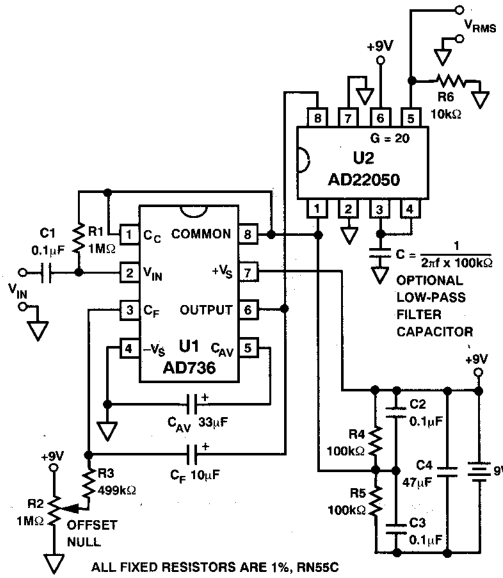


Figure 17. You can build a single supply circuit with X20 gain and optional filtering using the AD736 and AD22050.

This circuit builds on the techniques shown in the previous circuits. For example, the AD22050 provides a differential-in, single-ended out, current sensor. Here the AD22050 operates at a gain of ×20 as before, and drives the low impedance input (8 kΩ , Pin 1) of the AD736.

Because of their low power consumption, both the AD22050 and the AD736 can operate from 10 V supplied by the AD694's Pin 7 reference output. The AD694 operates from a +24 V single supply. Because this circuit

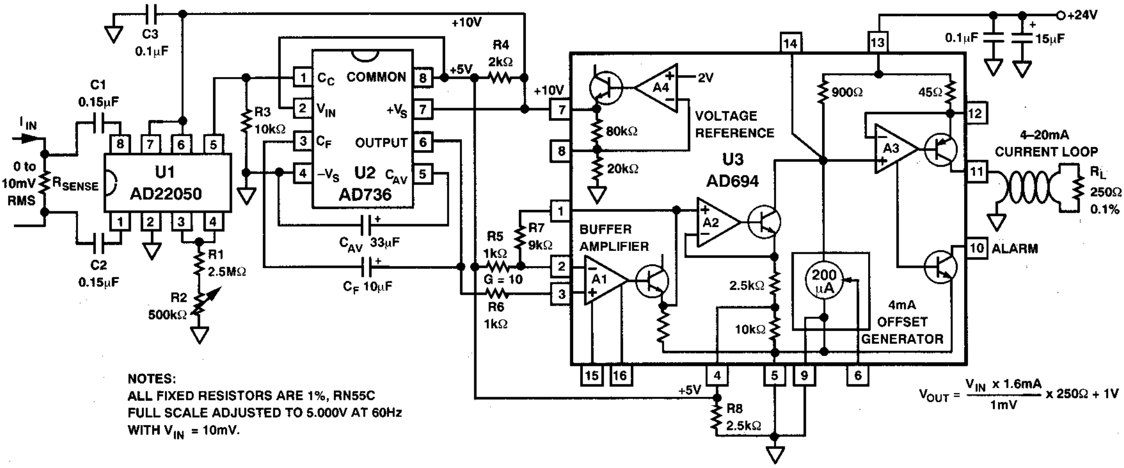


Figure 18. By using an AD694 4-to-20 mA current-loop interface IC, you can build a complete remote-monitoring system that measures true rms current or voltage with just three ICs. The entire circuit operates from a single +24 V supply.

operates from a single supply, you must bias the COMMON (Pin 8) input of the AD736 at 1/2 of the AD694's 10 V output, or 5 V. You can do this by creating a voltage divider at Pin 4 of the AD694 using R5 and R9, which is in parallel with the AD694's internal 10 kΩ resistor.

The AD694's buffer amplifier amplifies the *difference* between the AD736's output at Pin 6 and this 5 V rail—this *difference* ranges from 0 mV to 200 mV dc for a 0 mV to 10 mV rms input and produces a 4-to-20 mA current output from the AD694.

R2 serves as a gain adjustment. R5 and R7 set the gain of the AD694's amplifier A1 to ×10. R7 matches R5 to prevent offsets due to A1's input bias currents. This circuit's accuracy is 1.2% of reading from 20 Hz to 40 Hz and 1% of reading from 40 Hz to 1 kHz. Its −3 dB bandwidth is 33 kHz.

APPLICATIONS OF THE AD637

In the final section of this application note, we'll see how to enhance the accuracy of low level (<100 mV_{RMS}) measurements made with the AD637 rms-to-dc converter and reduce settling time.

Increased Accuracy for Low-Level Measurements

A problem in using all rms-to-dc converters occurs in making measurements of very low level signals. This problem arises due to slew rate limitations in the internal precision rectifier (or absolute value circuit) used to convert the bipolar input signal to a unipolar signal.

One way to circumvent this limitation is to shift the dynamic range of the input signal by amplifying it with a fixed gain stage. Figures 19's circuit uses an AD744 BiFET op amp configured as a fixed gain of ×10 amplifier. The AD744 was chosen because of its low cost, 13 MHz gain-bandwidth product (G = 2), and 75 V/μs slew rate. With a 75 V/μs slew rate, the AD744 can amplify a 10 V peak or 7 V rms signal at frequencies as high as 1.2 MHz without slew rate limiting.

The amplifier has a fixed gain of ×10. A 1 MΩ input resistor supplies the 100 pA maximum input bias current. Because the gain is ×10 and the circuit will be dc coupled to the AD637, an external offset trim is used. To minimize gain error, R4 and R5 were hand selected to set the gain of U1 at G = 1 + (R4 + R5)/R6 = 10.

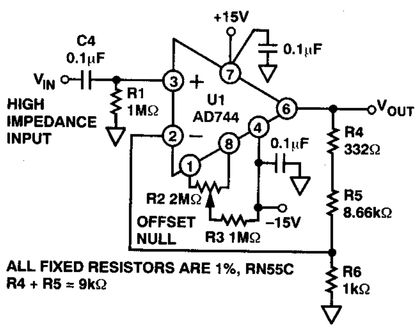


Figure 19. Adding an external preamplifier enhances the accuracy of the AD637 circuit for low level inputs. Here's an AD744 BiFET op amp configured as a gain of 10 amplifier.

A 3-Pole Ripple Filter for the AD637

The usual trade-off in designing any rms circuit is one of settling time versus accuracy and minimum output ripple. One way of reducing settling time is to use as small a value of C_{AV} as is practical while using a multipole output filter to reduce the residual ripple. Figure 20 shows how to construct a 5 Hz, 3-pole Bessel filter with constant phase using the AD637's uncommitted unity-gain buffer. Figures 21 and 22 show simulation results for the filter.

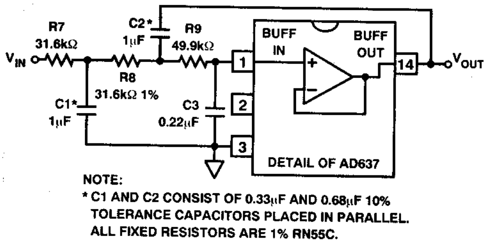


Figure 20. You can build a 3-pole Bessel filter using the AD637's on-chip buffer.

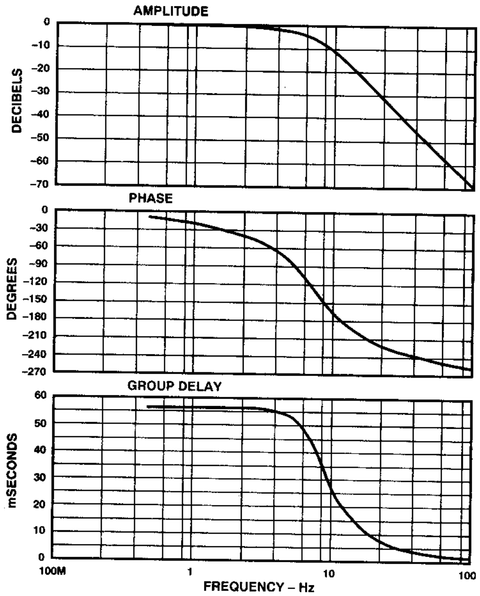


Figure 21. These simulation results show amplitude, phase, and delay for the AD637 ripple post filter.

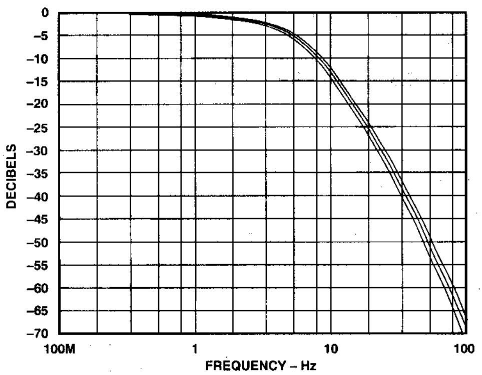


Figure 22. Results from Varying the Component Values for the AD637 Ripple Post Filter

High Accuracy AD637 Circuit for Low Level (<100 mV RMS) Measurements

Figure 23 shows how to use the preamplifier and 3-pole ripple post filter to enhance AD637 accuracy. With the preamplifier and the offsets nulled at V_{IN} = 10 mV at 1 kHz, the circuit's error is less than 0.5% of reading for inputs from 5 mV rms to 500 mV rms for frequencies from 40 Hz to 20 kHz. The 1% bandwidth of the AD744 preamplifier by itself (measured using a Fluke 931B RMS differential voltmeter) is 81 kHz at 10 mV input.

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Kitchin, Charles, and Counts, Lew, *RMS-to-DC Conversion Application Guide, Second Edition,* Analog Devices, Inc., Norwood, MA, 1986.

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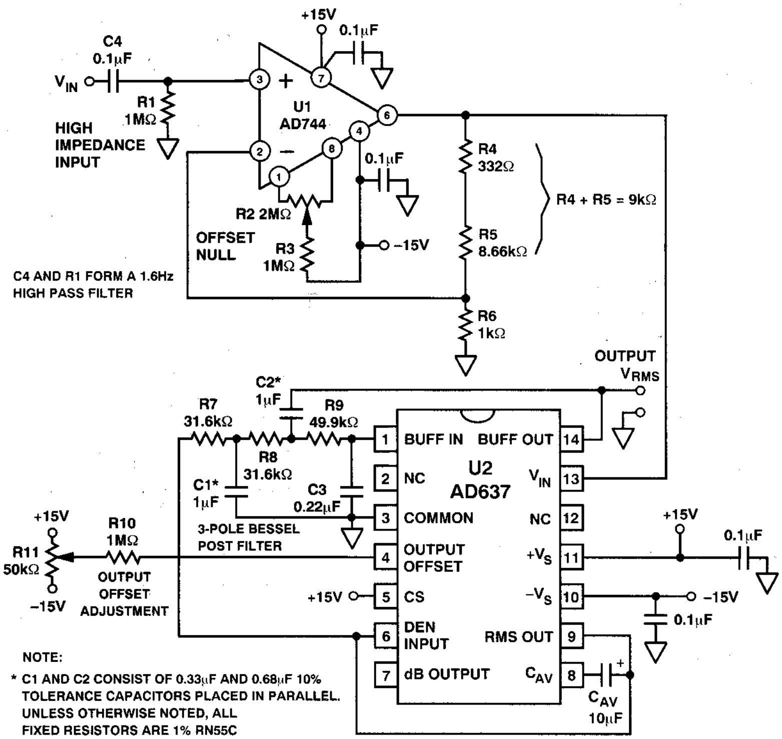

Figure 23. By combining a preamplifier and a 3-pole post filter, you can build an AD637 circuit that measures low level signals with precision and settles quickly for step inputs.