SECTION III

RMS APPLICATION CIRCUITS

AUTOMATIC GAIN CONTROL (AGC)

An rms – AGC Amplifier

Introduction

One of the major problems facing the designer of function generators is how to maintain a constant output level with variations in waveform, duty cycle, and frequency. Many conventional AGC (Automatic Gain Control) amplifiers suffer from poor performance, usually a restricted dynamic range and inadequate frequency response. These amplifiers usually provide a constant "average" output level, or instead employ a peak limiting scheme which controls the maximum "peak" output level. In many applications, it is more useful to have a constant rms output level from an AGC amplifier.

The rms-AGC amplifier presented here has many audio uses and will provide a virtually constant rms output voltage over an input variation of greater than 40dB (see Figure 40).

Circuit Description

As shown in Figure 39, the voltage output from the AD534 multiplier is converted to its dc equivalent voltage by an AD536A rms to dc converter.

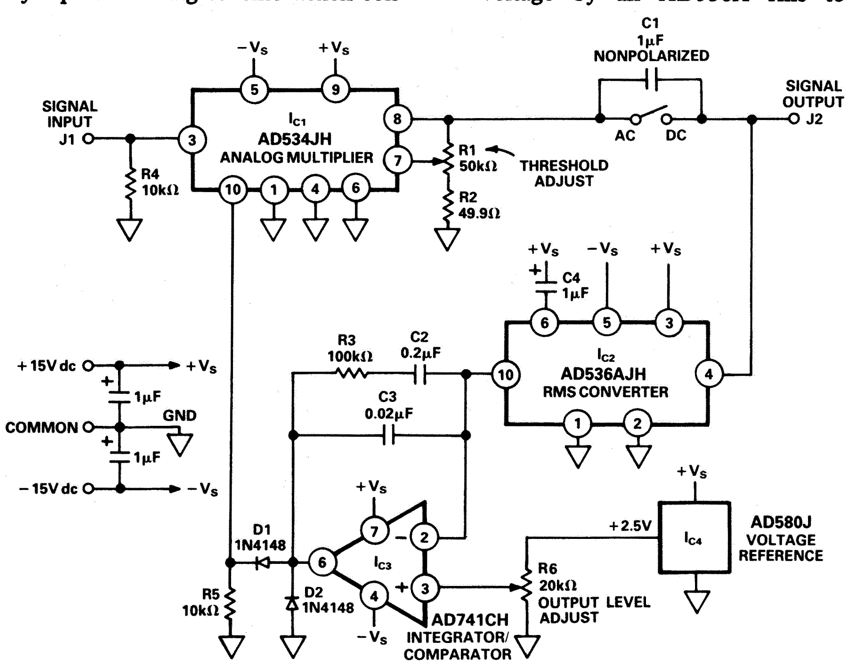


Figure 39. An rms AGC Amplifier

Capacitor C_4 is the rms converter's averaging capacitor. The output voltage from the converter is compared to a fixed reference voltage provided by an AD580 bandgap reference. An AD741 is used as an integrator/comparator amplifier whose output voltage sets the gain of the AD534 multiplier. The AD741 amplifies the difference between the rms output level from the AD536A and the preset dc voltage derived via the output level control, R_6. This amplified output voltage appears at pin 6 of the AD741 after a delay time set by the series/parallel RC combination of C_2C_3 and R_3 in the operational amplifier's feedback loop. The two diodes, D_1 and D_2, keep the output of the AD741 from going negative; this would change the phase of the control loop by 180°.

Performance Data

All measurements were taken with 300mV threshold and 1 volt output level. Note: Different types of waveforms may have widely varying crest factors; consequently, even though they have the same rms level, their peak values may be quite different. Since amplifier overload occurs due to the peak values of the signal waveforms, all voltage specifications in this circuit, and in the audio AGC amplifier, are given in peak to peak.

Input Range: 300mV to 28 volts (40dB)
Frequency Response: 10Hz to 400kHz @ 300mV input, dc to 1MHz @ 1 volt input.
Signal to noise ratio: 65dB
Attack/hold-in time: 100ms

An Audio rms – AGC Amplifier

Introduction

Here is an audio amplifier incorporating a smooth sounding automatic gain control with a gradual acting threshold level. This circuit eliminates the usual audible "thump" of most compressor amplifiers by slowly incorporating the AGC action (Figure 42).

This design offers a great amount of flexibility featuring: controls for input range, degree of compression, and output level.

Circuit Description (see Figure 41)

The audio input signal, adjusted by R_4, is amplified by an AD544 operational amplifier operating with a gain of 21. The AD544 voltage output drives the controlled gain stage, an AD534 analog multiplier. The 2.49kΩ resistor in series with pin 2 of the multiplier may be varied from 2kΩ to 3kΩ for the best compromise between bandwidth and signal to noise ratio. This resistor sets the trade-off of gain versus bandwidth of the analog multiplier.

The multiplier's output is ac coupled to the signal output jack and to an AD536A rms – dc converter. The rms converter's current output drives a comparator/amplifier, IC_4, which acts as a current to voltage converter. The comparator's output, a positive voltage, is *decreased* by the rms converter's output, reducing the gain of the multiplier. The comparator's threshold point is prevented from being too abrupt by resistor diode networks, D_4/R_7 and D_5/R_8. The attack/release times of the circuit are determined

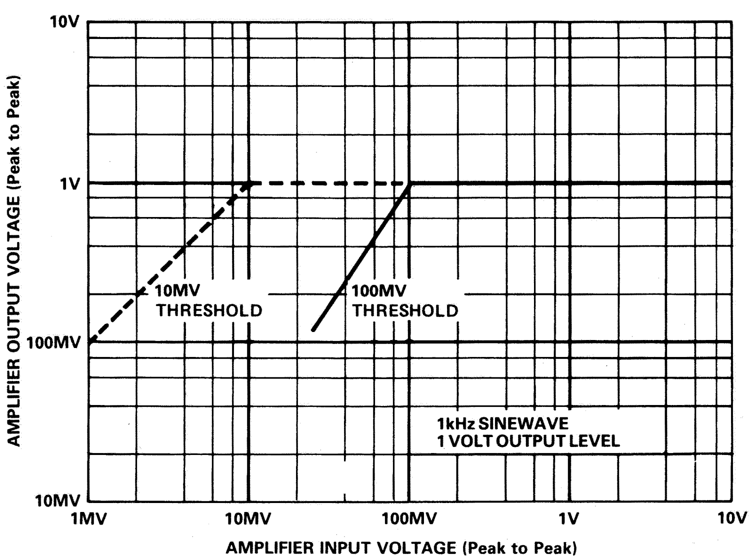


Figure 40. Input vs. Output – rms AGC Amplifier

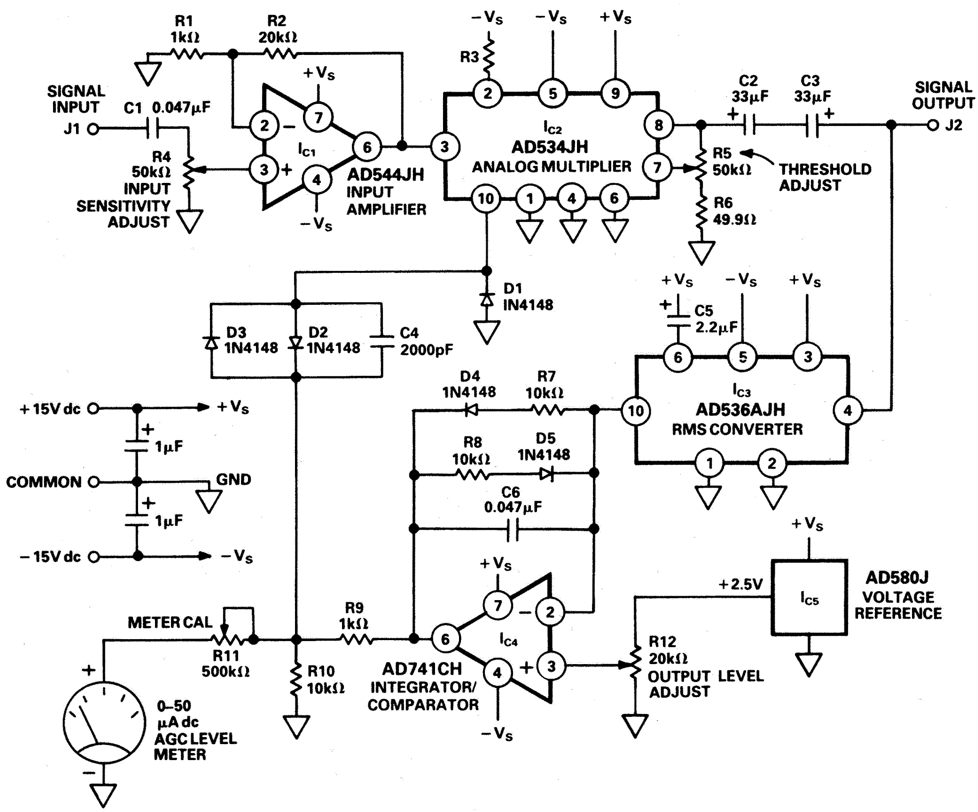


Figure 41. An Audio rms AGC Amplifier

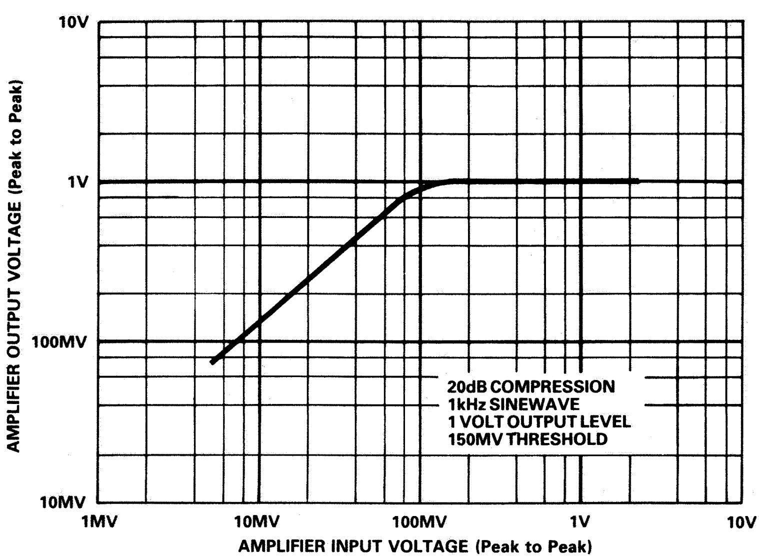


Figure 42. Input vs. Output – Audio rms AGC Amplifier

mainly by capacitors C_5, C_6 and resistor R_7 and to a lesser extent, resistor R_8. Diodes D_1, D_2, and D_3 prevent latch-up of the multiplier by voltage transients. The degree of AGC action is monitored by the 0-50μA analog meter.

Performance Data
All voltages are specified peak to peak

Input range: 150mV to 10 volts
Output level range: variable 0.5 to 2.5 volt p-p
Compression: variable from 0dB to 26dB
Threshold level ± 150mV to 1.6 volts

Frequency Response
10dB Compression
50Hz to 65kHz with 0.5V p-p input
50Hz to 100kHz with 1.0V p-p input
70Hz to 160kHz with 1.65V p-p input

20dB Compression
70Hz to 75kHz with 0.5V p-p input
70Hz to 120kHz with 1.0V p-p input
100Hz to 160kHz with 1.5V p-p input
With 1 volt p-p input and 1 volt p-p output level
attack time: 250ms, release time: 80ms

Signal to noise ratio
10dB compression 51dB, 20dB compression 45dB
Optimum output level: 1 volt p-p
Total harmonic distortion: 0.30%

To adjust the amplifier, first apply a 1 volt peak to peak sinewave at 1kHz to the input jack J_1. Adjust R_4 for a 2.5 volt peak to peak sinewave at the input of the multiplier, pin 3 of I_{C2}. Adjust R_5 to midposition. Next, adjust R_{12}, the output level control, for the desired output level, nominally 1 volt peak to peak. Adjust R_5 again, this time for the degree of compression desired. Trimpot R_{11} adjusts the output level meter's amplitude. Care should be taken when using the AGC amplifier to insure that the input is not overloaded (to avoid clipping the signal).

INSTRUMENTATION
RMS DIGITAL PANEL METERS (DPMs)

A Low Cost True rms Digital Panel Meter
This low cost DPM (Figure 43) features direct-reading true rms, a high impedance buffered input, four input ranges, and a minimum number of components. The DPM operates from a single 5 volt power supply requiring a total current of 100mA.

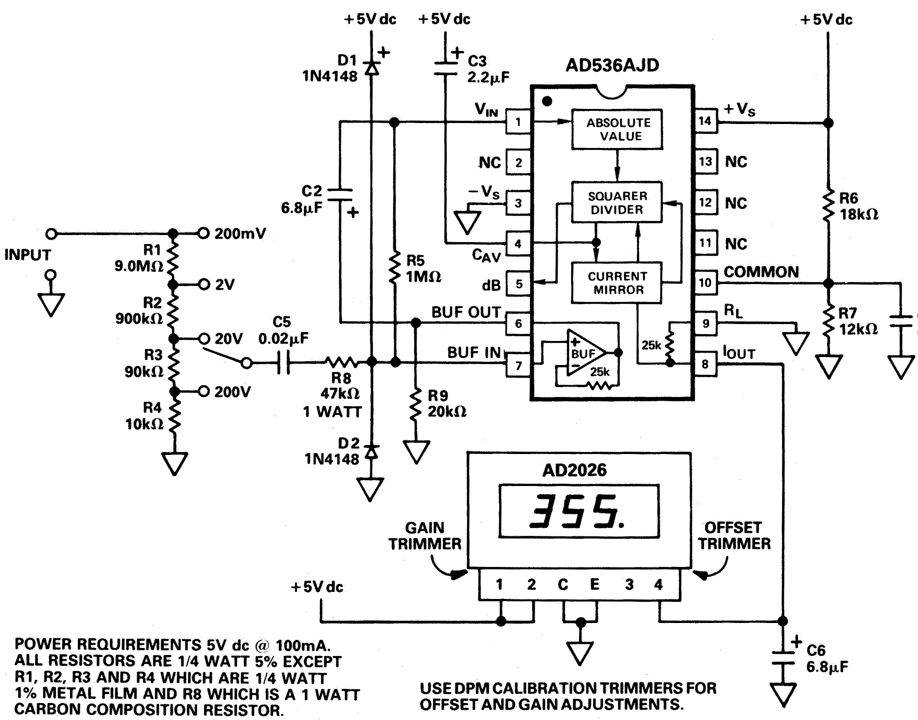


Figure 43. A Low Cost True-rms DPM

The input circuit of the device consists of a 10 meg ohm input attenuator with switch S_1 selecting the desired full scale input range. Capacitor C_5 ac couples the input of the rms converter's internal buffer amplifier (pin 7 of the AD536A) with resistor R_8 and diodes D_1 and D_2 providing input circuit protection. The output of the buffer, pin 6, is ac coupled to the rms converter input, pin 1. Resistor R_5 provides a "bootstrapping" return path for the buffer's input bias current; however, it does not affect the DPMs input impedance because the buffer is a unity gain follower, and pins 1 and 7 are at the same potential (see Appendix B). Resistor R_9 serves as a load resistor for the output of the buffer amplifier while resistors R_6 and R_7 provide a "floating" ground allowing single supply operation. Capacitor C_4 keeps the AD536A common, pin 10, at ac ground.

The output from the rms-dc converter (pin 8 of the AD536A) is low pass filtered by capacitor C_6; it then drives an AD2026 DPM. The rms meter's offset and scaling adjustments are made using the DPM's internal calibration trimmers.

An ac line-powered version of the AD2026 is available which will permit this circuit to operate from power supplied by the DPM itself, thus eliminating the need for an external 5 volt power supply.

A Portable High Impedance Input rms DPM and dB Meter Circuit

This high quality DPM/dB meter requires only two integrated circuits, their support circuitry, and a liquid crystal display.

As in the low cost DPM, the voltage input to the portable DPM runs through a 10 meg ohm input attenuator to pin 7 of the AD636. The 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. The 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 DPM chip's input. (The DPM chip is a 7106 type A/D converter.) The AD589 provides a stable 1.2V reference voltage which supplies the calibration circuitry.

To calibrate the meter, first adjust trim potentiometer R_9 for the 0dB reference point; next, set R_{14} for the dB scale factor, and finally, adjust R_{13} to set the linear scale factor. The total current consumption of the portable DPM is typically 2.9mA from a standard 9 volt transistor radio battery. This circuit utilizes the AD636 low power rms converter to extend battery life and to provide a 200mV full scale sen-

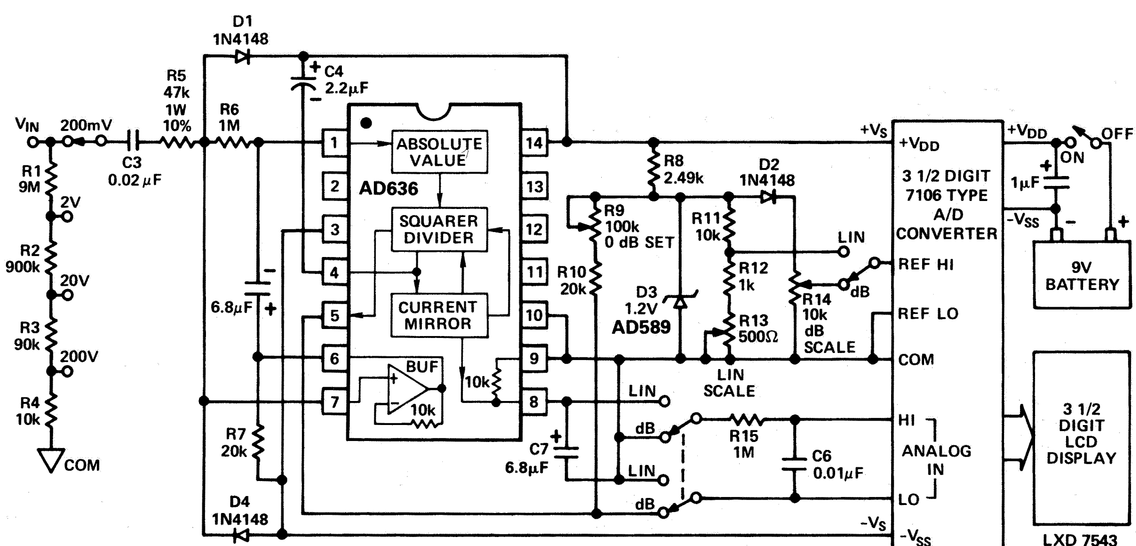


Figure 44. A Portable, High Z Input, rms DPM and dB Meter Circuit

sitivity. The AD636 gives better accuracy and bandwidth at 200mV rms inputs than the AD536A (the AD536A would require a gain of 10 preamplifier to achieve similar results at these levels.)

A Low Power, High Input Impedance dB Meter

Introduction

The portable dB meter circuit featured here combines the functions of the AD636 rms converter, the AD589 voltage reference, and a μA776 low power operational amplifier. It is also inexpensive, approximately $25.00. This meter offers excellent bandwidth and superior high and low level accuracy while consuming minimal power from a standard 9 volt transistor radio battery.

In this circuit, the built-in buffer amplifier of the AD636 is used as a "bootstrapped" (see Appendix B) input stage increasing the normal 6.7kΩ input Z to an input impedance of approximately $10^{10}\Omega$.

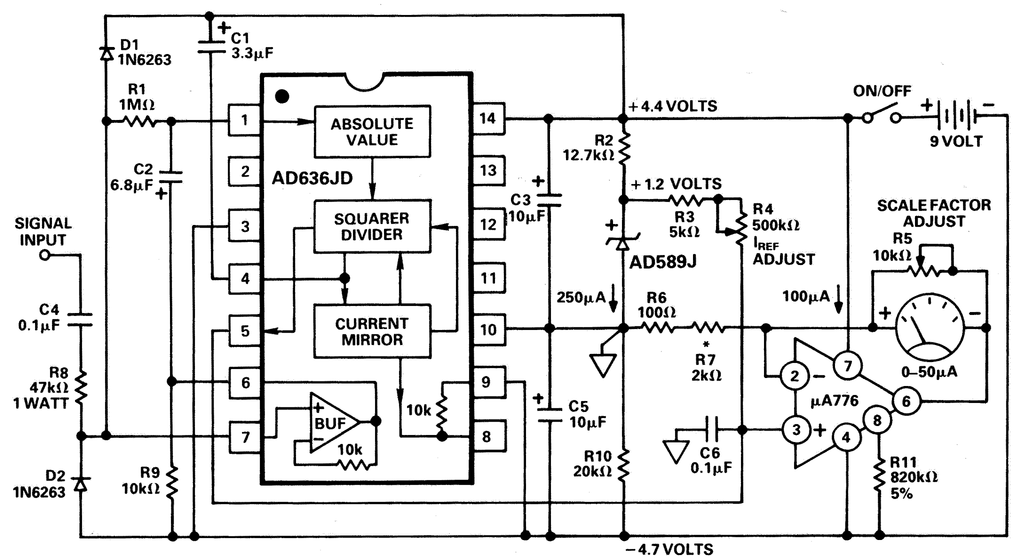
Circuit Description

The input voltage, V_{IN}, is ac coupled by C_4 while resistor R_8, together with diodes D_1 and D_2, provide high input voltage protection.

The buffer's output, pin 6, is ac coupled to the rms converter's input (pin 1) by capacitor C_2. Resistor R_9 is connected between the buffer's output, a Class A output stage, and the negative supply to increase the buffer amplifier's negative output swing (see Appendix B). Resistor R_1 is the amplifier's "bootstrapping" resistor.

With this circuit, single supply operation is made possible by setting "ground" at a point between the positive and negative sides of the battery. This is accomplished by sending 250μA from the positive battery terminal through resistor R_2, then through the 1.2 volt AD589 bandgap reference, and finally back to the negative side of the battery via resistor R_{10}. This sets ground at 1.2 volts +3.18 volts (250μA × 12.7kΩ) = 4.4 volts below the positive battery terminal and 5.0 volts (250μA × 20kΩ) above the negative battery terminal. Bypass capacitors C_3 and C_5 keep both sides of the battery at a low ac impedance to ground. The AD589 bandgap reference establishes the 1.2 volt regulated reference voltage which together with resistor R_3 and trimpot R_4 set the zero dB reference current, I_{REF}.

The 3mV/dB scale factor of the dB output (pin 5 of the AD636) is changed to a more convenient 10mV output per dB input by the μA776 operational amplifier. Resistor R_{11} sets the amplifier's quiescent current at 100μA. Temperature compensation is provided via the series combination of resistors R_6 and R_7 which together produce an equivalent 2.1kΩ + 3325ppm/°C TC resistor (see the decibel input provision section).



ALL RESISTORS 1/4 WATT 1% METAL FILM UNLESS OTHERWISE STATED EXCEPT
* WHICH IS 2kΩ +3500ppm 1% TC RESISTOR.

Figure 45. A Low Power, High Input Impedance dB Meter

Performance Data

0dB Reference Range = 0dBm (770mV) to −20dBm (77mV) rms
0dBm = 1 milliwatt in 600Ω
Input Range (at I_{REF} = 770mV) = 50dBm
Input Impedance = approximately $10^{10}\Omega$
V_{SUPPLY} Operating Range +5V dc to +20V dc
$I_{QUIESENT}$ = 1.8mA typical

Accuracy with 1kHz sinewave and 9 volt dc supply:
0dB to −40dBm ±0.1dBm
0dBm to −50dBm ±0.15dBm
+10dBm to −50dBm ±0.5dBm

Frequency Response ±3dBm:
Input
0dBm − 5Hz to 380kHz
−10dBm = 5Hz to 370kHz
−20dBm = 5Hz to 240kHz
−30dBm = 5Hz to 100kHz
−40dBm = 5Hz to 45kHz
−50dBm = 5Hz to 17kHz

Battery Life
Using a standard 250mA/hour 9 volt transistor radio battery, normal battery life with the meter left on will be between 100 and 150 hours. A ten-fold increase in battery life can be achieved using a 2500mA/hour mercury power pack battery which should operate the circuit continuously for about two months. If a 9 volt nickel cadmium rechargeable battery such as the Eveready N88 is used, it can be kept charged by solar cells, thus allowing maintenance-free operation. Requiring only about 1.8mA quiescent current, this meter lends itself well to many remote-site applications where changing batteries is inconvenient and expensive.

Calibration
1. First calibrate the zero dB reference level by applying a 1kHz sinewave from an audio oscillator at the desired zero dB amplitude. This may be anywhere from zero dBm (770mV rms – 2.2 volts p-p) to −20dBm (77mV rms 220mV – p-p). Adjust the I_{REF} cal trimmer for a zero indication on the analog meter.
2. The final step is to calibrate the meter scale factor or gain. Apply an input signal −40dB below the set zero dB reference and adjust the scale factor calibration trimmer for a 40μA reading on the analog meter.

Some final comments:
This meter is protected for input voltages up to 200 volts dc and has an input impedance of 10,000MΩ. Therefore, it is clearly superior to the circuits of Figures 37 and 38 for most all portable applications where the device may be exposed to all types of input signals and where very low power consumption is important.

Note: For the best possible resolution, the largest practical size analog meter movement should be chosen for use with this circuit.

The temperature compensation resistors for this circuit may be purchased from: *Tel Labs Inc*, 154 Harvey Road, P.O. Box 375, Londonderry, NH 03053, Part #Q332A 2kΩ 1% +3500ppm/°C or from *Precision Resistor Company*, 109 U.S. Highway 22, Hillside, NJ 07205, Part #PT146 2kΩ 1% +3500ppm/°C.

A Modem Line Monitor
This is a telephone line dB meter and line voltage sensor for the accurate monitoring and adjustment of telephone signal levels. The unit has a 600Ω line terminator and a voltage sensor to detect dc voltage on the line. When switched on, the line terminator also keeps ringing voltage (90 volts @ 20Hz) off the line being measured by making the phone line appear busy to the telephone switching equipment. The user may self-check the meter calibration by pressing the calibrate switch on the front panel.

The input signal is ac-coupled by C1 and runs through R_1 to pin 1 of the AD536A. Diodes D_1 and D_2 along with resistor R_1 protect the input of the AD536A from voltage spikes with capacitor C_2 and resistor R_1 forming a low pass filter.

The dB output from pin 5 of the AD536A runs directly to the input of the buffer amplifier. Zero adjust trimmer R_5 and resistor R_4 set the zero dB point on the analog meter by adjusting the amount of offset current from the AD580 voltage reference. Resistors R_9 and R_{10} form a voltage divider that "FLOATS" the AD536A common above ground, allowing single supply operation.

Capacitor C_4 is the averaging capacitor, while capacitors C_5 and C_6 are used for power supply bypassing. The dB output from the buffer amplifier (pin 6 AD536A) runs through meter calibration trimmer R_7 and resistor R_6 to a 50-0-50μA analog meter. Resistor R_8 provides a return path to ground.

To calibrate, first press the calibration switch and adjust R_5 to center the meter at zero with the chosen zero dB level applied. Decrease the input signal 30dB and adjust R_7 for −30dB.

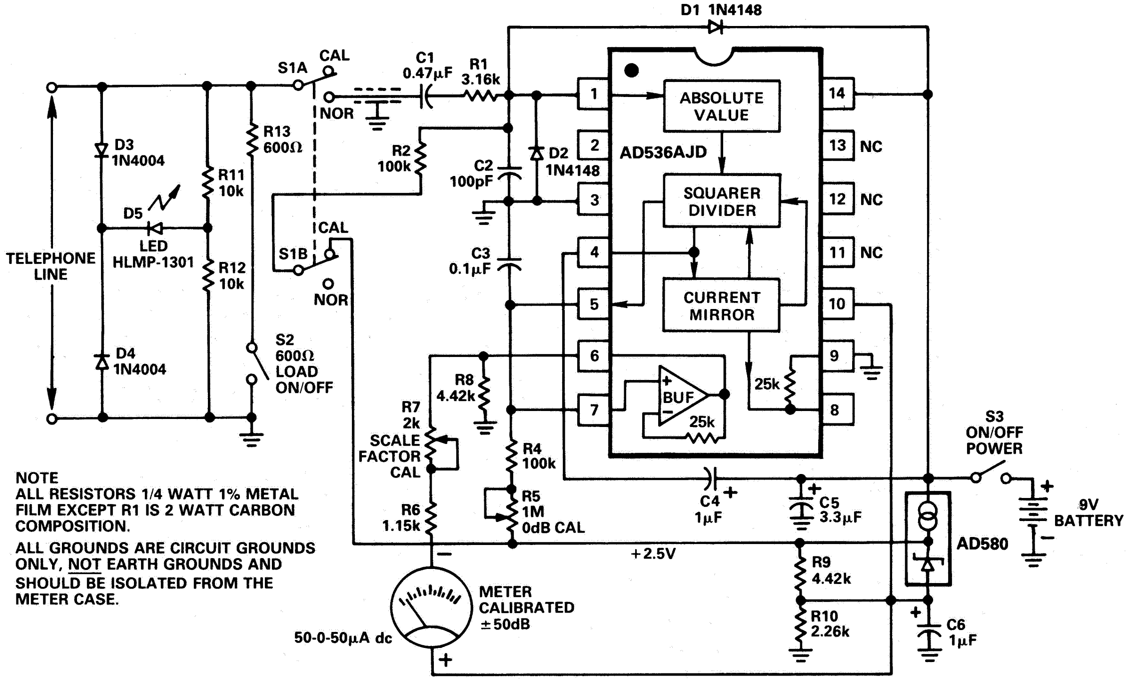


Figure 46. A Modem Line Monitor – A Telephone Line dB Meter

DATA ACQUISITION

A Programmable Gain rms Measurement System

Introduction

The rms measurement of complex waveforms of varying magnitude normally requires a high quality, compensated input attenuator. In contrast, the programmable gain rms preamplifier circuit of Figure 47 features an AD544 bifet operational amplifier as an inverting input buffer with four remotely switchable gain ranges: 200mV, 2 volts, 20 volts, and 200 volts full scale. Switching gain resistors in the buffers 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 500 volts peak.

Circuit Description

The input signal is connected to input jack J_1 with resistor R_1 and diodes D_1 and D_2 forming the amplifier's input circuit protection. The diodes will conduct whenever the voltage at pin 2 of the AD611 exceeds either of the power supply voltages by more than a diode V_{BE}. 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's 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 through C_7 are compensation capacitors which are adjusted for flat response at each gain setting. A_0, A_1, and A_2 are three address lines which select the desired input range of the preamplifier. R_4, R_6, R_{10}, and R_{12} are the gain calibration controls for each selected gain. The output of the AD611 operational amplifier is converted to its rms equivalent voltage by the AD536A rms-dc converter.

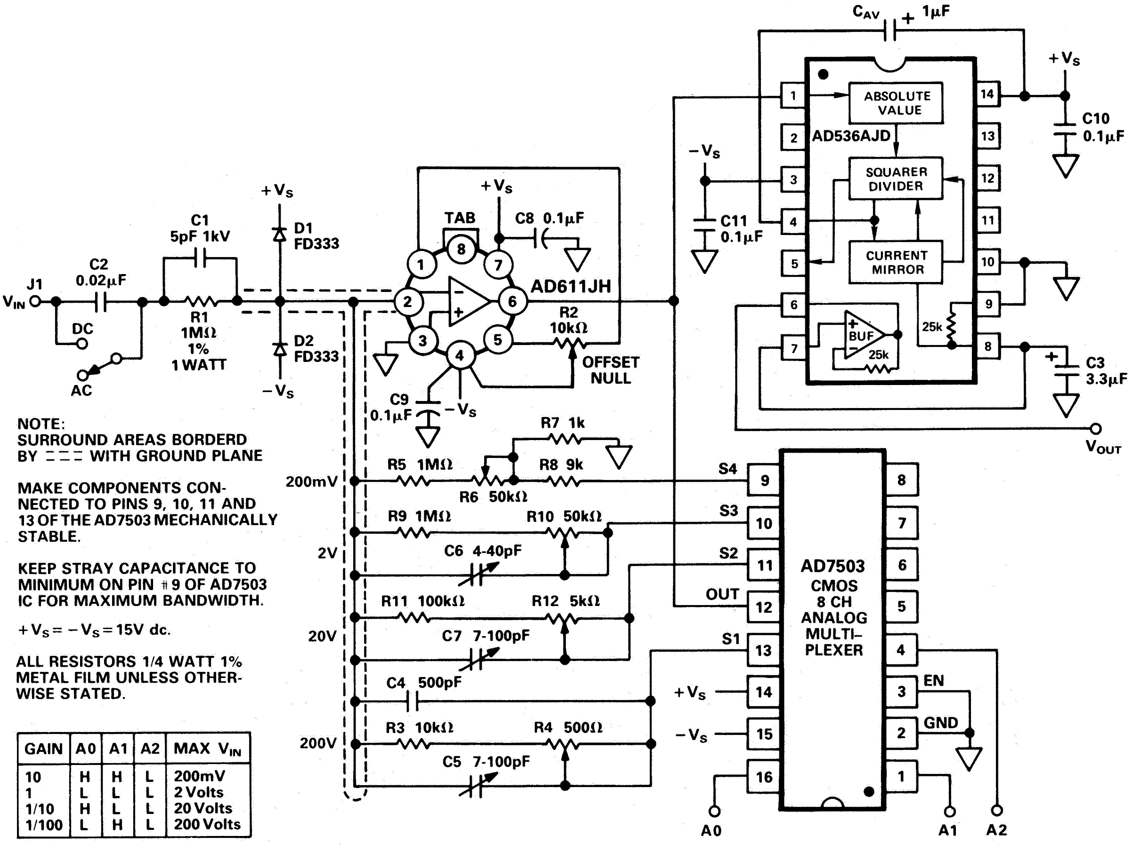
Performance Data:

Input Ranges: 200mV, 2 volts, 20 volts, 200 volts rms

– 3dB Bandwidth

200mV	≥4kHz*
2V	600kHz
20V	1.5MHz
200V	600kHz

*Bandwidth will vary with the degree of stray capacitance at pin 9 of the AD7503.



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

Figure 47. A Programmable Gain rms Measurement System

Noise referred to amplifier input: 360μV rms on 2 volt range, 75dB signal to noise ratio. RMS converter settling time: 397ms to within 1% of change in rms level of input. Power requirements: +5 volts dc @ 14mA ±15 volts dc @ 3mA.

Precautions

For maximum bandwidth and minimum input currents, capacitance at the summing junction (pin 2) of the AD544 must be kept to a minimum, and all points in the circuit connected to it should be Teflon insulated, or they should have a grounded guard ring surrounding them. The guard ring insures that leakage currents from the power supply pins or elsewhere are returned to ground and not to the summing junction.

As a safety precaution, the input jack and the wiring associated with it should be well insulated since potentially lethal voltages (200 volts rms) may be present.

Calibration

Address lines A_0, A_1, and A_2 should be set for each gain. The calibration trim potentiometers R_4, R_6, R_{10}, and R_{12} should be individually adjusted for the correct gain on each range.

The compensation capacitors C_5, C_6, and C_7 should be adjusted for flat response on each range by using a variable frequency sinewave input signal and either an oscilloscope to monitor the AD544 output, pin 6, or by using a digital voltmeter on its dc scale connected to the output of the rms converter.

Low Level rms Measurement Using an rms Instrumentation Amplifier

Introduction

The detection of low level signals can be made much easier and with greater accuracy by taking the required measurement differentially with an instrumentation amplifier (IA) rather than making an "unbalanced" measurement, employing an operational amplifier in one of the standard inverting or noninverting modes. To illustrate, an unbalanced input preamplifier, such as that shown conceptually in Figure 48, can be severely compromised in terms of input noise discrimination from several standpoints. Unless the input voltage V_{IN} is a completely floating source, it will be difficult to cleanly amplify the signal because of the common mode voltage, V_{CM}.

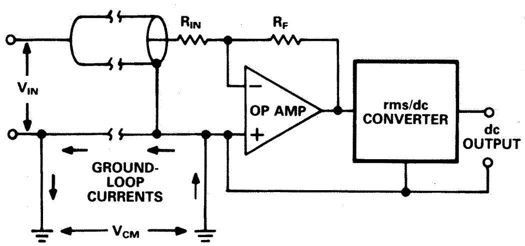


Figure 48. Noise in an Unbalanced System

In this system, both signal and ground loop currents flow in the shield line with the ground loop currents adding to the noise of the system. This added noise can make a low level measurement useless. In contrast to the operational amplifier, an instrumentation amplifier is a "gain block" which measures the *difference* between the voltages at its two inputs. This differential or balanced measurement method gives instrumentation amplifiers some distinct advantages, making them superior to standard operational amplifiers in many low level applications.

In the balanced measurement system of Figure 46, the shield line does not carry the input signal currents; it functions only as a shield. If there are any ground loop currents present, they simply are returned to common without adding to the noise of the system. Any noise that is picked up by the input lines

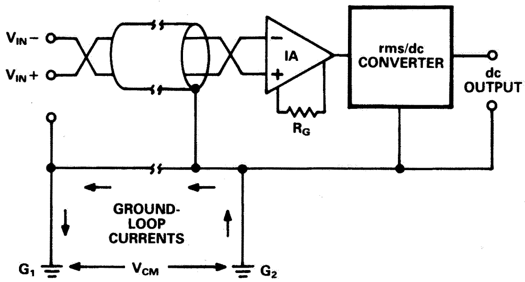


Figure 49. Noise in a Balanced System

will be "common mode" and cancelled out by the instrumentation amplifier. (V_{OUT} = $((+V_{IN}) - (-V_{IN})) \times$ GAIN).

Circuit Description

The rms-converter instrumentation amplifier scheme shown in Figure 50 uses an AD524 dual instrumentation amplifier.

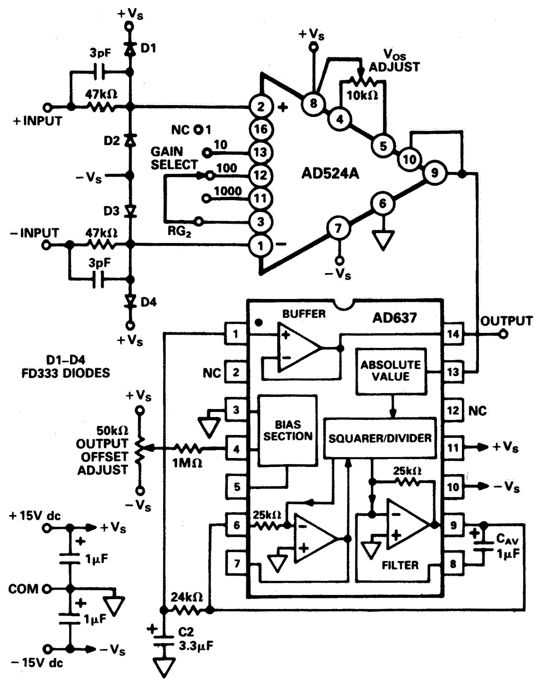


Figure 50. An rms Converter with an Instrumentation Amplifier Preamp

This, in conjunction with the AD637 rms-dc converter, forms a high quality system for low level rms measurement. The preamplifier section of this system offers superior overall performance featuring: excellent common mode rejection of up to 120dB, dc to 60Hz, and low input bias current.

This circuit features a very high input impedance of $10^9\Omega$; however, there must be a return path to the ground potential of the converter portion, which is to say the preamplifier section cannot operate with totally floating sources, such as transformers or thermocouples. For such instances, two high value bias resistors can be added (RB_1 and RB_2) each with a value of several megohms.

RMS NOISE MEASUREMENT

Introduction

Computing the root sum of squares is a well accepted method for measuring, evaluating, and comparing different types of noise signals. Practical noise measurement applications include the testing and grading of components–such as transistors, op amps, instrumentation amps and many other ICs.

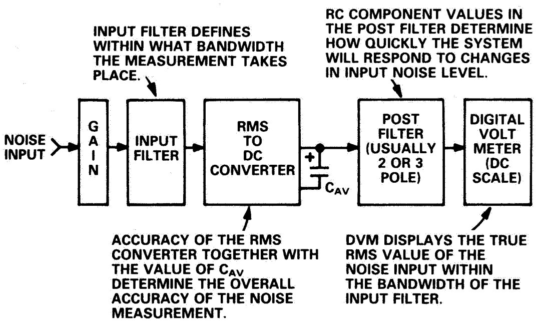


Figure 51. A Functional Breakdown of an rms Noise Measurement System

In general, a noise measurement system, such as that depicted in Figure 51, has the device under test–operating at some known noise gain being followed by an input filter whose output drives an rms converter. The input filter defines the bandwidth within which the measurement is to be made. A two or even a three pole active filter is normally used for the low pass section of the input filter, because these filters provide greater effective attenuation of out of band signals (such as harmonics). Noise, therefore, needs to be specified within the specific bandwidth of the filter; even more importantly, the area under the curve of noise amplitude vs. frequency needs to be known. Both the exact equations and a close cookbook approach for determining this area are detailed later on in this section.

The Effects of Input Coupling on Input Filter Performance

As shown in Figure 52, the input filter may have either dc or ac input coupling; the type of input coupling that is used will determine the response of the filter at low frequencies. DC input coupling allows the filter to respond to dc inputs but also permits high computational errors at very low frequencies, because the value of C_{AV} cannot be infinite. AC coupling produces a low frequency roll-off starting at frequency point F1. If C_{AV} is adequately large, this connection will maintain low error throughout the passband of the filter. There is, of course, no response to dc inputs when ac coupling is used.

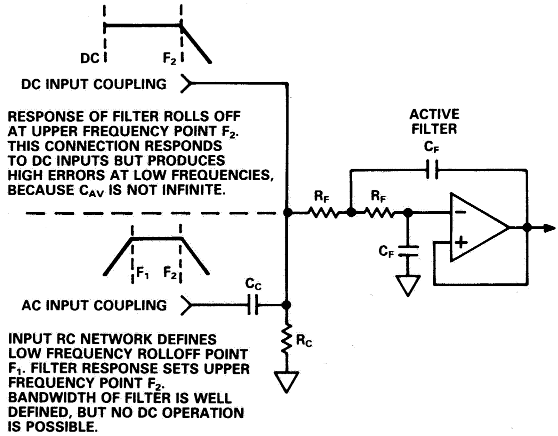


Figure 52. The Effects of Input Coupling on the Overall Response of an Active Input Filter

Determining the Noise Gain of the Input Filter

There are two important characteristics of the input filter's overall response that must be known in order to determine its noise gain. Once the noise gain is known, the noise of the device under test is easily found by dividing the (filtered) noise output of the rms converter by the noise gain of the circuit. The first characteristic which must be determined is the value of noise voltage vs. frequency (Nmax (f)). Typically, this peak will occur close to the center frequency, F_O, of the filter. Next, the noise equivalent bandwidth needs to be found.

A Cookbook Procedure

When the -3dB corner frequency F2 is more than 100 times higher in frequency than point F1, the following cookbook procedure may be used to closely approximate (to better than 1%) both the noise effective bandwidth of the input filter and its noise voltage vs. frequency, Nmax (f). This procedure applies to critically damped 2 and 3 pole input filters such as in Figure 53.

Simply multiply F2 by the following "correction factors" to obtain the noise effective bandwidth: for a one pole filter, multiply frequency F2 by 1.57; for a two pole, by 1.22 and for a three pole filter, multiply by 1.16. A correction factor is not required for F1, since, in this case, frequency F1 is significantly less than that of F2; F1 is, therefore, ignored. F2 times the correction factor closely equals the noise effective bandwidth of the input filter.

The Nmax (f) of an input filter with a 100 to 1 F2/F1 ratio can also be closely approximated by assuming a gain of one at the center frequency of the filter. This cookbook approach is acceptable, because, at the center frequency, the filter closely approximates a unity gain voltage follower. Over a 100 to 1 frequency ratio, any errors in this approximation will be minimal.

Once both the maximum gain of the filter and its noise effective bandwidth are known, the "noise gain" of the filter can be calculated by:

$$G(n) = \text{Nmax}(f) \times \sqrt{(\text{NEBW})}$$

Where G(n) is the noise gain of the input filter Nmax (f) is the maximum value of N(f) over the filter passband and NEBW is the noise equivalent bandwidth of the filter in volts per root Hz.

Determining the Exact Noise Gain

To find the *EXACT* noise gain, both the value of noise voltage vs. frequency, Nmax (f) and the noise equivalent bandwidth of the filter must be known to close tolerances. Although typically, Nmax (f) will occur close to the center frequency, F_O, of the filter, the exact gain at the peak response frequency of the filter needs to be measured. This measurement can be taken by utilizing a spectrum analyzer or by using an audio oscillator and an accurate ac voltmeter.

Next, the exact noise equivalent bandwidth needs to be determined; it may be calculated by using:

$$\text{NEBW} = \int_0^{\infty} \left[\frac{N(f)}{\text{Nmax}(f)}\right]^2 df$$

Where N(f) is the average output noise as a function of frequency. Nmax(f) is the maximum value of N(f) over the passband. Practical limits in the above equation are the points where N(f) is 40dB below Nmax(f). The −3dB bandwidth of the filter and its noise equivalent bandwidth become more equal as the filter's response approaches that of a "brick wall" filter (i.e., an ideal filter with infinitely straight–and fast–rising and falling edges).

Processing Noise with an rms Converter

Selecting the Value of C_{AV}

The buffered output of the input filter is connected to an rms to dc converter which computes the root-mean-square of the filtered noise fed to its input. The converter, by use of a very long-time constant, averages out short duration transients in the measurement. The input filter needs to have a low impedance buffered output (such as an op-amp) to allow it to drive the 6kΩ to 16kΩ input impedance of the rms converter.

It should be pointed out that an rms conveter is most useful for wide bandwidth noise measurement. Spot noise, where frequencies F1 and F2 are very close together in frequency, contains no significant harmonic content. Therefore, this type of noise can be just as accurately measured using an MAD rectifier (see Section I).

If an rms converter *IS* used for noise measurement, the value of its averaging capacitor is of critical importance. The C_{AV} must be sufficiently large to adequately process the lowest frequencies of interest; therefore, frequency point F1 normally determines the value of this capacitor (refer to filters and averaging section). The 5% and 1% error lines on the two pole filter chart in the filters and averaging section should be used to obtain practical values of C_{AV} and (post filter capacitors) for frequency F1.

The 5% error line will select a value of C_{AV} that will allow quick settling but still provide enough filtering to prevent the dc error from having any more than a minimal effect on accuracy. As an example: if the lower corner frequency point F1 were chosen as 10Hz, the recommended C_{AV} for fast settling would be 0.16μF with a 1.3μF C_2 and C_3 value in the post filter.

If very high accuracy is required, the 1% error curve will provide more than enough filtering; in fact, the predominate circuit error will now be due to the non-instantaneous response of the input filter. In other words, the one pole input filter has a gradual response of 6dB per octive and, therefore, it will exhibit gradually decreasing error vs. frequency near the corner frequency, F1.

A Practical Noise Measurement Circuit

Figure 53 is an example of an rms noise measurement system for audio frequencies, with practical component values for a −3dB bandwidth of 10.6Hz to 20.5kHz. Note that the high pass input coupling network C_C, R_C may be omitted and replaced with a single capacitor between the active filter output and the input of the rms converter. Moving this capacitor provides two advantages: first, the input offset of the op amp used for the input filter is no longer a problem. Also, the 15kΩ R_C is now unnecessary; the input resistance of the rms converter now serves this function.

On minor disadvantage of moving C_C is ‿at the

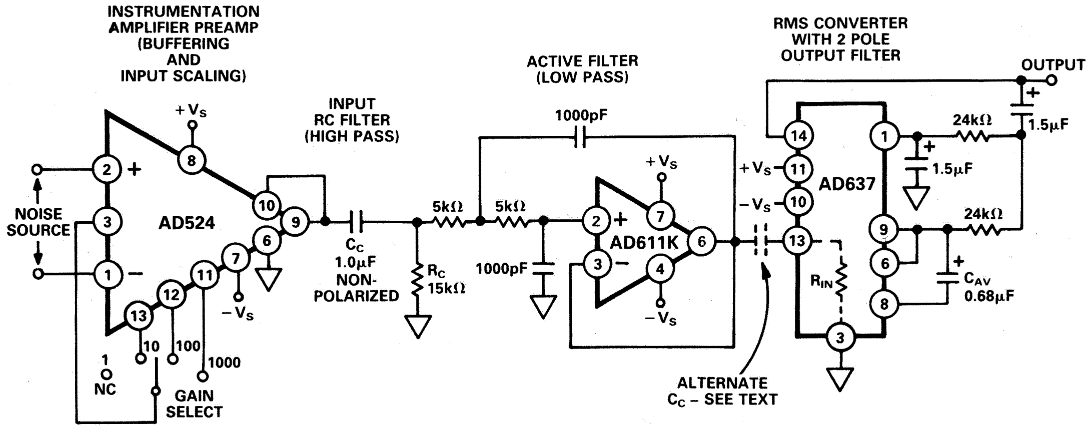


Figure 53. A Practical Audio Noise Measurement Circuit

input impedance of the rms converter will vary plus or minus twenty percent (due to thin-film process variations). Therefore, for a close bandwidth tolerance, the input resistance of the converter needs to be measured with a DVM and the value of coupling capacitor should then be changed appropriately (i.e.: 10% less resistance requires 10% greater capacitance, etc.).

The noise voltage vs. frequency, Nmax (f), of the filter illustrated in Figure 53 is approximately 1.0 at center frequency. The noise equivalent bandwidth of this circuit is approximately 20.51kHz times 1.22 or 25.0kHz. The cookbook value for noise gain is then 1.0 times the square root of the noise equivalent bandwidth: this equals 158.2. Multiply the desired gain of the instrumentation amplifier preamp (1, 10, 100, 1000) by 158.2 to arrive at the total circuit gain. The output reading is then divided by this number to determine the noise of the device under test in volts.

For preamp gains of 1 to 100, the bandwidth of the AD524 instrumentation amplifier is much greater than the 20kHz F2 corner frequency of the filter; therefore, the amplifier's high frequency roll-off has no effect on the overall accuracy of the measurement. However, at a gain of 1000, the bandwidth of the AD524 drops to 25kHz. This results in a reduction, of a few percent, in noise equivalent bandwidth for the pre-amp/filter section. But, this reduction in bandwidth is insignificant if tantalum capacitors and other wide tolerance filter components are used.

If the very best frequency stability and accuracy is needed, the resistors used in the input and post filter should be 1% metal film types. Their associated capacitors should be polypropylene or a similar low leakage variety with a close capacitance tolerance (or they can be hand-selected to be within 1% of the required value).

LOW FREQUENCY RMS MEASUREMENT

Introduction

As described in the previous sections, reducing the input frequency requires lengthening the averaging (and filtering) time contants to maintain the same levels of dc error and ripple. Consequently, successively larger values of C_{AV} are required as the input frequency is reduced. With the very large values of averaging capacitor necessary at frequencies below 10Hz, the physical size of the C_{AV} can occupy excessive board space and prohibit the use of the high quality, low leakage types that are the most useful at these frequencies.

Although the rms converter output filter section (or sections) can easily have their series resistance increased to give a longer averaging time constant (such as increasing Rx in Figure 27), the AD536A and AD636 averaging sections are not so flexible. Their fixed 25kΩ internal averaging resistors cannot be increased (see Figure 35). In these converters, averaging is carried out within the current mirror. The current, I_4, is averaged by C_{AV} and then ratioed (2×) to the output, via the current mirror, to the I_{OUT} terminal.

Fortunately, with the AD637 rms converter, averaging takes place within a filter stage which is externally accessible, shown in Figure 19. By reducing C_{AV} to 100pF (just enough capacitance to maintain stability), the filter stage becomes an output buffer, allowing external averaging. With this connection, very large resistance values (and therefore much smaller averaging capacitors) may be used. The circuits of Figures 54 and 55 use this concept to produce averaging times of several seconds, yet require relatively small averaging capacitor values.

A Low Frequency rms-dc Converter Circuit

Figure 54 is a low frequency rms to dc converter circuit optimized to give less than 0.1% averaging error for frequencies down to 1Hz. With this circuit, averaging is carried out *between* the rms output terminal, pin 9, and the input to the internal buffer amplifier, pin 1. The buffer amplifier successfully isolates the 25kΩ input impedance of the denominator input pin from the averaging section, preventing that section from being loaded down; it also provides the output buffering necessary to drive external circuitry. Rather than being directly connected to pin 9, the denominator input is now tied to pin 14, receiving its feedback via the output of the buffer amplifier. The rms output may be taken from either the buffer output pin for a 4.8% averaging error output (thus giving a high error output with minimum settling time) or from the output of the external output filter whose filtering reduces the averaging error to less than 0.1% (at the expense of increased settling time).

A useful by-product of utilizing the external component averaging scheme just mentioned is that by making the AD637 filter section a simple voltage follower, a new output function, V_{IN}^2/V_{rms}, now becomes available.

An Ultra-Low Frequency rms-dc Converter Circuit

The circuit of Figure 55 operates in a similar manner to that of Figure 54 except that it uses two very low input bias current amplifiers which permit even larger values of averaging resistance (in this case 10MΩ) to be used.

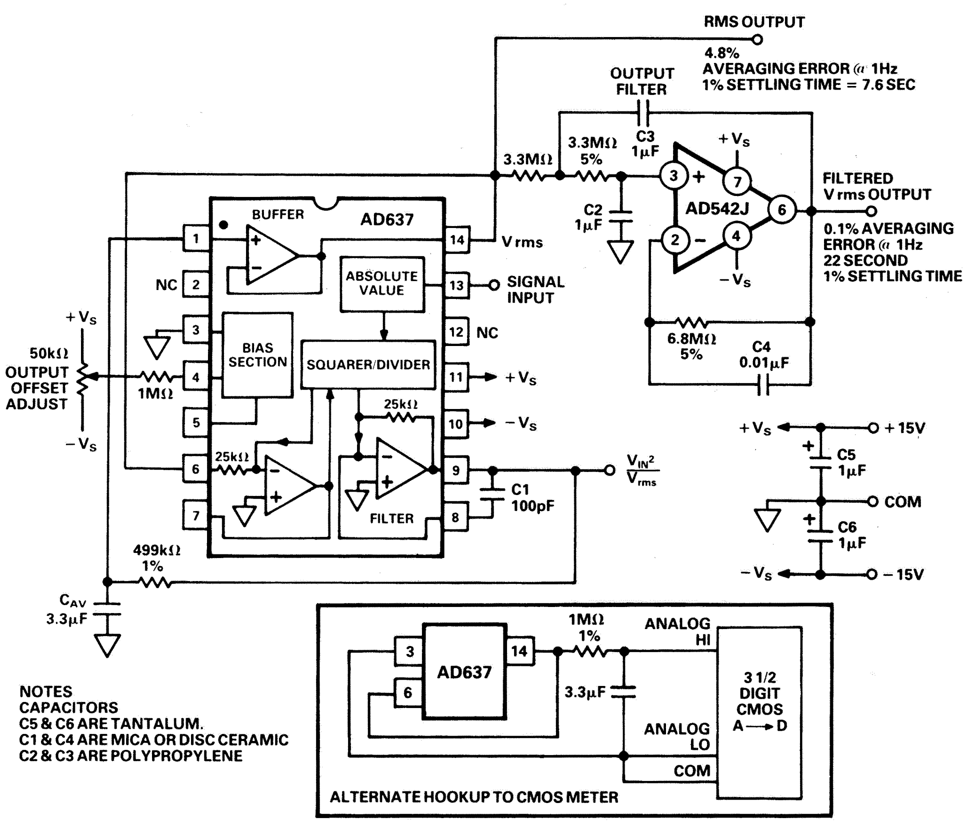
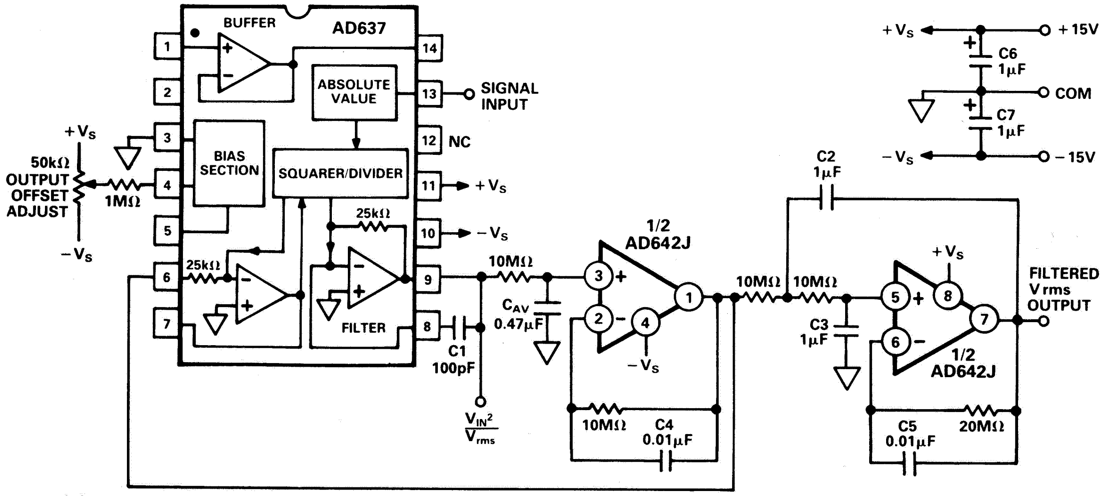


Figure 54. A Low Frequency rms to dc Converter Circuit



NOTE:
VALUES CHOSEN TO GIVE 0.1% AVERAGING ERROR @ 0.1Hz
WITH A 67 SECOND 1% SETTLING TIME

ALL RESISTORS 1/4 WATT 5% CARBON COMPOSITION.
CAPACITORS: C_6 & C_7 ARE TANTALUM.
C1, C4, C5 ARE DISC CERAMIC OR MICA – CAPACITORS
CAV, C2, and C3 ARE POLYPROPYLENE.

Figure 55. An Ultra-Low Frequency rms to dc Converter Circuit

This circuit has been optimized to exhibit less than 0.1% averaging error for input signals as low as 0.1Hz. As with the previous circuit, the V_{IN}^2/V_{rms} function appears at pin 9 of the AD637.

Note:

The two low frequency rms measurement circuits described in this section may overload on transient noise spikes, such as those at power line frequencies. This occurs because the filter stage averaging capacitor (normally called C_{AV}, but in these circuits, renamed C_1) has been drastically reduced. This allows the output at pin 9 of the AD637 to respond to the ***square*** of the input signal rather than to the ***average of the square*** of the input. For example, if a 1 volt peak transient appears at the input of the rms converter while the circuit is measuring a 10mV rms input signal, the output at pin 9 should theoretically equal:

$$V_{OUT} = \frac{V_{IN}^2}{V_{rms}} = \frac{(1\ \text{volt})^2}{0.01\ \text{volts}} = 100\ \text{volts!}$$

Obviously, the output will saturate long before it approaches 100 volts, creating a large error which ***may not be noticed*** as such at the filtered V_{rms} output point due to the extensive RC filtering between this point and pin 9 of the rms converter. Therefore, for general purpose applications where the V_{IN}^2/V_{rms} function is not needed or for applications where high crest factor-low frequency signals are to be measured, it is recommended that capacitor C_1 be increased to 3.3µF. This capacitor, in conjunction with the internal 25kΩ filtering resistor, will form a low pass filter with a 2Hz corner frequency. This will attenuate higher frequency signals, i.e., 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. Therefore, 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 increasing C_1 causes the 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 instanteous square of the input signal from appearing at pin 9 of the AD637.

MICROPROCESSOR CONTROLLED FUNCTIONS OF THE AD637

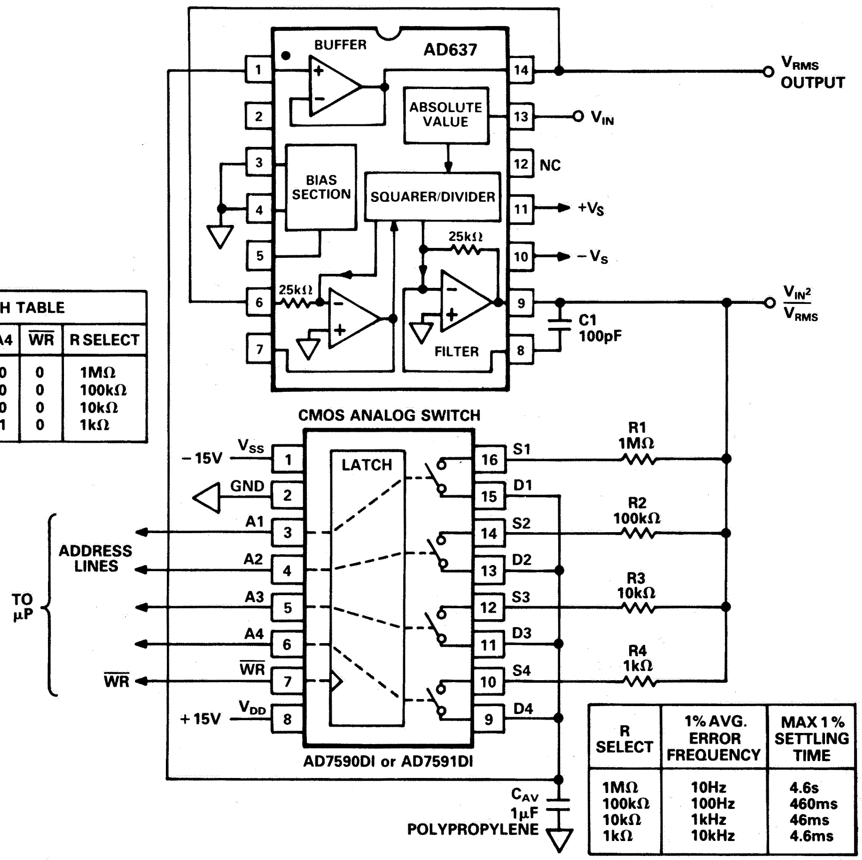
An rms Converter Circuit with a Microprocessor Controlled Averaging/Settling Time Constant

The circuit of Figure 56 gives a good indication of the power and versatility of the AD637's separate denominator input feature. This circuit allows a microprocessor to automatically change the averaging/settling time constant of an rms measurement system from a remote location (or at very high speed).

In this configuration, a very small value capacitor is connected between pins 8 and 9 of the AD637 to insure circuit stability. The actual averaging of the rms output does not occur here; instead, it is carried out within an RC filter network between the output of the AD637, pin 8, and its denominator input, pin 6. The internal buffer amplifier is used to isolate the external RC filter from the 25kΩ input impedance of the denominator input. A microprocessor addresses a CMOS analog switch which selects the appropriate series "R" to give the desired RC time constant.

For low frequency signals, such as 10Hz, a 4.6 second 1% settling time may be chosen to given the rms converter the sufficiently long time constant necessary for low dc error. However, at higher frequencies, (or for inputs which have most of their energy at higher frequencies such as 10kHz), the μP can quickly and automatically decrease this settling time to 4.6ms and get the 10kHz reading with a minimum of delay–instead of waiting 4.6 seconds!

Although values of RC time constant between 1ms and 1 second were chosen to give 1% settling times of 4.6ms to 4.6 seconds respectively (that is: 1ms × 4.6 time constants = 4.6ms, etc.,) in decade increments, other time constants (and increments such as logarithmic) may be selected to suit the particular application. The only restriction of this circuit is the (typically) 2nA input bias current of the AD637's internal buffer amplifier; this limits the maximum re-



TRUTH TABLE

A1	A2	A3	A4	$\overline{WR}$	R SELECT
1	0	0	0	0	1MΩ
0	1	0	0	0	100kΩ
0	0	1	0	0	10kΩ
0	0	0	1	0	1kΩ

R SELECT	1% AVG. ERROR FREQUENCY	MAX 1% SETTLING TIME
1MΩ	10Hz	4.6s
100kΩ	100Hz	460ms
10kΩ	1kHz	46ms
1kΩ	10kHz	4.6ms

Figure 56. A μP-Controlled Averaging/Settling Time rms Converter Circuit

sistance which can be connected in series with its input to less than 1 meg ohm (this is to keep the output offset of the buffer amplifier below 1mV). Therefore, for longer RC time constants increase the value of C_{AV}; for shorter time constants, replace resistors R_1 to R_4 with smaller values.

Note: As with the circuits in the Low Frequency Measurement section, this circuit may overload on transient noise spikes! Please refer to the ultra-low frequency rms-dc Converter circuit section for further details.

Quick-Reset an rms Converter with a VMOS FET

The quick reset rms scheme described here uses a VMOS power FET to rapidly discharge the energy stored in the averaging capacitor of an rms converter, thus permitting the converter's output to be rapidly reset after a measurement has been taken. This method of resetting the averaging capacitor is particulary useful when performing low frequency/long time constant measurements. The circuit is gated with standard TTL input levels.

This quick-reset circuit can be combined with the μP-controlled averaging/settling time circuit of Figure 56 to provide an automated measurement system capable of sampling, measuring, resetting and sampling again–with all these functions under μP control.

CIRCUIT DESCRIPTION

In addition to the AD637 rms converter IC, the quick reset rms circuit shown in Figure 57a also uses a VN46AF VMOS power FET and an MC14050B CMOS buffer. The input signal is applied to pin 13 of the AD637; the averaging capacitor, C_{AV}, is connected between pin 8 and the rms output, pin 9. The VN46AF has its source and drain connected directly across the averaging capacitor. With its gate at ground potential, this enhancement-mode FET has a very high source to drain resistance. In contrast, with its gate "high" the VMOS FET exhibits a resistance of approximately 8Ω, which will rapidly discharge C_{AV}.

A CMOS buffer/driver was chosen for this circuit because it allows the highest possible driving voltage for the VMOS FET gate (thus decreasing its "on" resistance) using a standard 5 volt logic power supply. In addition, the CMOS buffer has a very low standby current. The 1kΩ resistor connected between the input of the buffer/driver and ground insures that the

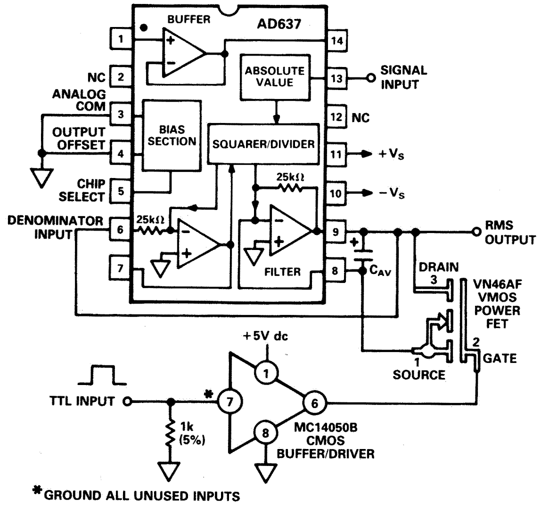


Figure 57a. A Quick-Reset rms to dc Converter Circuit Especially Suited for Low Frequency Measurements

CMOS buffer will stay in the "low" state unless it receives a TTL "high" level input.

Figure 57b is an oscilloscope photo displaying the output of the rms converter over time. At midscreen, the logic level at the CMOS buffer input has gone high, causing the VMOS FET to rapidly discharge C_{AV}. This photo was taken using a 10μF C_{AV}. In this case, the discharge time from 1 volt rms down to 10mV (1%) is approximately 120μs and will vary directly with the value of C_{AV} (however, the CHARGING time, requiring around 2.3 time constants, is approximately 450ms!)

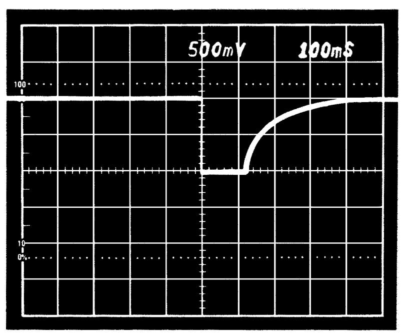


Figure 57b. Oscilliscope Photo Showing the rms Output Over Time Using a 10μF C_{AV}

A few words of caution: the effective series resistance of the averaging capacitor, Esr, can introduce an error which causes the exact settling time of the circuit to deviate from the ideal linear relationship of discharge time vs. C_{AV}. Also, avoid using large "leaky" electrolytic capacitors for C_{AV} since their high dielectric absorption and loose tolerance values may contribute to additional errors.

A final note: The C_{AV} discharge time may be further reduced (to better than one third of the remaining time) by driving the gate of the VMOS FET with a voltage higher than +5V dc. One way to accomplish this is to increase the power supply voltage powering the CMOS buffer/driver while *simultaneously* increasing the TTL "high" level driving the buffer.

A Microprocessor-Controlled Two Quadrant Analog Squarer

A second variation (on the same theme) of the variable time constant rms converter is the μP controlled variable gain squarer circuit of Figure 58. This circuit provides an excellent performance/price value by using an rms converter as a precision squarer–instead of using a high cost analog multiplier! It also allows both the square and the mean square of the input signal to be accurately computed.

In this circuit, the filter stage averaging capacitor, C_{AV}, has been reduced to the absolute minimum value which still maintains stability since stability NOT averaging is C_{AV}'s function here. The denominator input of the AD637, pin 6, normally connected directly to the output terminal, pin 9, is now used as a gain control terminal. The rms converter now responds to the *instantaneous square* of the input signal rather than to the *average of the square*, as in the normal rms connection. The output voltage at pin 9 of the AD637 equals:

$$V_{OUT} = \frac{V_{IN}^2}{V_{denominator}}$$

Since the output level of the rms converter is now dependent on the voltage applied to the denominator input, a microprocessor can subsequently be used to vary the output level of the squarer. Consisting of an AD581 voltage reference, an AD7590 CMOS analog

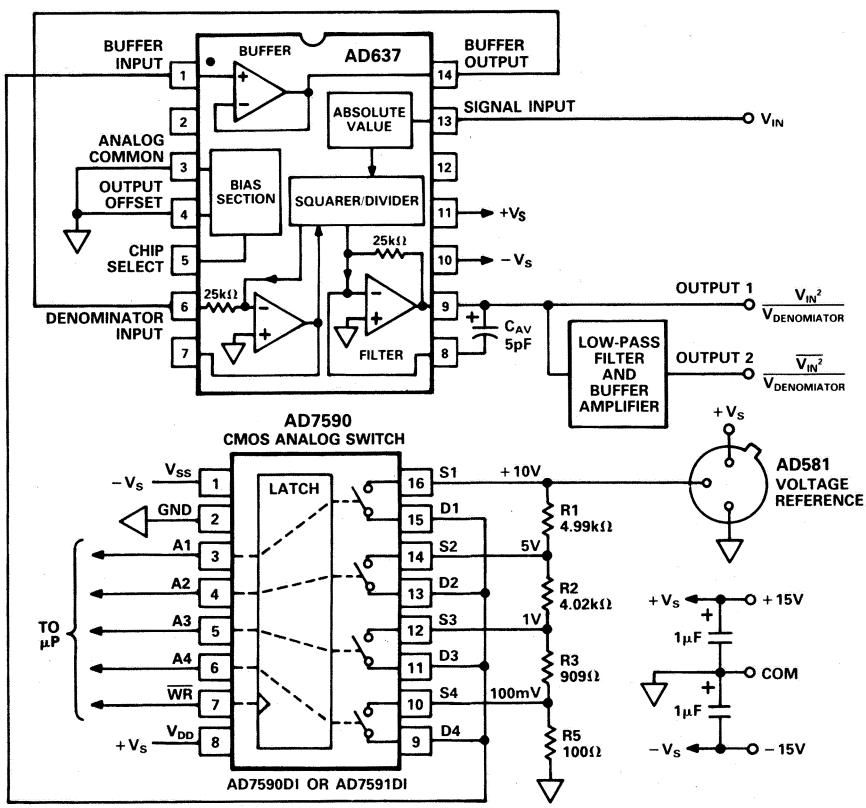


Figure 58. A μP-Controlled Two Quadrant Analog Squarer

switch, and a resistive voltage divider, the µP-controlled voltage source used in this circuit provides automatic and remote gain control via several digitally-selectable denominator voltages.

The dc performance of this circuit is excellent, with accuracies typically within 0.2% or better, in percent of reading, for signal input levels between 100mV and 10 volts and denominator inputs of 10V, 5V, 1V, and 100mV. These numbers do not include denominator voltage errors which may be caused by non-ideal resistor dividers.

Please note: since the instantaneous square function requires that C_{AV} be reduced to the minimum value which still maintains stability (i.e., NO averaging being carried out) transient input spikes may easily exceed the + 12 volt swing of the AD637 filter amplifier! For this reason, optimum performance will be set by a tradeoff between the best overload sensitivity and the maximum bandwidth of this circuit. That is: as the value of C_{AV} increases, better overload protection is provided, but bandwidth suffers as a result of the high frequency rolloff due to the RC time constant created by C_{AV} and the AD637's internal 25kΩ filter stage feedback resistor.

As shown by Figure 58, a low pass filter (such as an AD741 in an active filter configuration) may be added to obtain an output proportional to the AVERAGE of the square of the input voltage, thus providing both a squared and mean-squared output function. Alternatively, increasing C_{AV} is another way to produce an output that responds to the average of the square. This has the additional benefit of greatly improving the circuit's transient overload characteristics! Note, however, that now the output from pin 9 of the AD637 will be a filtered dc voltage equal to the average of the square of the input voltage (rather than an ac absolute value waveform).

MISCELLANEOUS MATHEMATICAL COMPUTATIONS

Vector Summation

The denominator input of the AD637 permits this chip to be used as a cascadeable n-dimensional vector sum, or root sum of squares, building block. Figure 59 shows a two dimensional version of this summation circuit; each additional AD637 would add one extra variable. In this circuit, the V_X variable is applied to the input of IC_1; likewise, the V_Y input is applied to IC_2. Assuming a constant voltage denominator level at pin 6, IC_1 produces an output voltage of $\frac{V_X{}^2}{V_D}$. This voltage is then inverted by the unity gain summation amplifier IC_3 and applied through scaling resistors, R_4 and R_5, to the filter section summing junction of IC_2. Potentiometer R_4 is adjusted to produce an exact unity gain output from A_1 to the output of IC_2; this output voltage also appears at IC_2's output as $V_O = \frac{+V_X{}^2}{V_D}$.

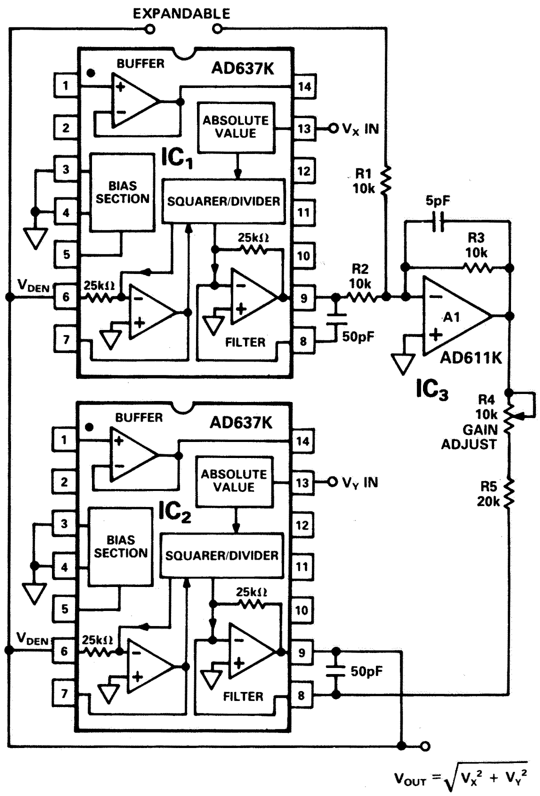


Figure 59. A Vector Summation Circuit

In this circuit, the input to IC_2, V_Y, also appears at IC_2's output as V_{Y2}/V_D. The total output equals:

$$V_O = (V_X{}^2/V_D) + (V_Y{}^2/V_D)$$

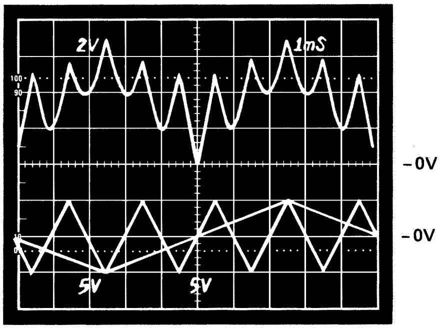
Since the denominator inputs are tied in parallel back to V_O, the equation can also be considered as:

$$V_O = (V_X{}^2/V_O) + (V_Y{}^2/V_O)$$

Multiplying by V_O: $V_O{}^2 = V_X{}^2 + V_Y{}^2$

Solving for V_O: $V_O = \sqrt{V_X{}^2 + V_Y{}^2}$

This implicit or feedback solution gives greater dynamic range and accuracy than a straightforward or explicit approach. The explicit solution would normally require the use of fixed gain squarers for each variable as well as a square rooter to process the root of squares. The dynamic range of this circuit is 10V to 10mV (60dB), limited only by the 0.5mV input offset voltage of the AD637K. As one of the inputs passes through zero, the output will exhibit an error due to this offset term. This 60dB range is about thirty times greater than the 10V to 0.3V (30dB) range of a comparable performance explicit summation circuit.



TWO OR MORE AD637S CAN BE USED TO CALCULATE ROOT SUMS SUCH AS $\sqrt{V_X^2+V_Y^2 \ldots V_{IN}}$ BY IMPLICIT ANALOG COMPUTATION.

Figure 60. The Upper Trace of the Scope Photo is the Root Sum of the Two Triangular Waves That Are Shown in the Lower Trace

The upper waveform in Figure 60 shows the vector sum output (2V vertical scale) of two triangular waves of 100Hz and 500Hz (5V vertical scale). As the input goes through zero, the output appears as a sharp "V" with its point on zero. That is:

$$V_O = \sqrt{(0^2) + (V_y^2)} = 1V_Y$$

This increase in magnitude of the 100Hz triangle wave from this point results in a parabolic waveform. The reason for this is that when $V_X > V_Y$ in the equation:

$$V_O = \sqrt{(V_X^2) + (V_Y^2)}$$

the equation can then be approximated as:

$$V_O = V_X + [(1/2\, V_Y^2)/V_X^2]$$

The useful bandwidth of the circuit is about 100kHz for a 1000:1 input dynamic range. Each of the inputs can go as high as ±15V as long as the instantaneous vector sum is below the 13V clipping level of the amplifiers.

The two dimensional circuit of Figure 59 is expandable to accommodate four or more input variables by first tying together the input pins of additional AD637s and then by summing each of their output pins through 10kΩ resistors to the summing junction of IC_1.

POWER MEASUREMENT

Introduction

In any circuit, the product of the voltage across and the current through a given load will equal the power dissipated through the load. Problems arise (and much confusion persists) when the load is not a pure resistance. Inductive and/or capacitive circuit components introduce a phase shift between the voltage across and the current through a given load. This phase shift becomes more pronounced as the reactance of these components increase (as with increasing frequency); they then become a greater portion of the total impedance of the load. In a purely inductive circuit, the voltage will lead the current by 90 degrees; conversely, the voltage lags the current in a purely capacitive circuit.

Of Volt-Amperes, Watts, and Vars.

There are three primary methods for defining and measuring the (sine-wave) power dissipated in a given load impedance: apparent power (Pa), average power (P) and reactive power (Pr).

Apparent power, measured in volt-amperes, is simply the product of the rms value of the voltage across a given load times the rms value of the current through the load. That is:

$$Pa = V_{rms} \times I_{rms}$$

Where V is in volts and I is in amperes. The volt-ampere rating is often used in specifying electrical equipment since volt-amperes may be used to directly compute the current requirements of individual pieces of equipment.

Average or real power, measured in watts, is equivalent to the apparent power multiplied by the cosine of the phase angle separating the voltage and current waveforms. That is:

$$P = Pa \cos\theta = V_{rms} I_{rms} \cos\theta$$

Where V is in volts, I is in amperes, θ is in degrees. Most commonly used, average power specifies the overall power consumption of a particular circuit, regardless of the dissipation of its individual components–some of which may be reactive. The cosine of the phase angle, θ, is also referred to as the power factor and is the ratio of a circuit's average power to its apparent power. A highly reactive load exhibits a low power factor with a correspondingly low power consumption.

Because of the importance of defining power consumption within individual reactive components in a circuit, a third power specification, reactive power was created. Reactive power, in vars. (volt-amp reactive), is used to directly measure the peak power consumption of individual inductive components in a circuit, even though their average power consumption (ideally) is zero. Reactive power is very important to electrical power companies since they must still supply this energy during a portion of every cycle, even though (on the average) no energy is actually dissipated.

Reactive power is equal to:

$$Pr = Pa \sin\theta = V_{rms} I_{rms} \sin\theta$$

Where V is in volts, I is in amperes, θ is in degrees.

Practical Power Measurement

The fact that averaging is carried out in performing rms computation means that whatever phase information existed in the original signal will be lost after rms computation; this fact precludes the use of rms converters for measuring power into nonresistive loads. Measurement of complex power is normally carried out using analog multipliers, since they will preserve the voltage/current phase information.

Figure 61 shows the basic building blocks for a practical power measurement system which can accurately measure both real and reactive power. As shown by the figure, rms converters are used for real-time monitoring of the rms value of the voltage and current waveforms being processed. With their dc outputs, the converters can directly drive both analog panel meters or DVM chips.

As shown by the figure, the output of the analog multiplier is VI cos θ; at this point, the unfiltered multiplier output equals the instantaneous power dissipation through the load. As shown, if the output is low pass filtered, it will then equal the average or real power dissipated. Likewise, if only the negative half cycle of the output waveform is detected and filtered, this output will respond to the reactive power dissipated in the load.

Figure 62 is a block diagram of a practical power measurement circuit which measures apparent power by calculation $\frac{V^2}{R}$. A voltage sensor measures the voltage across a RESISTIVE load; the AD637 rms converter then squares this voltage; this squared output is then scaled by the denominator input voltage at pin 6. The denominator voltage must be set to give the required output voltage scaling for each particular load resistance. Since $\frac{V^2}{R}$ varies with the value of R, the circuit must be recalibrated each time the value of load resistance is changed. One volt per milliwatt or one volt per watt would be practical scale factors for this circuit. Since a squaring operation is being performed by the AD637, the scaling voltage

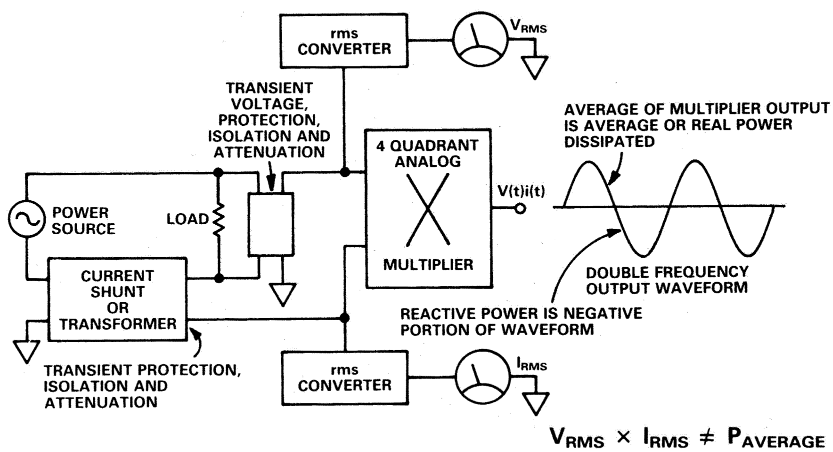


Figure 61. A Block Diagram of a Practical Power Measurement System

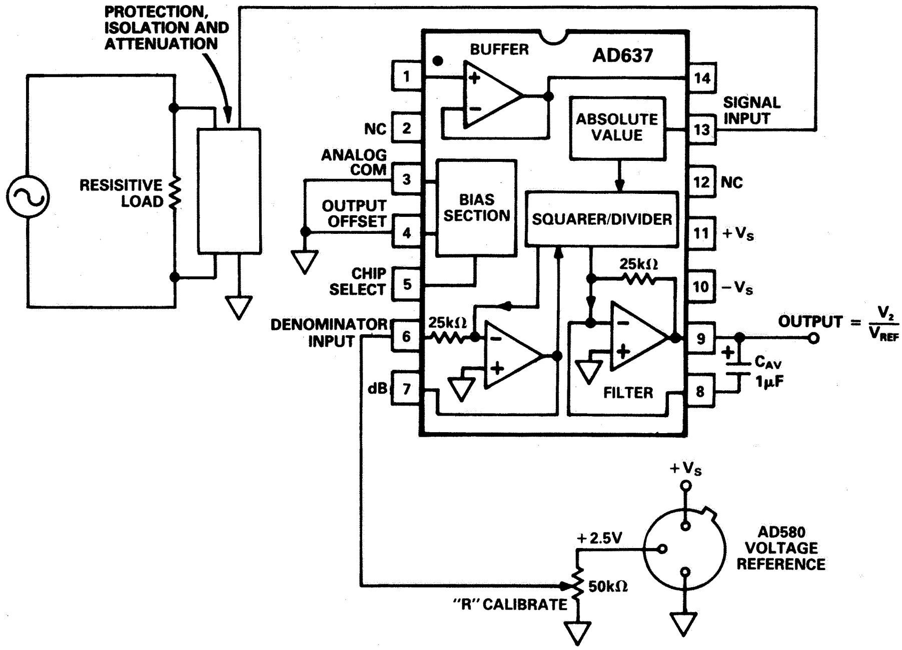


Figure 62. Apparent Power Measurement Using an rms Converter

must be carefully chosen to provide sufficient headroom to allow the rms converter to process the maximum full-scale input level without clipping. Thus, there will be a tradeoff between maximum input level and low level sensitivity.