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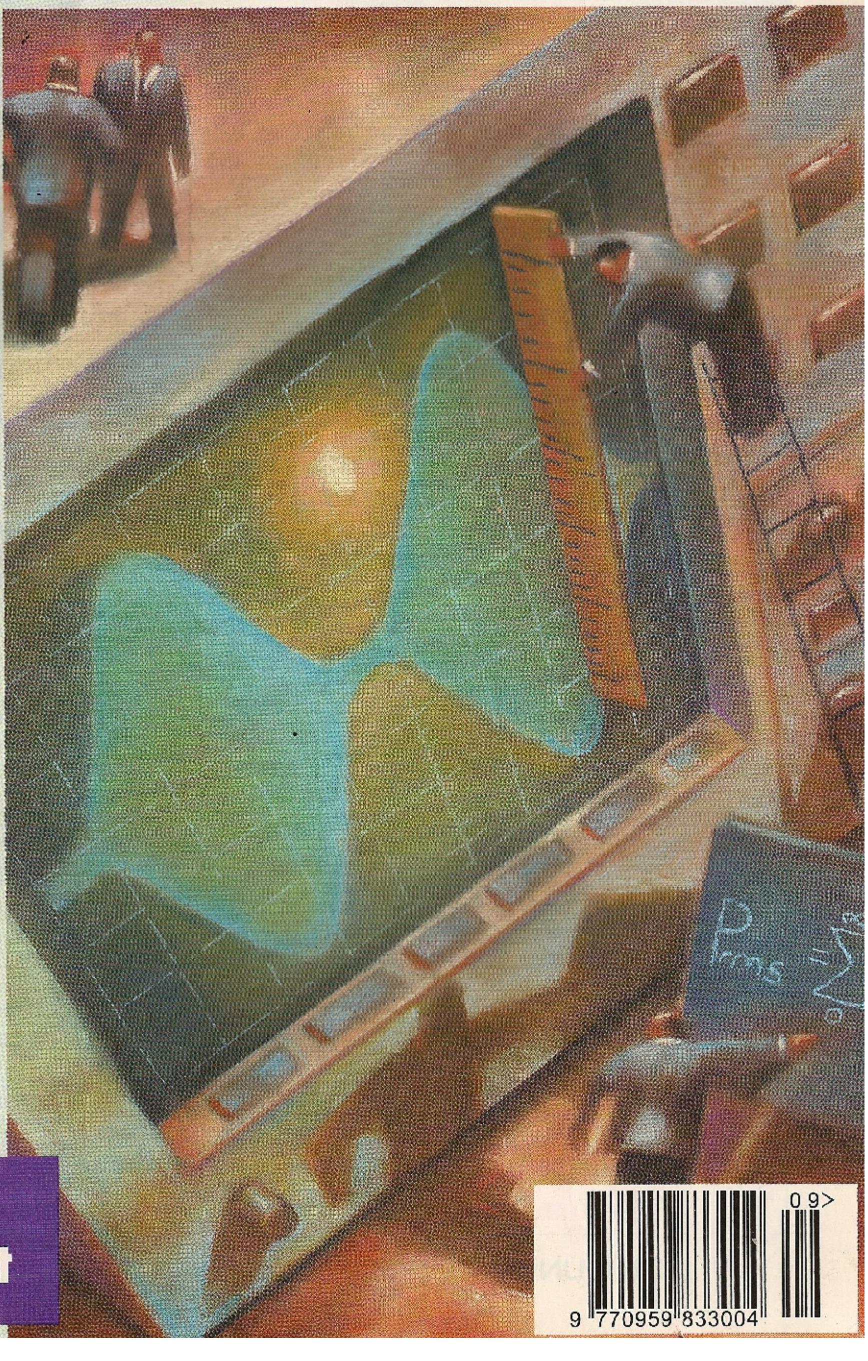

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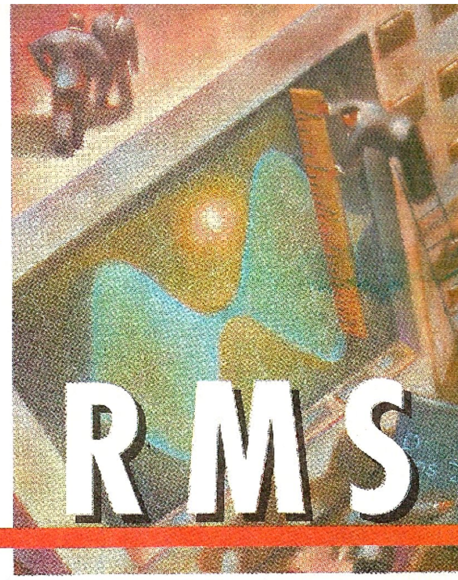
True rms measurement has become straightforward thanks to dedicated ICs. The information in this article has been compiled by Dan Ayers and *EW+WW* staff.



READING RMS

Since the advent of digital multimeters, engineers have been able to make quick, simple and relatively accurate voltage measurements for both dc and ac. A common and unhappy side effect of this however is an over-reliance on the seemingly definitive number on the readout.

Reference to the meter's specifications often shows that the last digit displayed may be far from the real value. A more fundamental question is whether even the range is appropriate. Although a low to middle priced multimeter is adequate for dc and certain ac measurements, a crucial range is usually missing - true rms.

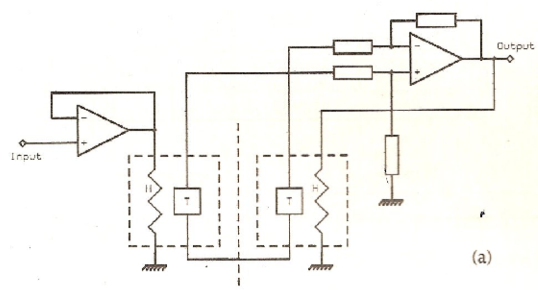
Why rms ?

Virtually all electronic systems call for some means of monitoring ac voltage. It is easy to obtain the peak, or peak-to-peak, value of a signal by pumping a capacitor with a rectifier, and subsequent op-amp buffering is straightforward. This is useful to indicate when an amplifier or similar system is approaching its clipping limits.

A strategy used in many ac voltmeters is to show the mean average deviation, or MAD, of a signal from a predetermined reference, usually the mean. This so-called ac average can be useful, but a more versatile measure is the rms voltage of a signal. This fundamental quantity provides information about the energy available or used over time.

When applied to a resistive load for a given period of time, any signal of the same rms voltage would cause the same amount of heat dissipation. Sometimes described as effective voltage, rms corresponds to the dc voltage that would produce the same heating effect.

Often, the mean average deviation is displayed on a scale calibrated in rms volts. But this setup only shows a correct reading when the waveform applied is of the same shape as the waveform used to calibrate the meter. Many digital multimeters only give a valid ac reading for fairly low frequency, sinusoidal waveforms below around 400Hz.



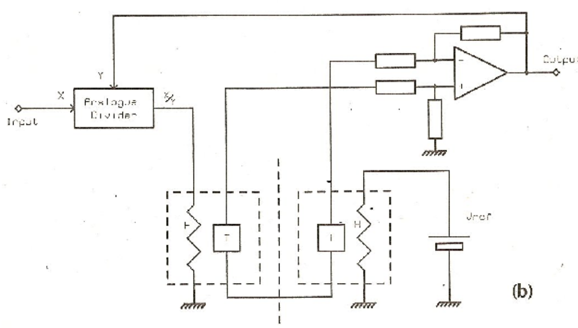


Fig. 1a). Deriving the rms value of a signal using two heater-sensor combinations. A DC voltage applied to a resistor produces exactly the same amount of heat as its equivalent in rms.
b). Dynamic range of the thermal converter is improved by adding analogue divider.

As long as the waveform is known, the true rms value of a signal can be calculated from the MAD. With many real-world signals such as noise and those associated with distortion however, this can cause problems. Comparing the MAD values with the true rms values for differing waveshapes clearly demonstrates the limitations, **Table 1**.

It is helpful that if unrelated signals are summed, then the rms of their sum is equal to the square root of the sum of the squares of their individual rms values. The rms value is also convenient for assessing signals with random characteristics. It represents the statistical standard deviation of a stationary zero-mean random process[1].

Circuit methods for true rms

For high accuracy, thermal methods of deriving the rms level of a signal are the most appropriate. This is because the heating effect of an ac voltage corresponds directly with the rms value, ie. that of the dc voltage required to produce the same heating in the same load. There are many drawbacks here, mainly due to the time taken for the temperature of different parts of the system to stabilise.

In the simplified thermal converter of **Fig. 1b**), two units, each comprising a heater *H*, thermally coupled to a temperature sensor *T*, are thermally insulated from each other. The first is heated by the applied signal, the second is forced by the difference amplifier to the same temperature. If both units have identical thermal paths to the environment, then the output voltage is proportional to the rms value of the input.

A practical system might have thermocouple sensors and a chopper-stabilised device for the difference amplifier. This configuration suffers from limited dynamic range. Power through the heaters is proportional to the *square* of the rms.voltage, and heater overload is a distinct possibility.

This problem is overcome in **Fig. 1b**), Here, the output amplifier still strives to maintain the temperature difference at zero, but now the power in the second heater is fixed. An analogue divider maintains equilibrium as its control voltage Y is proportional to the rms of input voltage X. As a result, the rms function is provided without the heaters having to function over an unmanageable range[1].

Table 1. Comparison between mean absolute deviation and rms voltages for common waveforms. Mean absolute deviation is also known as ac average.

Waveform		RMS	MAD	CF
V_p	DC	V_p	V_p	1
V_p	Sine	$\frac{V_p}{\sqrt{2}}$	$\frac{2V_p}{\pi}$	$\sqrt{2}$
V_p	Triangle	$\frac{V_p}{\sqrt{3}}$	$\frac{V_p}{2}$	$\sqrt{3}$
V_p, t, T	Pulse	$V_p\sqrt{\frac{t}{T}}$	$\frac{V_p t}{T}$	$\frac{1}{\sqrt{\frac{t}{T}}}$
$2V_p$	Gaussian (white) noise	—	$RMS \times \sqrt{\frac{2}{\pi}}$	∞ typically 1 - 6

Convenience is much enhanced by using computational elements to obtain the rms value. Analogue-to-digital converters and digital processing are relatively expensive however. Fortunately, old-fashioned analogue techniques with modern manufacturing methods have resulted in accurate and easy to use integrated circuits.

The complete function required is:

$$E_{rms} = \sqrt{\left(\frac{1}{T}\int_0^T V_{in}^2 dt\right)}$$

Computation is simplified by considering the integration and division by T as a running average. In practice, this is valid for most types of signal encountered, so:

$$V_{rms} = \sqrt{\overline{V_{in}^2}}$$

There are two basic approaches to obtaining the true rms value of a signal – explicit and implicit[1]. The explicit or direct approach is shown in **Fig. 2 (a)**. Two inputs of a four-quadrant multiplier are fed with the input signal, producing a squaring function. Positive-going voltage created is averaged over time, and the square root of this dc value is taken. This can be done by inserting a squarer into the negative feedback loop of an amplifier.

Although good accuracy is possible, this approach is more complex and more expensive. In addition, dynamic range is at least an order of magnitude narrower than with a comparable implicit arrangement.

Dynamic range is particularly significant when measuring signals with a high crest factor, or cf. This is the ratio of peak to rms voltage. Obtaining a valid measure of a signal with a large crest factor needs a proportionately greater headroom.

The implicit approach follows from a little manipulation of the rms equation to:

$$V_{rms} = \frac{\overline{V_{in}^2}}{V_{rms}}$$

producing the more elegant configuration Fig. 2b). Assuming an adequate CR time constant, the rms voltage output is held constant over the period of the signal being averaged and division by this value can be carried out before the average is taken.

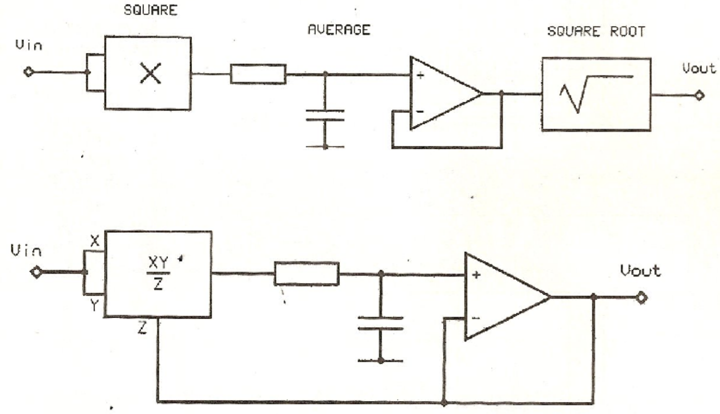


Fig. 2. Computation of rms voltage can be explicit, but implicit computation, (b lower), provides greater dynamic range.

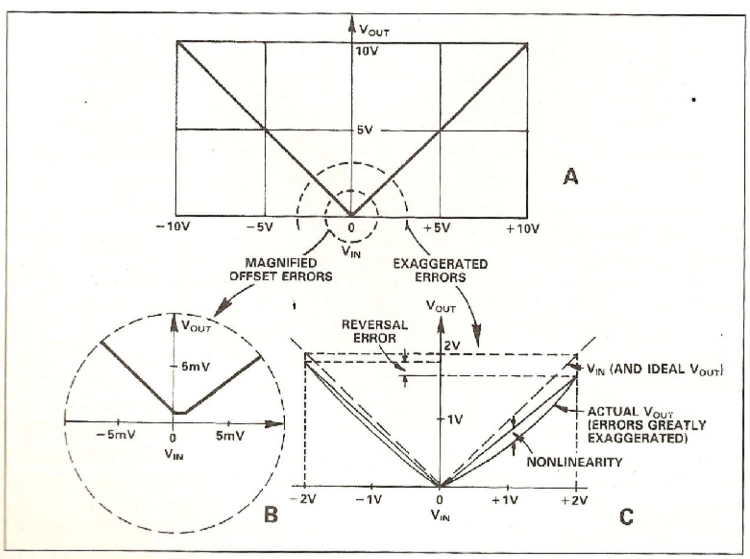


Fig. 3 (bottom). Static errors in rms-to-dc converters. These errors are combined and expressed as a percentage of reading plus a constant.

Error sources in rms conversion

An ideal rms converter provides a dc output voltage exactly equal to the rms value of its input voltage, regardless of the amplitude, frequency, or shape of the input waveform. Of course a practical rms converter has errors.

Static errors are offsets and scale factor errors that apply to dc and low-frequency sinewave to about1kHz. Under these conditions, the finite bandwidth of the converter – and the effective averaging time – can be made negligible compared to the input and output offset, and scale factor errors. Here, rms can be interpreted as the square root of the low pass filtered, or averaged, square of the input voltage.

An rms to dc converter's overall 'static' error is specified in percent of reading plus a constant. As shown in **Table 2**, the *AD637J* is specified at 1mV +0.5% of reading. This should be interpreted to mean that at any point within the *AD637J*'s 0V to 7V rms input dynamic range, converter output voltage will differ from the precise value of the rms input by at most 1mV plus 0.5% of the correct rms level. Note that this is less absolute error than the *AD536AJ* converter.

To illustrate this point, consider a sinewave input of 1V rms at 1kHz applied to the input of an *AD637J*. Actual *AD637* output voltage will be within: ±(1mV+0.5%×1V)=±(1mV+5mV). This is 6mV from the ideal output of 1.0V, or between 0.994 and 1.006V dc. These static errors can be classified into the standard categories of offset voltage, scale factor (gain) error, and nonlinearity errors.

Every practical rms converter has an input/output transfer characteristic that deviates from the ideal. The detailed error explanation given by **Figures 3a,b)** illustrate the major classes of errors commonly encountered.

At low levels, the rms converter's input offset voltages can flatten the point of the ideal absolute value transfer and take it more positive relative to the zero output voltage level with zero input voltage applied. Practical effects of these offset errors determine both the resolution and accuracy of the converter for low-level input signals.

For the ICs discussed here, the combined total of offset errors is typically less than 1mV. At higher input levels, of the order of few hundred millivolts, scale factor and linearity errors may dominate offset errors. A scale factor error is defined as the difference between the average slope of the actual input/output transfer and the ideal *I* to *I* transfer. If a 100mV rms input change produces a 99mV change in output, then the scale factor error is 1%.

In addition to the single polarity example just given, there can be a different scale factor for both negative and positive input voltages. The difference in these scale factors, termed the 'dc reversal error', is shown in Fig. 3c). When testing this parameter, a dc voltage is applied to the converter's input, say +2V, and then the polarity of the input voltage is reversed to –2V. Difference between the two readings will equal the dc reversal error.

Nonlinearity, as its name implies, is the curved portion of the input/output transfer characteristic. This is shown in an exaggerated form in Fig. 3c. This error is due to non-ideal behaviour in the rms computing section and cannot be reduced by trimming offset or scale factor.

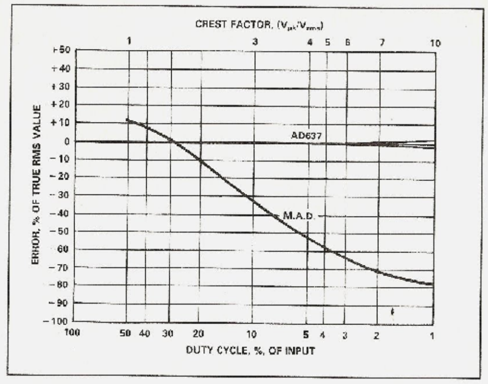


Fig. 4. Error versus duty cycle for an MAD ac detector and AD637-based rms converter.

Therefore, nonlinearity sets a limit on the ultimate best case accuracy of the rms converter.

For the *AD637*, nonlinearity is typically better than 1mV (0.05%) over a 2V full-scale rms range; for the *AD536A* the nonlinearity equals 5mV or less. Typically the *AD636* has less than 1mV nonlinearity over its 0 to 200mV specified input range.

As shown by **Fig. 4**, the errors of true rms to dc converters, although varied, are considerably lower than those errors found in precision mean-absolute deviation rectifiers when the duty cycle of the input waveform is varied.

TABLE 2: Typical rms-to-dc converter specifications.

	AD536AJ	AD637J	AD636J
Input dynamic range	7Vrms	7Vrms	1V rms
Nominal fsd rms	2V rms	2V rms	200mV
Peak trans. Input	±20V	±15V	±2.8V
Max total error			
No external trim			
mV/% reading	5mV ±0.5	1mV ±0.5	0.5mV ±1
Bandwidth, (-3dB)			
Full Scale	2MHz	8MHz	1.3MHz
0.1 V rms	300kHz	600kHz	800kHz
Error at Crest Factor			
of 5, rms	-0.3%@1V	±0.15%@1V	-
0.5%@200mV			
Power supply			
Volts min	±3	±3	+2/-2.5
max	±18	±18	±12
Current typ.	1mA	2mA	800μA
max	2mA	3mA	1mA

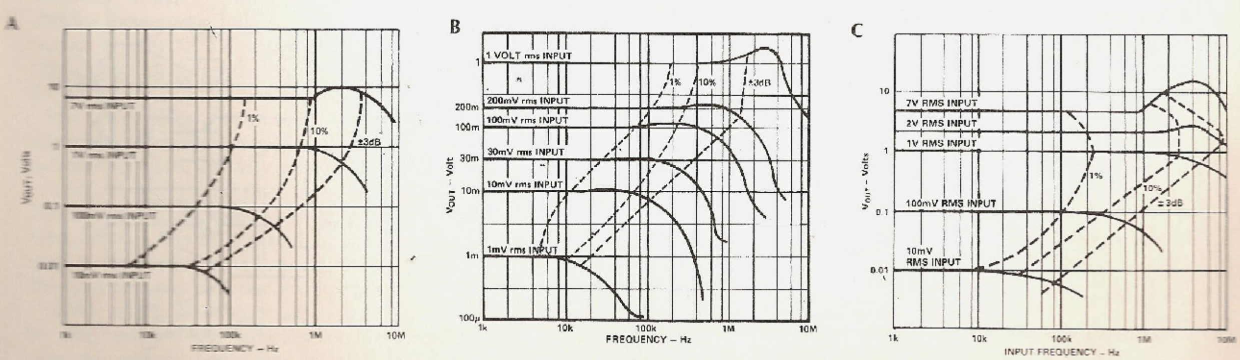


Fig. 5. High frequency response for the three converters – AD536A *at a),* AD636 *at b) and* AD637 *at c).*

Bandwidth considerations

In practice, ac inputs are of the most interest to users of rms converters. For 1kHz sinewave inputs, there is negligible difference between readings at this frequency and performance at dc. As a result, dc measurements provide a convenient way of determining errors at around 1kHz.

At higher input frequencies, bandwidth characteristics of the rms converter become most important. As shown by **Figs 5a,b,c)**, ac bandwidth drops off as the input level is reduced; this is primarily due to gain-bandwidth limitations in the absolute value circuits.

Caution should always be used when designing rms measuring systems which must deal with complex waveform amplitudes above 1V rms. Trimming is recommended for applications needing the lowest possible offset and scale factor errors, **Figs 6a,b,c)**. Ground the signal input point, *V*, and adjust trimmer R_4 for an output of zero volts. Alternatively, *R* can be adjusted to give the correct output with the lowest expected value of V_{IN} applied. This second method allows the lowest possible error over the expected input range, but results in higher errors below this range.

Connect a 1kHz calibrated full scale input to V_{IN}. Adjust trimmer R_1 to give the same output voltage. This adjustment provides specified accuracy with a 1kHz sinewave input and slightly less accuracy with other input waveforms.

With correct trimming, the remaining errors in an rms converter will be due to nonlinearity effects of the device; unfortunately, nonlinearity errors cannot be reduced by external trimming.

Practical circuits

Although it is possible to produce close approximations to squaring and square root functions relatively directly, log/antilog blocks can give greater accuracy and simplify initial setting up[2]. These blocks are often based on the exponential response of transistors.

Figure 7 uses two standard chips to produce a log/antilog implicit rms converter which is adequate for many applications; the separate computing elements are clear to see.

Several companies produce dedicated rms chips, and the circuit of **Fig. 8** shows how straightforward such devices are to apply. The *SSM2110* is a particularly versatile device. With a minimum of external components it can provide rms, absolute value and peak conversion, or alternatively the log of any of these[3]. **Figure 9** would be suitable for a meter calibrated in decibels – very useful for audio work.

High impedance input rms dpm and dB meter

Only two integrated circuits and a liquid crystal display are needed to produce high quality, dpm/dB meter.

Voltage input to the meter feeds through a 10MΩ input attenuator to pin 7 of the *AD636*. Buffer output, pin 6, is ac coupled to the rms converter's input, pin 1. Resistor R_6 provides a 'bootstrapped' circuit to keep the input impedance high.

Output from the rms converter is selected by the linear/dB switch; selecting pin 8 for linear, pin 5 for dB. The selected output travels from the linear/dB switch through low pass filter R_{15}, C_6 to the input of the meter chip, which is a *7106* type a-to-d converter. The *AD589*

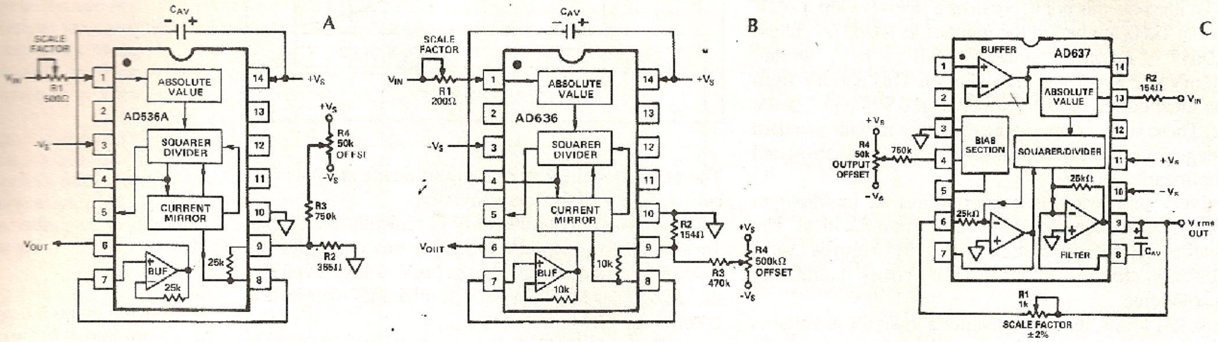


Fig. 6. Internal functions together with external offset and scale factor trimming circuits for AD536A *at a),* AD636 *at b) and* AD637 *at c).*

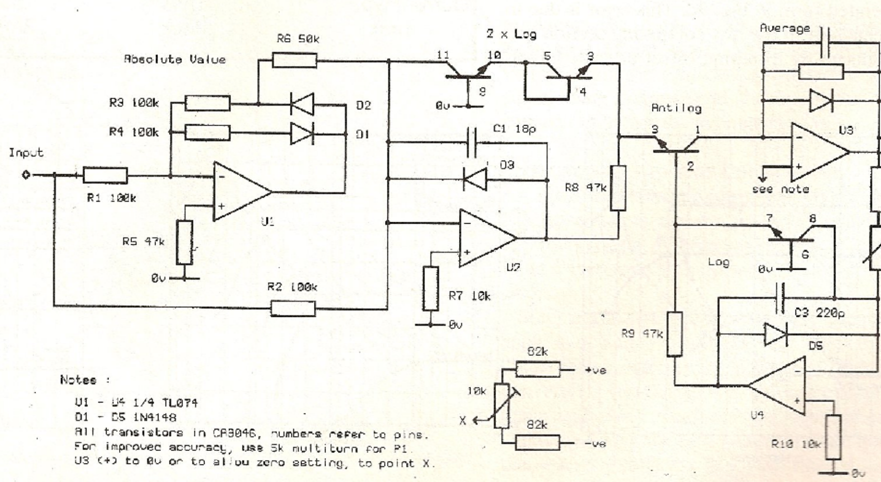


Fig. 7. Converter for rms measurement using standard chips shows log-antilog calculation of square/square-root functions.

provides a stable 1.2V reference voltage for supplying the calibration circuitry.

To calibrate, first adjust trimmer R_9 for the 0dB reference point. Next, set R_{14} for the decibel scale factor, and finally, adjust R_{13} to set the linear scale factor. Total current consumption is typically 2.9mA from a standard 9V transistor radio battery.

This circuit uses the *AD636* low power rms converter to extend battery life and provide a 200mV full scale sensitivity. It provides better accuracy and bandwidth at 200mV rms input than the *AD536A*, which would need preamplifier to achieve similar results.

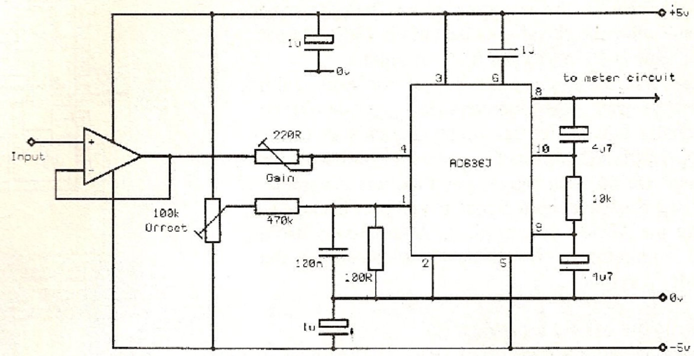


Fig. 8. Dedicated rms converter chips can reduce component count and improve accuracy.

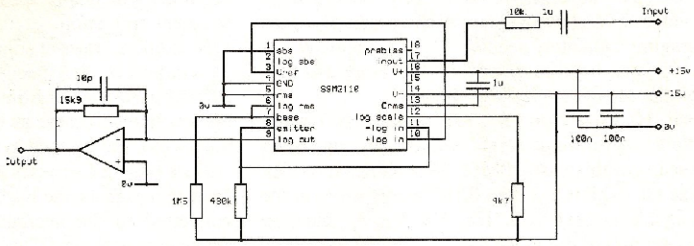


Fig. 9. Versatile converter chip configured for log of rms conversion. This configuration is useful for audio decibel metering.

Programmable-gain rms measurement

Measurement of the rms of complex waveforms of varying magnitude normally requires a high quality, compensated input attenuator. In contrast, the programmable gain rms preamplifier circuit of **Figure 10** features an *AD544* bifet operational amplifier as an inverting input buffer with four remotely switchable gain ranges: 200mV, 2V, 20V, and 200V full scale.

Switching gain resistors in the buffer feedback loop allows the use of a low voltage cmos multiplexer to remotely control the gain of potentially high voltage input signals. The preamplifier's input is well protected on all ranges for input voltages up to 500V peak.

Input connects to J_1, with R_1 and diodes $D_{1,2}$ forming the amplifier's input protection. Capacitor C_1 prevents high frequency roll-off, which would occur due to the *R/C* time constant of the 1MΩ input resistor and the stray capacitance at the *AD544* summing junction. The *AD7503* cmos multiplexer switches the appropriate feedback resistor for each gain connecting the resistor between the operational amplifier output, pin 6, and its summing junction, pin 2.

Capacitors $C_{4\text{-}7}$ are compensation capacitors which are adjusted for flat response at each gain setting. Address lines $A_{0\text{-}2}$ select the desired input range of the preamplifier. Resistors $R_{4,6,10,12}$ are gain calibration controls for each selected gain. Output of the *AD611* operational amplifier is converted to its rms equivalent voltage by the *AD536A* rms-dc converter.

Input ranges are 200mV, 2V, 20V and 200V rms. For the respective ranges, –3dB bandwidth points are >4kHz, 600kHz,1.5MHz and 600kHz. For the lowest range, bandwidth will vary with the degree of stray capacitance at pin 9 of the *AD7503*.

Testing converters

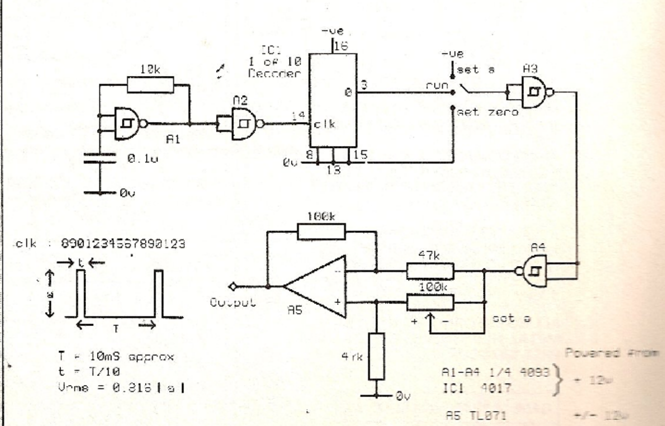
To calibrate and assess the accuracy of an rms converter, many factors need to be considered – particularly the dc response (offset), frequency response (gain) and dynamic range. Laboratory equipment is desirable, but a good overall picture can be gained by feeding a pulse waveform of known amplitude and mark/space ratio into the converter. This is because the pulse contains frequency components extending to infinity – in theory at least – and calculation of the crest factor and true rms value is straightforward.

The circuit shown generates reasonable pulses with variable amplitude and a mark/space ratio fixed at 1:10. A simple clock with a frequency around 1kHz is built around a Schmitt trigger. This frequency may be varied over a wide range by altering the resistor and/or capacitor values. Clock output is sent to a 1 of 10 decoder to fix the mark/space ratio and the pulses are cleaned up by the remaining Schmitt triggers.

For controlling amplitude, an op-amp is configured to provide variable gain giving an output pulses from –10V to +10V referred to ground. The circuit can suffer from ringing on the pulse edges. This affects the rms level, especially at higher pulse rates. Should this be a problem, it is advisable to strap a variable resistance of around 47kΩ between the op-amp input pins and trim for best shape.

Before testing an rms circuit, the zero should first be checked and any offset noted. A suitable dc reference voltage should then be set at, say, 5V. It is important that the circuit under test is connected before setting the reference to avoid loading errors. This will also confirm that the converter is responding correctly to dc.

Switching the pulse generator switch to run should produce a dc voltage at the output of the converter of around 1.6V. Its true rms value is 0.316×5=1.58V.



Simple pulse generator with fixed mark/space ratio and variable amplitude allows easy assessment of rms converter accuracy.

Noise referred to the amplifier input is 360μV on the 2V range while the signal-to-noise ratio is 75dB. Output settling time is 397ms to reach 1% of input.

Address lines $A_{0\text{-}2}$ should be set for each gain. Calibration trim potentiometers $R_{4,6,10,12}$ should be individually adjusted for the correct gain on each range.

Compensation capacitors $C_{5,6,7}$ should be adjusted for flat response on each range. For this, use a variable frequency sinewave input signal and an oscilloscope to monitor the *AD544* output, pin 6. Alternatively use a digital voltmeter on its dc scale connected to the converter's output.

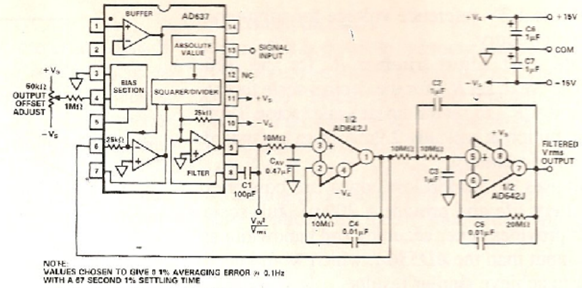


Fig. 10. With signals as low as 0.1Hz, this circuit exhibits less than 0.1% averaging error.

Reading ultra-low frequencies

Reducing input frequency requires lengthening the averaging and filtering time constants to maintain the same levels of dc error. Consequently, successively larger values of C_{AV} are needed. With very large values of averaging capacitor, needed for frequencies below 10Hz, C_{AV} can become physically too large and also prohibit the use of low-leakage devices.

Figure 10 uses two very low input bias current amplifiers, permitting large values of averaging resistance – in this case 10MΩ. This circuit has been optimized to exhibit less than 0.1% averaging error for input signals as low as 0.1Hz. The V_{IN}^2/V function appears at pin 9 of the *AD637*.

As a result of transient noise spikes, the circuit may overload because the filter stage averaging capacitor has been drastically reduced. Normally, the averaging capacitor is called C_{AV} but in this case it has been renamed C_1. Reducing the capacitor allows output at pin 9 of the *AD637* to respond to the square of the input signal rather than to the average of the input square

For applications where high crest factor-low frequency signals are to be measured, C_1 should be increased to 3.3μF. In conjunction with the internal 25kΩ filtering resistor, this capacitor forms a low-pass filter with a 2Hz corner frequency. This attenuates higher frequency signals – transients – by the ratio of the transient frequency to that of 2Hz. This means that in the case of 60Hz transients, they will be reduced by 60Hz/2Hz or 30 times. Practically speaking, there will be effective transient protection.

In addition, larger or smaller values of C_1 may be used as required by the specific application. If a low-pass filter is used ahead of the *AD637*, out-of-band signals are less likely to cause an overload. This allows smaller values of C_1 to be used in these circuits.

Since raising C_1 causes increased averaging of higher frequency signals, the V_{IN}^2/V_{rms} function will be linearly converted to the average of V_{IN}^2/V_{rms} as the input frequency goes up. This prevents the instantaneous square of the input signal from appearing at pin 9 of the *AD637*. ■

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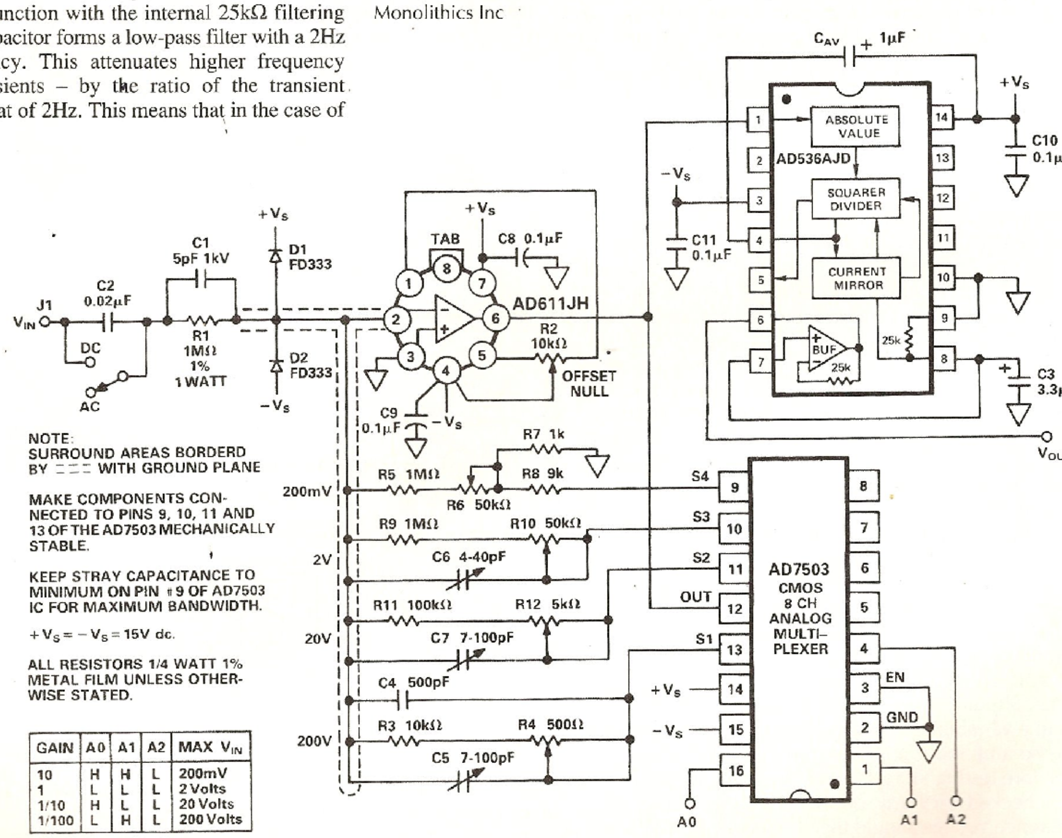
2. D.Ayers, *The Twisted World of Non-Linear Electronics*, *Electronics World & Wireless World*, Feb 1993.
3. *SSM Audio Products Audio Handbook*, Vol. 1, Precision Monolithics Inc

Fig. 11. Measuring rms of complex signals normally involves an expensive attenuator. This programmable-gain circuit, with 200mV, 2V, 20V and 200V ranges, does the same job.



GAIN	A0	A1	A2	MAX V_{IN}
10	H	H	L	200mV
1	L	L	L	2 Volts
1/10	H	L	L	20 Volts
1/100	L	H	L	200 Volts

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