APPLICATIONS

INSTRUMENTATION FUNCTIONS

WATTMETERS

Multipliers are well suited for wattmeter designs. Figure 7 shows a simple arrangement that measures the power output of an audio amplifier into a load. The 18kΩ-10kΩ divider scales the amplifier's output voltage swing to a maximum of 10V (from a maximum of 28V, representing about 100W peak power), well within the AD534's input range. The voltage is measured across the load, with the tap of the divider connected to X_1 and the lower end of the load to X_2. The power dissipated in the divider is negligible (1/3500 of the power in the load).

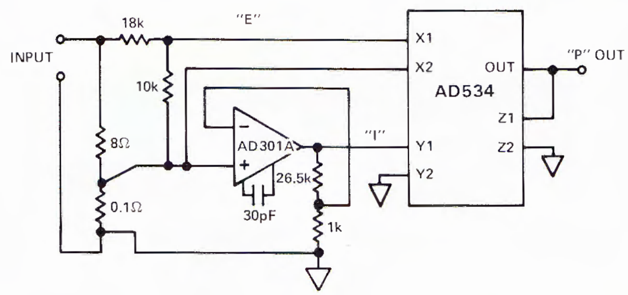


Figure 7. Wattmeter measures audio-amplifier output power into dummy load

Current is measured with a 4-terminal, 0.1Ω shunt. The 0.1Ω "steals" only about 1/80 of the voltage coming out of the amplifier, but the differential voltage measurement ensures that even this small difference between the amplifier output and the load voltage does not affect the measurement of the power dissipated in the load. The AD301A op amp is used to amplify the signal up to manageable levels and to present it to the multiplier's Y input. The output of the multiplier is $(X_1 - X_2)(Y)/10$, which is proportional to the product of load voltage and current, hence the power dissipated in the load.

In practice, the output of the amplifier under test should be connected to the input of the circuit through number 16 wire. Connections should be made with soldered lugs to minimize contact resistance, and the load should be a high-wattage-capacity, non-reactive 8Ω resistor. A number of vendors supply 8Ω, 10W resistors, encased in finned, integral heat sinks, which may be bolted to aluminum plates for optimum dissipation characteristics. An 8Ω loudspeaker could be used as a load, but the cost of the necessary anechoic chamber or concert hall (to avoid loss of friends in the laboratory – or one's hearing) should be considered. This system can be used to test instantaneous amplifier power into a resistive load as a function of frequency and waveform. With an averaging output, it will test average power. With a loudspeaker, it will determine how much power is delivered to a real load.

Some interesting characteristics of incandescent lamps can be observed if the 8Ω load is replaced by a light bulb and a transistor switch. This configuration was used to determine the optimum price-performance breakpoint in a computer-controlled scoreboard. Trace A of Figure 8 is the bulb voltage, trace B is the AD301A "current" amplifier output, and trace C is the AD534 "power" output.

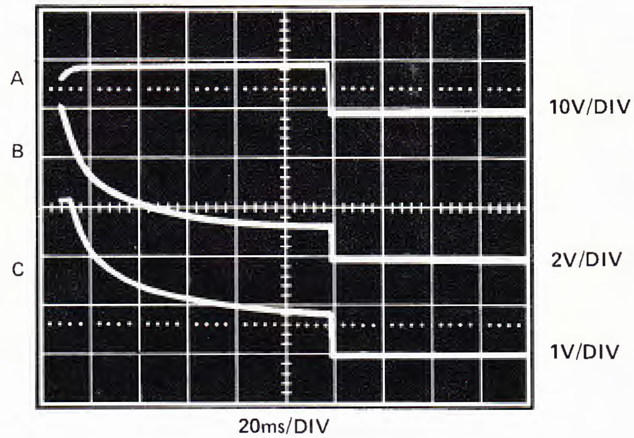


Figure 8. Turn-on power of cold incandescent lamp

The waveforms show that the bulb pulls almost 2.5 times the power at turn-on as it does in the steady state. In Figure 9, the bulb has been pre-biased just below the illumination level by connecting a "leak" resistor across the switch. This pre-heating of the filament dramatically reduces the turn-on power requirements, which results in an increase of bulb life. In addition, the transient demand on the power supply (which also runs the computer) is reduced, eliminating logic-deranging spikes.

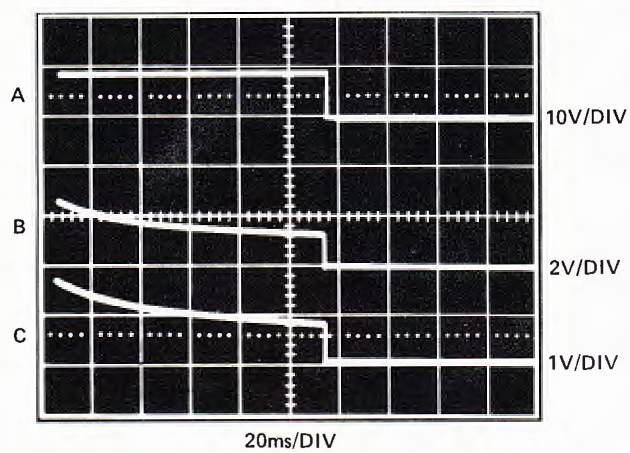


Figure 9. Turn-on power of incadescent lamp from standby

For three-phase wattage measurements, voltages proportional to the voltages and currents in each phase are multiplied in each of three analog multipliers and summed in an operational amplifier.[3]

WATT-HOUR METER

A power-measuring circuit similar to that used in Figure 8 is employed in the watt-hour meter circuit of Figure 10. An isolation amplifier allows the multiplier output to be measured without safety hazard. The Model 284J isolation amplifier is chosen for economy and because it furnishes a fully floating dual supply, which drives the active circuitry. The output of the 284J represents the instantaneous power delivered to the load, which is a "40-watt" light bulb.

[3] "Detection and Measurement of Three-Phase Power, Reactive Power, and Power Factor, with Minimum Time Delay," by I. R. Smith and L. A. Snyder, *Proc. IEEE*, November, 1970, p. 1866.

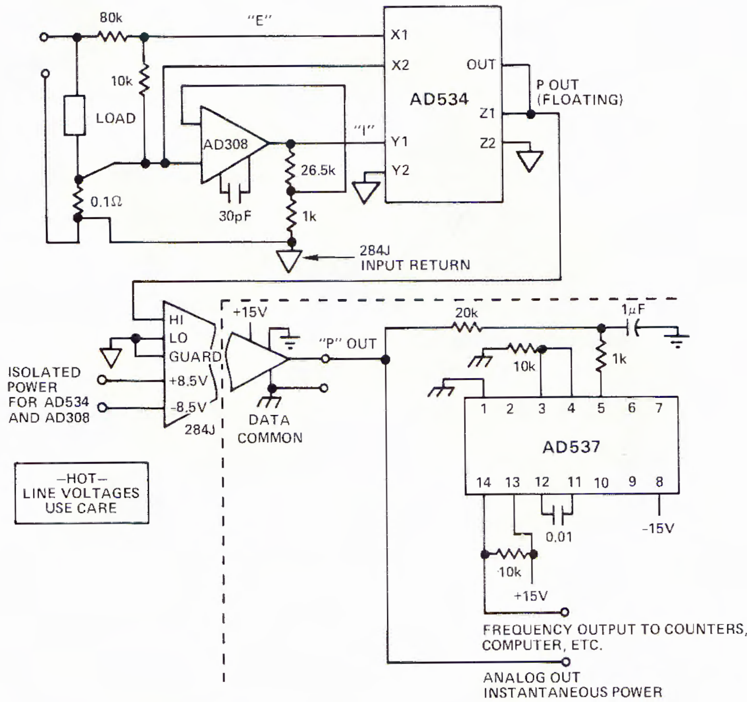


Figure 10. Line wattmeter — watt-hour meter

To obtain the watt-hour function, this signal is averaged and converted to frequency by an AD537 V/f converter. The pulse repetition rate of the AD537 will vary in direct proportion to the average power consumed by the load. A counter will determine the total energy consumed over the desired interval.

Differing sensitivities (watt-minutes, watt-milliseconds, etc.) can be obtained by altering the scale factor of the AD537, the gain to the 284, or the count ratio on the AD537 output.

If an analog output is desired, an analog integrator may be successfully employed within the limitations of the components chosen. Accuracies to within ±1% or better are achievable with time constants of 100ks, but 1000s is perhaps a more practical limit.

FLOWMETER

The accurate measurement of the flow rate of liquids flowing at slow speeds presents a difficult transduction problem. A flow transducer of wide dynamic range at low flow rates may be configured with the aid of analog multipliers and a pair of well-matched, linear-responding temperature sensors.

The theory behind the flowmeter can be understood by referring to Figure 11. The conceptual transducer is composed of a section of tubing, a heater dissipating a constant amount of power into the tubing, and two temperature sensors. The entire assembly is wrapped in Fiberglas to limit and smooth the thermal loss rate.

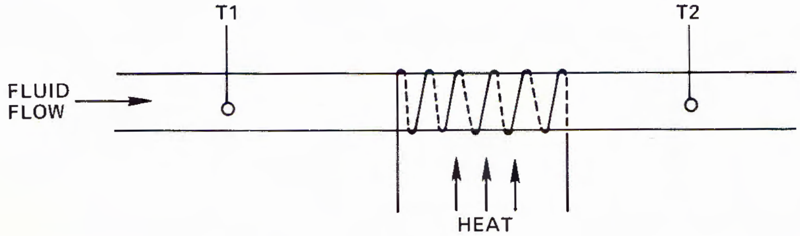


Figure 11. Flowmeter principle

With no flow through the pipe, the power is dissipated into the medium in a symmetrical manner. Under these conditions, the temperature sensors, T1 and T2, will indicate a net temperature difference of zero. As flow begins, T1 will continue to assume the value of the upstream temperature, but T2 will be influenced by the power dissipated into the moving stream.

The temperature difference between T_1 and T_2 will be solely a function of flow rate, so long as the specific heat of the fluid remains constant. Changes in stream or ambient temperature will be effectively offset by T_1, which serves as a baseline for the measurement.

It is worth noting that the flowmeter is bidirectional. If the flow reverses direction, the measurement is still valid. By detecting which sensor is hotter, and by how much, the transducer will indicate what the flow rate is and which way it is going.

A working version of the theoretical model is shown in Figure 12. The transducer comprises a ¼″ inside-diameter stainless steel tube which has been turned down and force fit through the core of a standard 20W ceramic-coated resistor. Silicon grease is used to facilitate thermal transfer between the resistor and the tubing. At each end of the tube, stainless-steel mesh has been inserted to break up the laminar flow of the liquid and promote mixing. The requirement for linear, matched temperature sensors is met with distinction by the AD590 current-output temperature transducers.

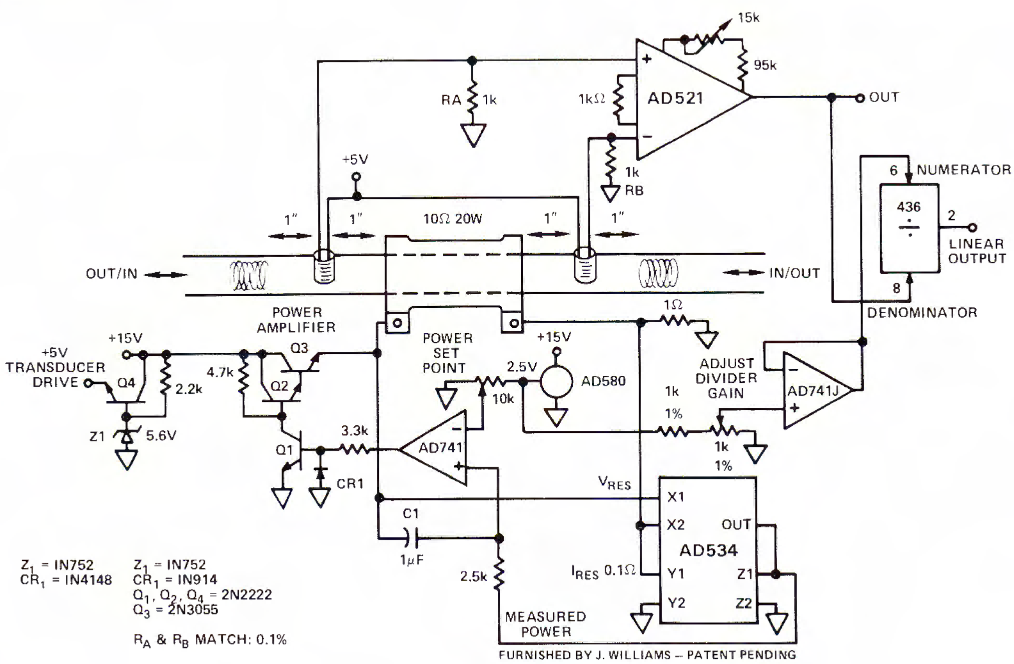


Figure 12. Flowmeter circuit

In theory, the power dissipated by the resistor can be held constant by driving it with a fixed voltage. In practice, changes in the resistor's value over time and temperature (remember, the resistor is being used as a heater, so the temperature rise will be quite substantial) mandate a requirement for a constant *wattage* regulator. The regulator determines the V• I product at the load (in similar manner to the wattmeters described in Figures 7 and 10), compares it to a dc set-point reference, and feeds the amplified difference back to the load.

At the wattmeter, X measures the voltage, Y measures the current in the 1Ω shunt, and the output of the AD534 therefore measures the power. The AD741 is used as the servo amplifier, and an AD580 bandgap reference provides the set point, which is compared with the power measured by the AD534. Boosted power to heat the resistor is provided via Q1, which provides voltage gain, and the Darlington pair, Q2-Q3, which provide current boost. Since Q1 inverts the AD741's output, feedback is returned to the "+" input terminal. C1 provides dynamic feedback for stability and noise reduction. The D1-Q4 combination provides the 5V nominal drive potential to the AD590 temperature transducers.

The 1μA/K current output of the AD590's is converted to voltage by R_A and R_B, and the difference is amplified by the AD521 instrumentation amplifier, set for a gain of 100V/V.

Figure 13 is a plot of flow rate vs. the output voltage of the AD521. The function is hyperbolic (flow *period* is measured linearly) and may be converted to a linear measure of the flow rate by performing a division (Const/x) to obtain the reciprocal. The function could be performed in a divider-connected general-purpose multiplier, like the AD532 or the AD534, but for wider dynamic range and better performance at low-level outputs from the AD521, a high-accuracy dedicated divider, such as the 436, can be used, as shown in the figure.

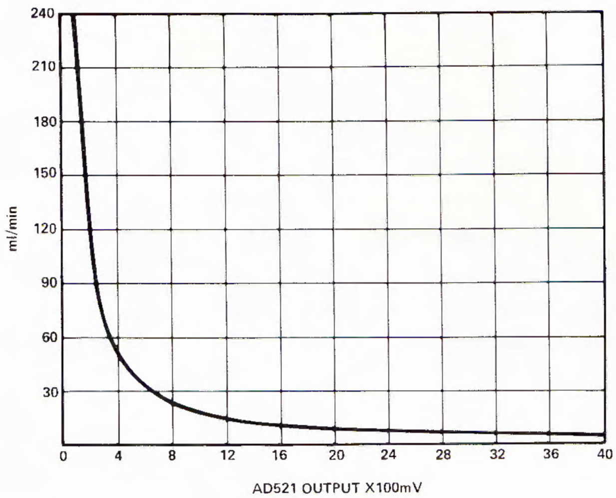


Figure 13. Flowrate vs. output voltage of the AD521, showing inverse relationship

The output of the divider can be calibrated to read (for example) 10V at full scale. Other scale factors are possible by changing the gain of the AD521 or of the divider. The time response of the flowmeter is slow – of the order of 10-15 seconds. In most low flow-rate applications, this is not a severe penalty, unless the flow rate changes quickly over a wide dynamic range.

DENSITOMETER

In this application, a mathematical function that is not easy to achieve with a single multiplier is needed for linearizing. The problem has been resolved by the use of a Model 433 multifunction component, which embodies the mathematical relationship $Y(Z/X)^m$. In Figure 14, the density of a liquid moving in a pipe is computed by measuring the difference between velocity-produced pressure and static pressure in the pipe. This output is nonlinearly related to density, and the 433 performs the linearization.

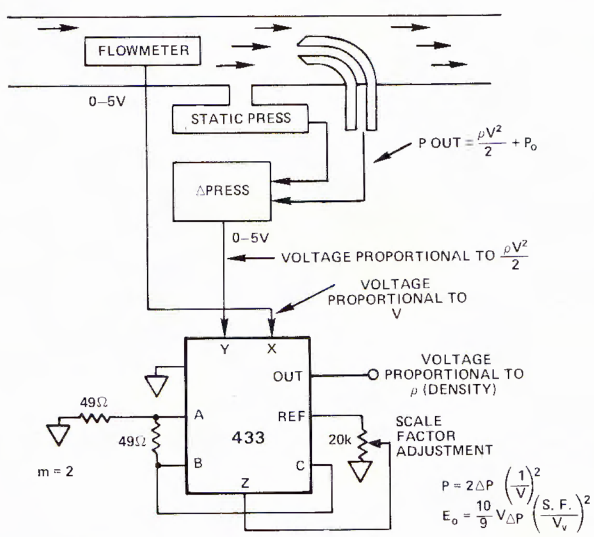


Figure 14. Block diagram of densitometer

The static pressure in the pipe is taken through a simple orifice in the pipe wall. The velocity-produced pressure is generated by a Pitot tube. The pressure produced in the Pitot tube is governed by the relationship,

$$P_{out} = \rho V^2/2 + P_0$$

The differential pressure at the output of the pressure sensor is:

$$\Delta P = P_{out} - P_o = \frac{\rho V^2}{2}$$

Therefore, the density, ρ, is

$$\rho = 2\Delta P \left(\frac{1}{V}\right)^2$$

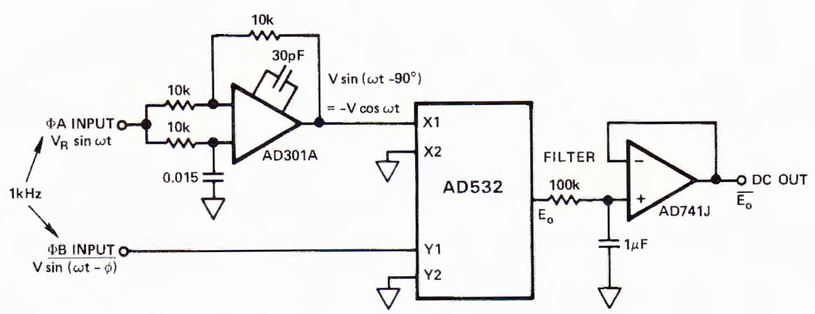
If ΔP and V are measured electrically, the analogous voltage corresponding to ρ can be computed electrically using a 433, with Y analogous to ΔP, X analogous to V, Z a scale constant, and m = 2.

The choice of flowmeter and pressure sensor depends upon the ranges of the variables and the conditions of measurement. In the case of the actual device illustrated here, a high-level-output LX-series pressure transducer was used, together with a flowmeter consisting of a paddle wheel, which interrupted a light beam from a light-emitting diode, and a frequency-to-analog converter.

PHASE MEASUREMENT AND PHASE-SENSITIVE DETECTION

Although there has been a vast increase in the use of square waves and pulses, in their various degrees of freedom (amplitude, frequency, phase, duty cycle, code), for conveying information, sine waves are still widely used. The analog multiplier is a simple and useful way of recovering information from sine waves.

Figure 15a shows the AD532 connected as a phasemeter. At a given frequency, if the phase reference signal is applied to one input of the multiplier, and the phase-shifted signal is applied to the other input, and a further 90° phase shift is introduced between the inputs, then the average value of the product is proportional to the sine of the phase angle. Since the sine is approximately equal to the angle (in radians) for small angles, this circuit provides a good linear measure of phase for small angles (within 0.0025 radian for angles up to 0.25 radian, or 14°), and a sinusoidal measure for angles between ±90°.



$$E_o = -\frac{V V_R}{20}\left[\sin 2\omega t \cos\phi - \cos 2\omega t \sin\phi - \sin\phi\right]$$

$$\overline{E_o} = \frac{V V_R}{20}\sin\phi \cong \frac{V V_R}{20}\phi$$

Figure 15a. Phasemeter for sinusoidal signals

Figure 15b shows a phase-sensitive detector, a "soft" rectifier for sinusoidal signals, with positive output if the signal is in phase with the reference (about 0°) and negative output if the signal is out of phase (about 180°). The average value of the output is

$$\overline{E_o} = \frac{1}{20} V \cdot V_R \cos\phi$$

Thus, the output is pretty much independent of small phase variations around 0° to 180° (2.5° corresponds to 0.1%, 8° corresponds to 1%).

Incidentally, the form of the equation indicates that the same configuration will also serve as a computation of "real power" in an ac system, while the configuration of 15a will compute reactive power.

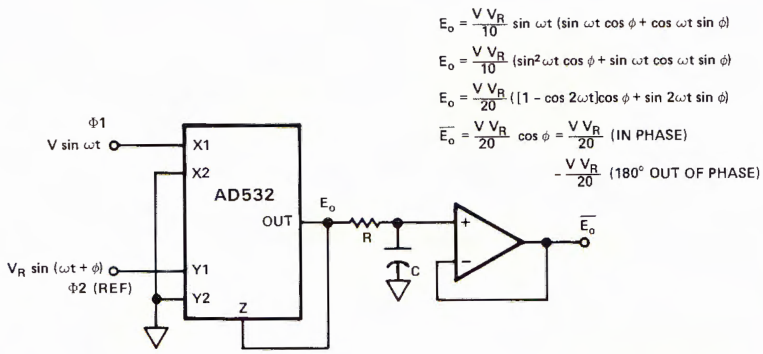


Figure 15b. Phase-sensitive detector for sinusoidal signals. Measures magnitude of in-phase or 180°-out-of-phase input with proper polarity, depending on phase relationship to reference, with less than 1% nominal error for phase shift between signal and reference of less than 0.14 radian (8°)

Figure 15c shows a phase-sensitive detector in which the reference signal is a square wave. If the signal and the reference are in phase, the output is positive; if they are 180° out-of-phase, the output will be negative. The output magnitude will be $\frac{1}{10}\ V \cdot V_R\,|\sin \omega t\,|$, and the average output will be the ac average $0.636\,V_{in}V_{ref}$ for sine waves. The circuit will tolerate small phase shifts between the signal and the reference.

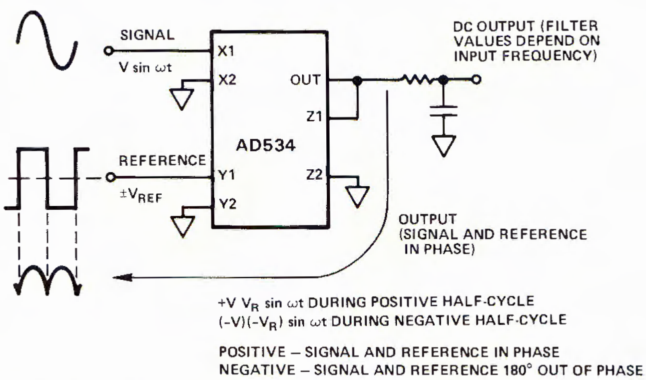


Figure 15c. Phase-sensitive detector, square-wave reference

ACOUSTIC THERMOMETER

A novel and wide-range temperature sensor can be constructed by using the relationship between the speed of sound and absolute temperature. Acoustic thermometers rely on the principle that the speed of sound varies predictably with temperature in a known medium. They are usually implemented as either clocked (pulsed) systems or oscillators. In both modes, the sensor is, in effect, a thermally dependent delay line.

In theory, the speed of sound through a medium is predictable and reproducible over temperature ranges from cryogenic to thousands of degrees. Acoustic techniques will function at extremes which other sensors cannot tolerate. The relationship between the speed of sound in dry air (for example) and temperature is:

$$c = 331.5\sqrt{\frac{T}{273}}\ \text{m/s.}$$

Thus, for a measured value of c, the absolute temperature is

$$T = \frac{273}{331.5^2} \cdot c^2$$

If c is measured in terms of the length of time, Δt, required for a sound impulse to travel a given distance, λ, the absolute temperature is

$$T = \frac{273\,\lambda^2}{331.5^2}\left(\frac{1}{\Delta t}\right)^2$$

Thus, in addition to the physical hardware for implementing the measurement, a means of computing the inverse square is needed. It could be the 433, as used in Figure 14. However, in this case, a Model 436 high-accuracy dedicated divider was used for reciprocation, and an AD534 multiplier was used for squaring.

A linearized temperature sensor using analog components is shown in Figure 16. Some of the critical waveforms are shown in Figure 17. The clock pulse (trace A) simultaneously sets the flip-flop, drives the piezoelectric 40kHz transducer, and triggers the 74121 one-shot into its output *low* state. This causes Q6 to turn off, allowing the AD812-AD820 current source to begin charging C4, the 0.04μF integrating capacitor (trace C).

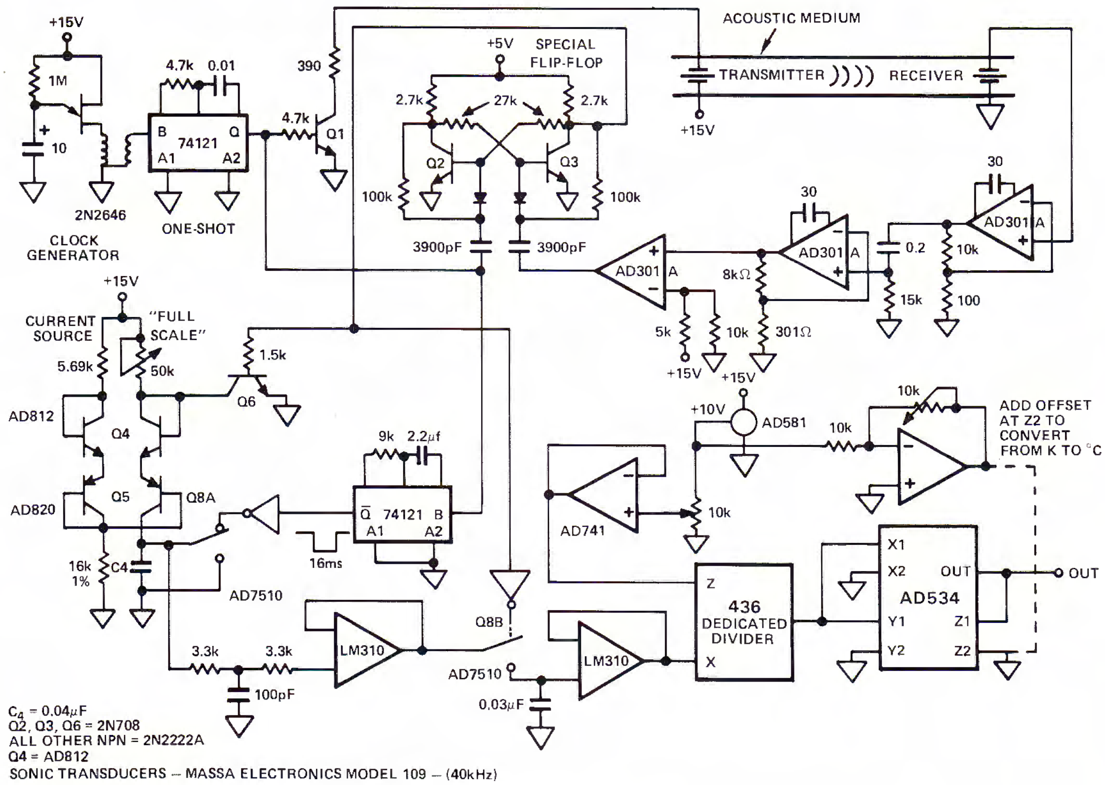


Figure 16. Acoustic-thermometer analog output

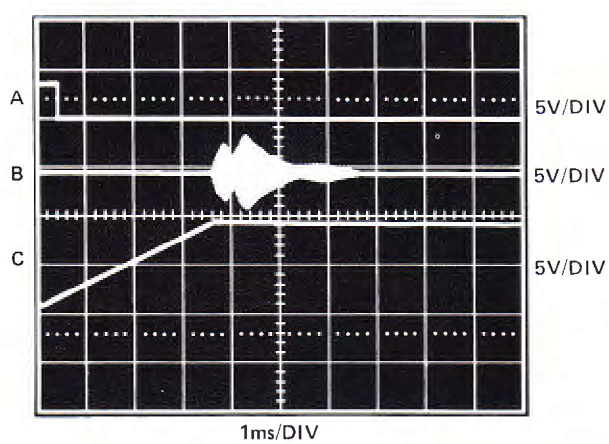


Figure 17. Waveforms in acoustic thermometer

When the acoustic pulse reaches the receiver transducer, the output is amplified by the AD301A amplifiers, producing the damped pulse train of trace B – the output of the second AD301A. The leading edge of the pulse train is used to reset the Q2-Q3 flip-flop, via the third

AD301A, which functions as a comparator. When the flip-flop resets, Q6 is turned on, and the current source turns off very quickly, typically in 10ns.

The voltage on C4 at this time is a function of the length of the tube and its temperature. The full-scale trim is adjusted by altering the slope of the ramp at the current source with the 50kΩ potentiometer in the AD812 collector line. The voltage at C4 is unloaded by the first LM310 follower, and the value is stored in the sample-hold formed by switch Q8, the 0.03μF capacitor, and the second LM310. The stored voltage is proportional to the transit time of sound in the tube.

To obtain an output reading proportional to temperature, the output of the sample-hold is applied to the denominator input of the Model 436 precision divider; a constant derived from the AD581 reference is applied to the numerator. The output of the 436 is squared in an AD534, thus obtaining an output voltage which is a measure of the absolute temperature. A voltage derived from the AD581 may be optionally applied to input Z2 of the AD534 to subtract a constant, if the analog readout is desired in degrees Celsius. The AD534 output will be related to the temperature at the transducer to a typical accuracy of better than 1% over a 0° to 100°C range.

The prospective constructor of this circuit should be aware of the difficulty of developing a good transducer design. The design used here is crude in comparison to what can be achieved. Acoustic thermometers involve a great deal of engineering to compensate for errors in the sonic transducers, thermal expansion effects in the tube walls, wave dispersion inside the tube, and sundry other potential problem areas. This example, while certainly workable, is primarily intended to show the key role of the analog multiplier and divider in obtaining an output that linearly represents temperature, once a time interval can be measured.

SIGNAL GENERATION AND FILTERS

WIEN-BRIDGE OSCILLATOR

Figure 18 is a schematic diagram of a stabilized Wien-bridge oscillator. The AD534 serves as a variable-gain amplifier for the feedback signal from output to input, via the Wien bridge. The peak-rectifier & filter combination applies sufficient voltage to the X (denominator) input to support stable oscillation with about 0.2% ripple. The circuit has no startup problems,

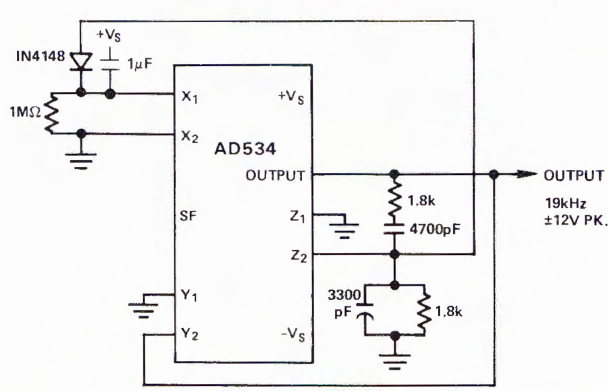


Figure 18. Stabilized Wien-bridge oscillator

since X is small and the gain very high, allowing rapid buildup of the oscillation. Tighter amplitude control is possible with other schemes at the expense of simplicity. This circuit will typically stay within 0.01dB of amplitude over 10°C temperature range and ±1V supply variation.

VOLTAGE-CONTROLLED SINE-WAVE OSCILLATOR

Voltage-to-frequency converters using charge balancing and other techniques (for example, Models 456, 454, 460, and the monolithic AD537) are readily available and feature excellent performance at low cost. However, *voltage-controlled oscillators with sine-wave output* are not so plentiful and constitute a non-trivial design task if reasonable performance is desired.

Figure 19 shows two multipliers being used to form integrators with controllable time constants in a 2nd-order-differential-equation feedback loop. R2 and R5 are connected for controlled current-output operation. The currents are integrated in capacitors C1 and C3, and the resulting voltages at high impedance are unloaded by the X inputs of the "next" AD534. The frequency-control input, E_{in}, connected to the Y inputs, varies the integrator gains, with a sensitivity of 100Hz/V, to 10V. C2 (proportional to C1 and C3), R3, and R4 provide regenerative damping to start and maintain oscillation. The diode bridge, CR1-CR4 and Zener diode Z1 provide economical temperature-compensated amplitude stabilization at ±8.5V by degenerative damping. Figure 20 shows the VCO's response to a ramp input.

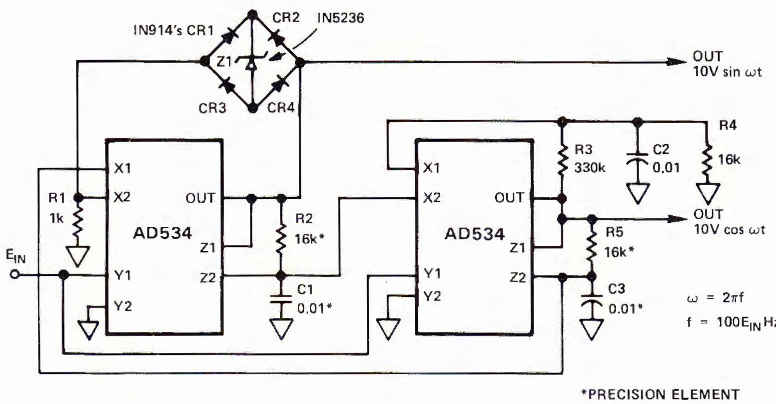


Figure 19. Voltage-controlled sine-wave oscillator

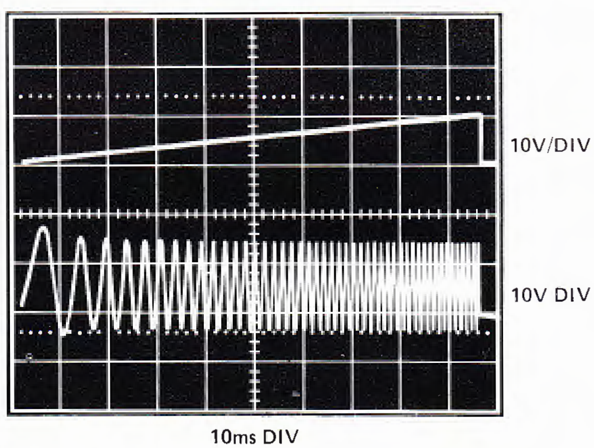


Figure 20. Ramp-modulated output of V.C.O.

CRYSTAL OSCILLATOR WITH AMPLITUDE-MODULATED OUTPUT

Fast amplitude slewing and settling of a crystal-stabilized oscillator are provided by an AD534 in the circuit of Figure 21. This arrangement was used to test 32.768kHz clock crystals for Q vs. amplitude.

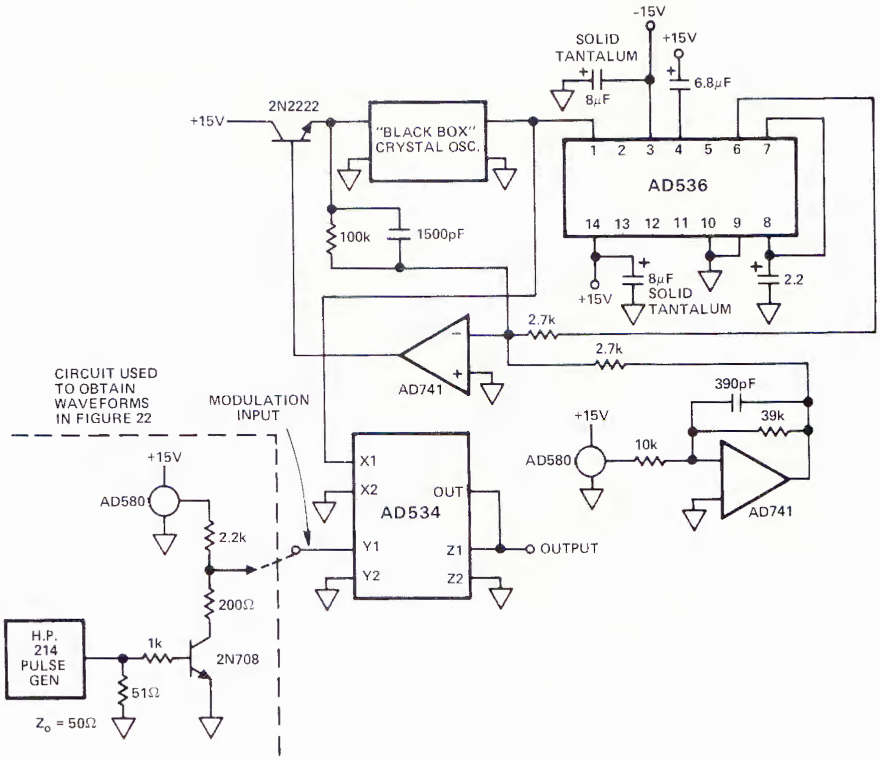


Figure 21. Crystal oscillator circuit, showing high-speed gain adjustment. The circuit at the lower left was used to obtain the test waveforms in Figure 22.

The "black-box" crystal oscillator's output is converted to dc by the AD536 rms/dc converter. The AD536 output is summed in an AD741 with the dc reference voltage obtained by inverting and amplifying the output of an AD580 band-gap reference.* The AD741 drives the 2N2222 control transistor to close the feedback loop around the oscillator by adjusting its supply voltage. The op amp runs at a gain of 35V/V, as determined by its feedback circuit. The 1500pF capacitor stabilizes the loop.

Amplitude modulation could be obtained by changing the reference voltage, but settling time would be long because of time constants in the AD536 filter circuit, as well as any unknown poles in the "black-box" oscillator.

Fast settling is conveniently obtained by using an AD534 as an *amplitude modulator.* The oscilloscope photo of Figure 22 shows the 32.768kHz waveform being switched by a fast step

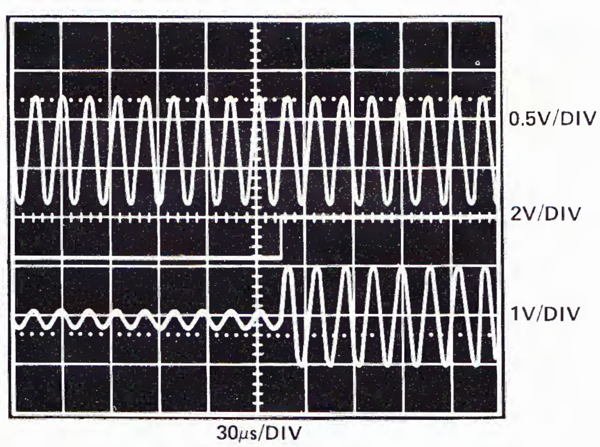


Figure 22. Amplitude modulation of crystal oscillator output. Upper trace is unmodulated output, middle trace is modulating waveform, lower trace is modulated output.

*The newly available AD581 could also have been used, in its connection as a two-terminal -10V reference.

from 0.3 to 2.2 volts peak-to-peak, with no ringing, overshoot, or other untoward dynamics. Since an rms-to-dc converter is used in the control loop to measure the output for setting the amplitude, the output of the circuit can be readily calibrated in rms volts, if desired.

LOW-DISTORTION OSCILLATOR

A low-distortion (0.01%) oscillator is depicted in Figure 23. Amplifier Al is connected as a non-inverting gain of about 3. The band-pass filter, R1, C1, R5, C2, provides notched feedback at 1kHz, causing the circuit to oscillate at $f_0 = (2\pi RC)^{-1}$, where $R_1 = R_5 = R$ and $C_1 = C_2 = C$. The output amplitude is measured and compared with the set-point voltage by R6 and R7 at the input of A2. Integrator A2 accumulates the error and applies voltage to input Y1 of the AD532 multiplier. This will increase or decrease the gain in the damping feedback loop around A1. This loop can be visualized as a "smart" resistor, of large value, that parallel-trims R4 and adjusts the overall gain of A1 to keep the oscillation stable.

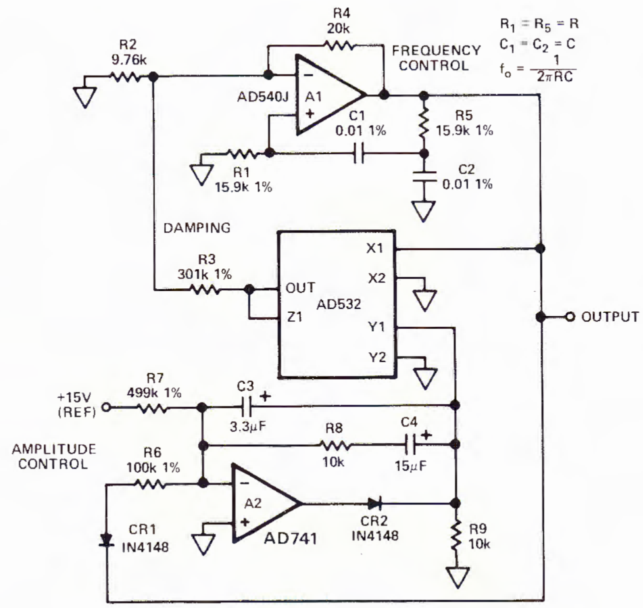


Figure 23. Low-distortion sine-wave oscillator

Since the multiplier is linear, and its output is attenuated to provide a "vernier" gain adjustment on the oscillator amplifier, its distortion has negligible effect upon the output. The distortion is due primarily to the nonlinearity of the op amp. An AD540J provides 0.01% distortion at the frequency in this example, 1kHz. Figure 24 shows the output waveform in trace A and the crossover distortion, greatly amplified (from the output of a distortion analyzer) in trace B.

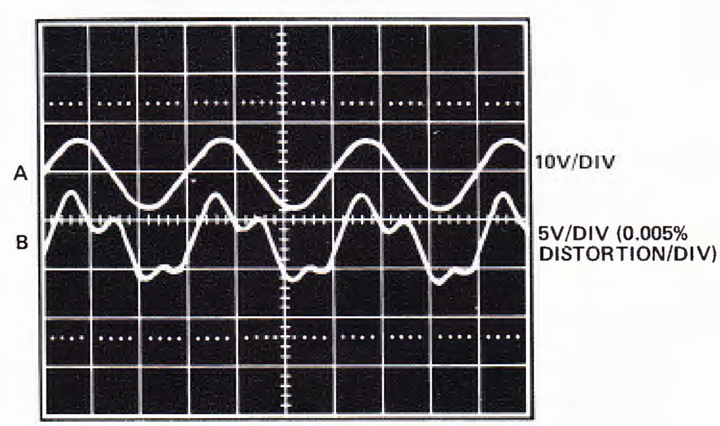


Figure 24. Waveforms in low-distortion oscillator – output and distortion

VOLTAGE-CONTROLLED LOW-PASS FILTER

Figure 25 is the circuit of a controlled low-pass filter, and Figure 26 shows its response to a square-wave input, as the control input ramps the time constant from very slow to quite fast response.

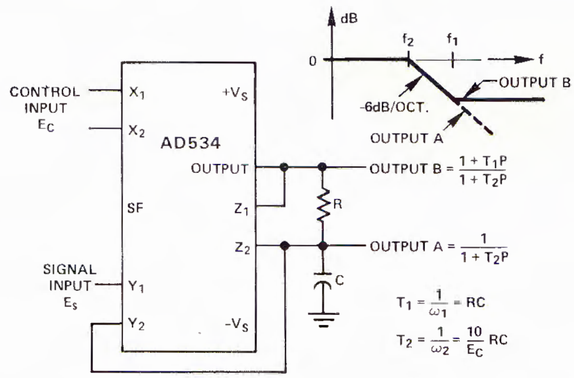


Figure 25. Voltage-controlled low-pass filter

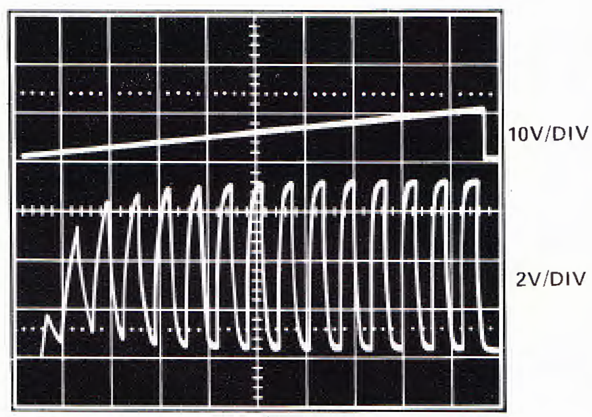


Figure 26. Response of low-pass filter to square wave as cutoff frequency is linearly increased

The voltage at output A, which should be unloaded by a follower, responds as though the input signal, E_S, were applied directly to the filter, but its break frequency is modulated by E_C, the control input. The break frequency, f_2, is equal to $E_C/(20\pi RC)$, and the rolloff is at -6dB per octave.

Output B, the direct output of the AD534, has the same response up to frequency f_1, the "natural" break point of the RC filter $(1/(2\pi RC))$, then levels off at a constant attenuation of $f_2/f_1 = E_C/10$.

For example, if R = 8kΩ, C = 0.002μF, Output A has a pole at frequencies from 100Hz to 10kHz, for E_C ranging from 100mV to 10V. Output B has an additional zero at 10kHz (and can be loaded, since it's the low-impedance multiplier output). The circuit can be converted to high pass by interchanging C and R.

DERIVATIVE-CONTROLLED LOW-PASS FILTER

Figure 27 shows an interesting variation of the voltage-controlled low-pass filter. A well-known difficulty with conventional linear filters is that long time constants provide excellent filtering, but they also require a great deal of time to settle. The filter described here settles rapidly in response to step changes, then assumes a long time constant for filtering (small amounts of) noise.

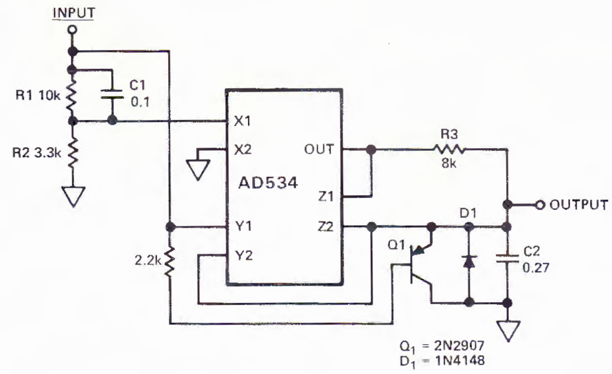


Figure 27. Derivative-controlled low-pass filter

In the circuit of Figure 27, a controllable low-pass filter, like that of Figure 25, has its control input driven from a high-pass filter (i.e., a differentiator), consisting of R1, R2, and C1. When a step voltage is applied to the input, the time constant is immediately determined by the voltage applied to X1, the amplitude of the step; then the control voltage exponentially decays until it is about 25% of the step, (the divider ratio $R_2/(R_1 + R_2)$), which increases the filter time constant fourfold.

Figure 28 shows salient waveforms in the circuit. Trace A is the input step, trace B is the control input, showing the immediate jump in cutoff frequency, and the decay to a cutoff frequency of about ¼. Trace C is the signal output, showing a rapid response – faster, in fact, than that of the control input, followed by a long tail, indicating the low steady-state cutoff frequency.

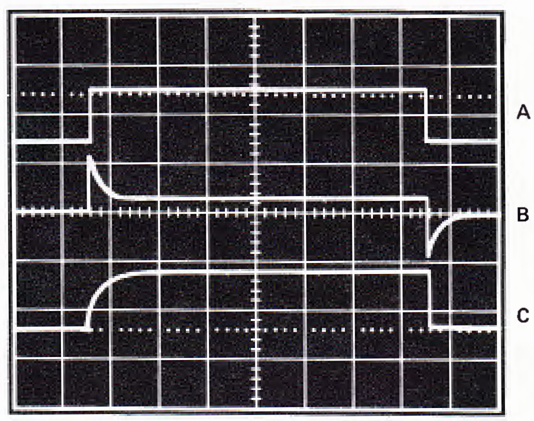


Figure 28. Waveforms in derivative-controlled low-pass filter

Since the spike produced by the R1, R2, C1 network will go in the wrong direction, tending to greatly lengthen the time constant, when the square-wave returns, a clamping circuit, consisting of Q1 and D1 rapidly resets C1 to zero. If derivative control of the falling edge were desired, an absolute value amplifier could be inserted between the X1 input and R1-R2 junction.

In this example, the ultimate time constant is determined by the height of the step and/or the ratio of the resistors R1-R2. If there are to be other influences on the time constant, appropriate additional voltages could be summed in at the X2 input. It is important to note that this circuit will function properly only if the noise is relatively small and has lower-frequency components than the initial step.

This kind of approach to filtering has proven useful in electronic weighing applications, where long time-constants are undesirable when weighing an object, yet "floor noise" has to be filtered out.

In deference to objectivity, another nonlinear approximation, using simple circuitry but not as fast-settling, can exhibit related behavior. Figure 29 shows a pair of diodes and an R-C filter. This circuit will only exhibit low-pass characteristics after the capacitor has charged to within 0.6V of the applied signal. Further, the diode could be replaced by a computer-controlled switch, allowing the "breakdown voltage" to be any desired value.

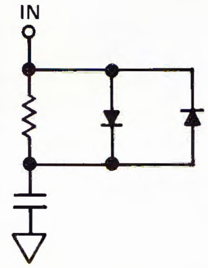


Figure 29. Diode-controlled low-pass filter

MISCELLANEOUS MULTIPLIER APPLICATIONS

% DEVIATION – RATIO COMPUTER

Figure 30 shows a circuit that computes the percent deviation between its two inputs. The scale factor in this arrangement is 1% per volt, but other scale factors can be obtained by altering the resistor ratios. The percentage deviation function is of practical value for many applications in measurement, testing, and control. For example, the output of this circuit might be applied to a pair of biased comparators to stimulate particular actions or displays depending on whether the gain of a circuit under test were within limits, or deviating by a preset amount in either direction.

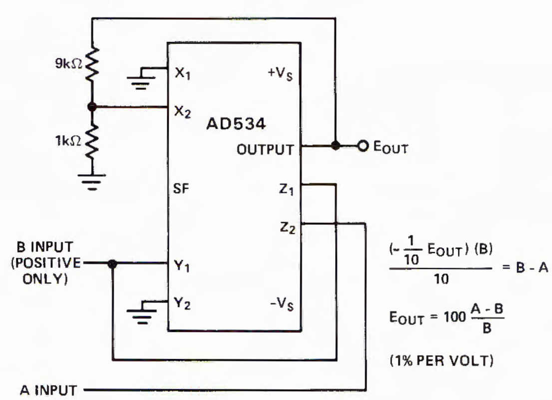


Figure 30. Δ% ratio computer

COMPONENT SORTER

The circuit of Figure 30 forms the basis for the component sorter shown in Figure 31. This circuit will grade capacitors, resistors, or Zener diodes by percentage deviation from a settable value. The circuit comprises a switchable current source, a clock, a sample-hold network, and some timing logic. Waveforms are shown in Figure 32.

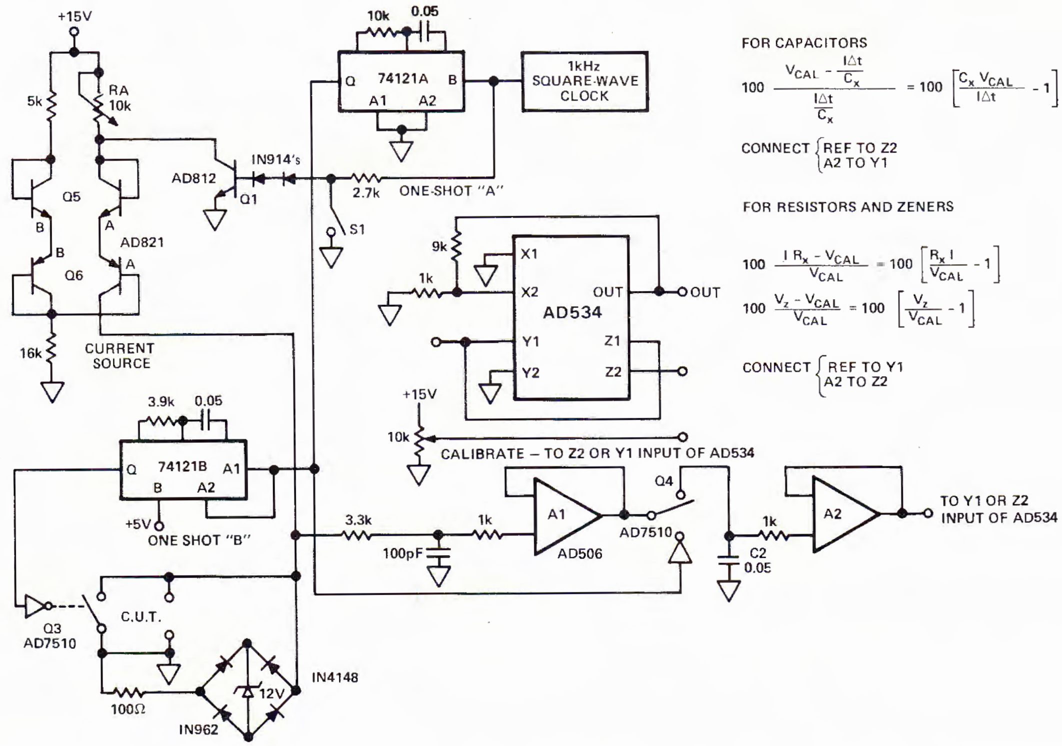


Figure 31. Capacitor, resistor, and Zener-diode sorter

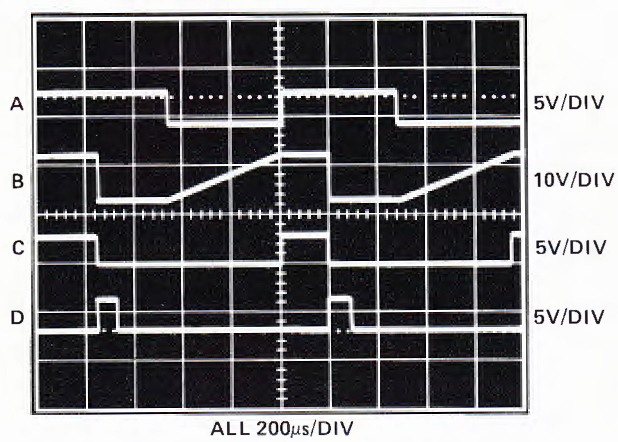


Figure 32. Waveforms in C, R, Zener sorter

The "A" portion of the AD821 functions as the current-source transistor, and the "B" portion provides temperature compensation. The AD812B prevents the AD821 pair from conducting in the reverse direction whenever the voltage across the component under test (C.U.T.) exceeds the AD821B emitter voltage (when Q1 is on). The adjustable 10kΩ resistor, R_A, sets the output of the current source. The 1kHz square-wave clock (trace A) is applied to Q1, turning the current source on and off. The capacitor under test, in this case 0.01µF, is allowed to charge until the clock goes high, turning off the current source (trace B). The voltage the capacitor sits at is inversely proportional to its absolute value.

The AD506 follows this potential and feeds the simple sample-hold circuit, Q4, C2, A2. The sample-hold is enabled by the one-shot "A" for 200µs when the clock goes high (trace C). After this time, "A" goes low, triggering one-shot "B" on for 100µs. This pulse drives Q3 on (trace D) and discharges the C.U.T.

The output of the sample-hold is applied to the percent deviation circuit. In practice, a 0.01µF standard capacitor is put into the fixture, and the reference voltage is adjusted so that the AD534 output reads zero. The circuit is now ready for use at a scale factor of 1V/%

deviation. When capacitors are measured, the unknown is applied to input Y1, and the calibration input to Z2.

The 3.3kΩ-100pF network prevents the sample-hold from catching any portion of the discharge of the c.u.t. at the beginning of *hold* by introducing a slight delay into A1's response. The Zener-diode-bridge clamp provides protection for the tester in the event a charged capacitor is placed in the test fixture. The discrete-component current-source facilitates a ground-referenced two-terminal, high-noise-rejection test fixture.

The same two-terminal fixture may be used to check resistors and Zener diodes, by closing S1. This allows the current source to run all the time (necessary, since resistors and diodes have no "memory"). Because the test voltage will be proportional to the resistance or the Zener voltage, the connections to the Δ% circuit are reversed, the calibration reference is connected to Y1, and the test voltage is connected to Z2.

BRIDGE LINEARIZATION

If one arm of Wheatstone bridge varies from its nominal value by a factor, (1 + 2w), the voltage or current output of the bridge will be (with appropriate polarities and scale factors):

$$y = \frac{w}{1 + w}$$

Linear response requires very small values of w (to make the denominator essentially independent of w) and, as a consequence, preamplification.

The circuit shown in Figure 33 enables large-deviation bridges to be used without losing linearity or resorting to high attenuation. The circuit computes the inverse of the bridge function, i.e.,

$$w = \frac{y}{1 - y}$$

Depending on which arm of the bridge varies, it may be necessary to reverse the polarity of the X connections.

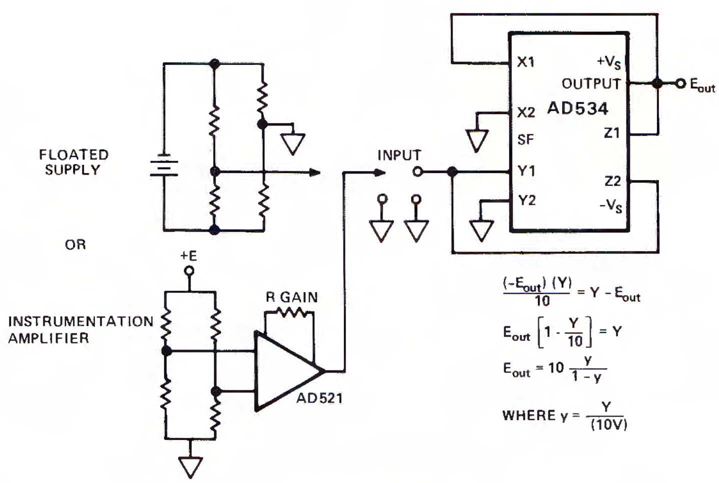


Figure 33. Bridge-linearization circuit

Since the input to this circuit is single-ended, the bridge must either float with respect to ground, or an instrumentation amplifier may be used to translate the bridge output to the AD534's common. Any resistive, linearly responding transducer (i.e., one or more legs of the bridge proportional to the phenomenon being measured) may profit from the application of this circuit. Examples include position servos, linear thermistors, platinum-resistance-wire sensors (nearly linear over wide ranges), pressure transducers, strain gages, etc.

HIGH-PERFORMANCE RMS-TO-DC CONVERSION CIRCUIT

For applications calling for greater accuracy and bandwidth, but with lesser dynamic range than the AD536 rms-to-dc converter can easily handle, the circuit of Figure 34, employing the AD534, will be found useful.

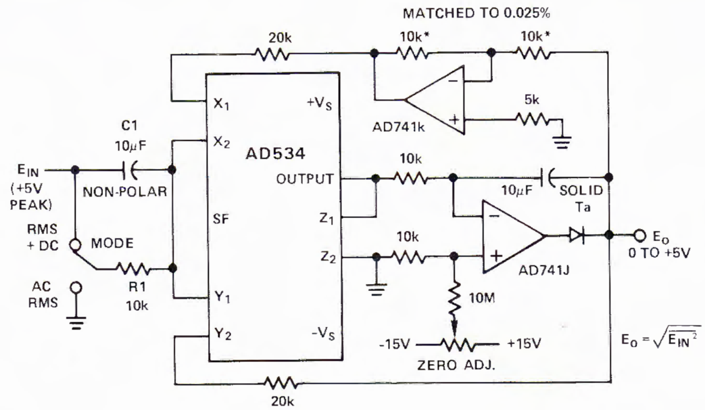


Figure 34. Wideband, high crest, rms-to-dc converter

The balance equation of this circuit is:

$$-(E_{in} + E_o)(E_{in} - E_o) = -10RC \frac{dE_o}{dt}$$

For steady state (or very slowly varying) values of E_o, the right-hand term approaches zero, the average value of $E_{in}{}^2$ is equal to $E_o{}^2$, and hence E_o measures the rms value of E_{in}.

After calibration, error <0.1% is maintained for frequencies up to 100kHz; it increases to 0.5% at 1MHz at 4V rms. Crest factors up to 10 have little effect on accuracy.

To calibrate, with the mode switch at "RMS + DC", apply an input of (say) 1.00V dc. Adjust the zero until the output reads the same as the input. Check for inputs of ±5V.

FREQUENCY DOUBLING

By use of trigonometric identities that can be found in any mathematical formulary, circuits can be assembled that accept sinusoidal inputs and generate outputs at two, three, four, and five-or-more times the input frequency.

Frequency doubling, in its simplest form, makes use of the identity,

$$\cos 2\,\theta = 1 - \sin^2\theta$$

Since, for a sine wave, $\theta = \omega t$,

$$\cos 2\omega t = 1 - \sin^2\omega t$$

If an input E sin ωt, is applied to a multiplier, connected as a squarer, the output will be

$$E_o = \frac{1}{20} E^2 (1 - \cos 2\omega t)$$

The dc term may be eliminated from the output by capacitive coupling, or by applying a bias voltage (which can be done conveniently with the AD534). Figure 35 shows a circuit which accepts a sinusoidal signal with a 10V amplitude and produces a double-frequency signal also having a 10V amplitude with no dc offset.

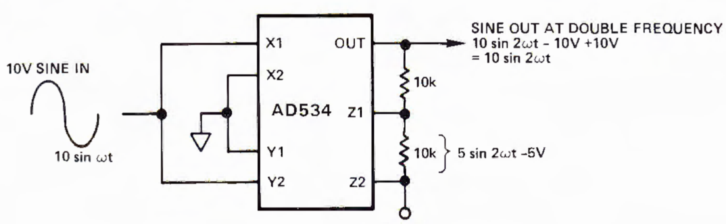


Figure 35. Frequency doubler

An obvious problem is that variations of input amplitude result in dc errors at the output. If the output is ac-coupled, abrupt changes of input level cause the output to bounce, which in some applications is troublesome. Figure 36 shows a circuit, using a different trigonometric identity, which produces frequency doubling, at a given frequency, with no dc offset, hence no bounce. It uses the identity:

$$\cos \omega t \sin \omega t = \tfrac{1}{2}\sin 2\omega t$$

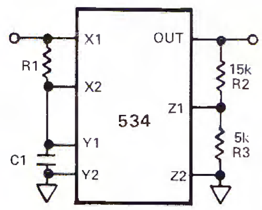


Figure 36. "No-bounce" frequency doubler

The X input leads the input signal, E sin ωt, by 45° (and is attenuated by $\sqrt{2}$), and the Y input lags the input by 45°, and is also attenuated by $\sqrt{2}$. Since the X and Y input are 90° out of phase (at the frequency $f = (2\pi RC)^{-1}$), the response equation of the circuit will be

$$\frac{1}{10}\frac{E}{\sqrt{2}} \cos \omega t \frac{E}{\sqrt{2}} \sin \omega t = \frac{E^2}{40} \sin 2\omega t = \frac{1}{4} E_o$$

The right-hand side reflects the attenuation of the output at the Z input. Hence,

$$E_o = \frac{E^2}{10} \sin 2\omega t$$

While this circuit is not wideband (as Figure 35 is), considerable frequency deviation can be tolerated without causing appreciable change in output amplitude. A ±10% frequency error causes a ±0.5% amplitude error. Obviously, frequency quadrupling can be effected by cascading doublers. The waveforms of Figure 37 show a sine wave that is doubled and then doubled again by repeating the circuit of Figure 35.

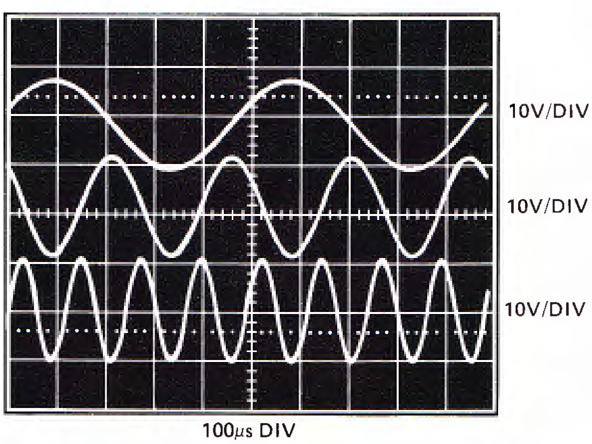


Figure 37. Sine wave – doubled and redoubled (and not vulnerable)

In frequency-doubling applications, amplitude is usually not critical. This is fortunate, because the squaring of the input amplitude doubles the sensitivity to amplitude errors, and distorts the modulation envelope. A circuit that uses feedback of the filtered output to the scale-factor input of the AD531 to produce frequency doubling with *proportional* amplitude is shown in Figure 38. A discussion of this circuit may be found on page 502 of the NON-LINEAR CIRCUITS HANDBOOK.

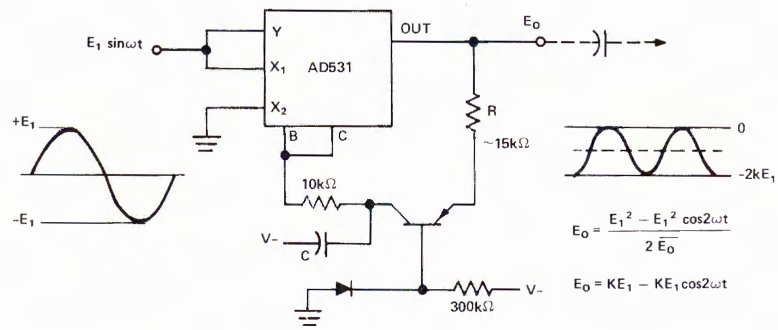


Figure 38. Frequency doubler with linear amplitude response

FILTER TESTER USING WIDEBAND MULTIPLIER

The high-speed testing of high frequency filters can pose some knotty problems. A circuit for testing a "black box" 1MHz bandpass filter for insertion loss (gain) vs. input amplitude is shown in Figure 39. The test is performed by sweeping the amplitude of the input signal and comparing the envelopes of the input and output signals.

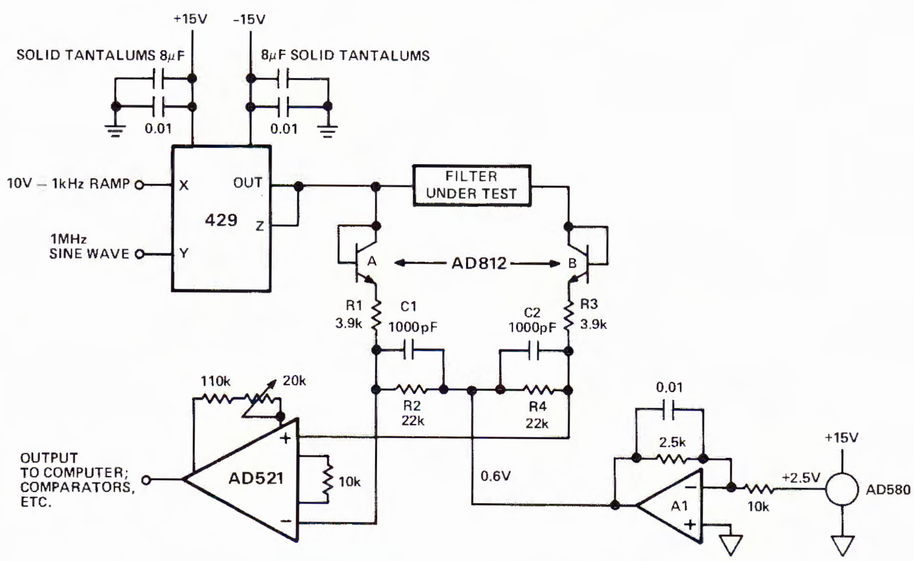


Figure 39. 1MHz bandpass filter insertion-loss tester

The test signal is produced by modulating a 1MHz sinusoidal carrier with a 10V, 1kHz ramp, using a high-speed Model 429 multiplier. The test signal is demodulated and filtered by the diode-connected AD812A and the network consisting of R1, R2, C1. The signal coming out of the filter is passed through an identical envelope-demodulation network, consisting of the AD812B, and R3, R4, C2.

The filter under test is specified to have zero insertion loss at 1MHz. Therefore, as the 1kHz ramp-modulated 1MHz signal is driven through the filter, the voltages across C1 and C2 should be identical, within an acceptable error band. Any discrepancy between the input signal and the filter response will manifest itself as a voltage difference between the two capacitors. Since both the input and the output are demodulated in the same way, any peculiarities introduced by the demodulator and its filter become common-mode factors.

The AD521 instrumentation amplifier compares the voltages at a gain of about 12. The offset error due to the AD812's approximately 0.6V drop is reduced to second-order by returning the capacitors to a –0.6V potential, provided by A1. Common-mode gain errors in the resistive dividers are calibrated out by fine trimming of the gain of the AD521. Other values of overall gain could be used for increased or decreased sensitivity, depending on the magnitudes of the errors that are being investigated, and their tolerances.

The AD521 output is a direct indication of whether the filter is "peaking" (positive output) or attenuating (negative output), by how much (absolute value of the output), and at what amplitude (the input ramp is also available to the monitoring processor, comparator, or whatever). The waveforms in Figure 40 illustrate the input ramp (trace A), the modulated test waveform (trace B), the demodulated ramp (trace C) and the demodulated filter output (trace D). Though not shown in the photograph, the AD521 output would provide a sensitive measurement of the difference between the two lower traces, which appear quite similar in the photo.

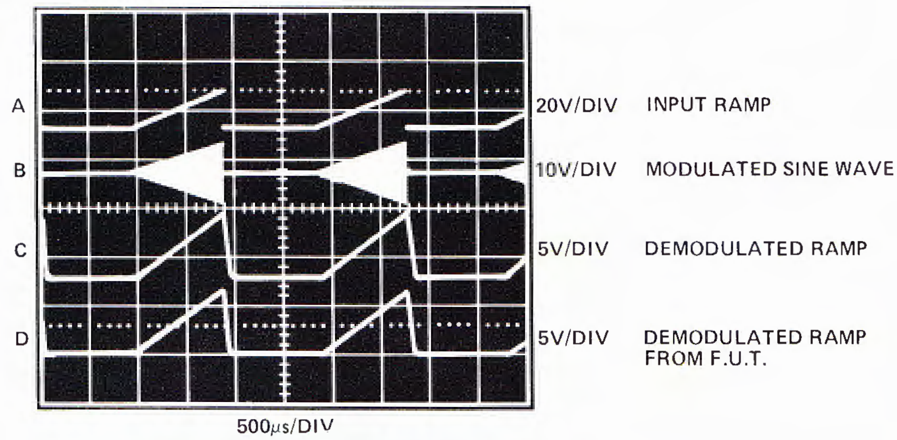


Figure 40. Waveforms in filter tester

PERFORMANCE AUGMENTATION

INCREASED ACCURACY WITH MULTIPLYING DAC'S

The present state of the analog multiplier mart allows 0.1% absolute accuracy at reasonably high speeds, using *translinear** multiplier circuitry (Model 435K), which has, by and large, superseded the slower *pulse-height-pulse-width* technique for high-precision applications.

Higher-accuracy performance can be achieved for many applications by the use of 4-quadrant multiplying d/a converters, especially for asymmetrical-input functions involving variable gains and modulation. Multiplying DAC's, which are commonly thought of as d/a converters requiring external reference voltage, can also be thought of as digitally controlled attenuators of analog signals. When considered for use as a multiplier, a 12-bit multiplying DAC, with its digital gain-setting resolution of one part in 4096, can be seen to be approaching an order-of-magnitude better performance than the best all-analog multipliers. The analog inputs of multiplying DAC's provide comparable accuracy and linearity, with quite low feedthrough, principally because the analog signal is conditioned by an attenuator consisting of high-quality linear resistors.

*These concepts are discussed briefly in the "Theory" section, and at greater length in the NONLINEAR CIRCUITS HANDBOOK.

Figure 41 is a simplified block diagram of the DAC1125 12-bit 2- or 4-quadrant multiplying DAC. The waveforms in Figure 42a were produced from measurements on a DAC1125. A square wave (trace A) multiplies a 40kHz sine wave, and the result is shown in trace B. The square wave in trace A was obtained by exercising the most-significant bit with a digital input (which could be the output of an a/d converter, when providing multiplication of two analog inputs). The gated 40kHz waveform appears cleanly at the DAC1125's output and is 12-bit accurate (±0.025%). Figure 42b shows the feedthrough of the analog input. With a ±10V 50kHz sine wave applied to the analog input and all digital inputs held high (the DAC1125 is complementary-coded in 2-quadrant operation), the feedthrough is seen to be comfortably less than 1mV peak-to-peak (1/20,000).

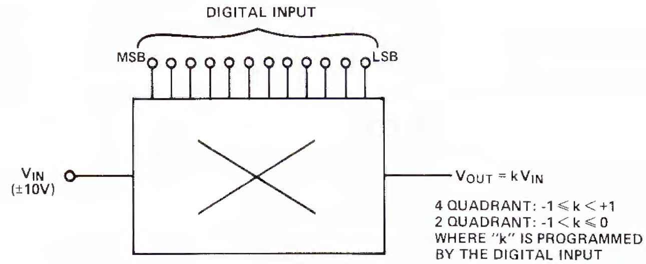


Figure 41. Functional diagram of the DAC1125 multiplying d/a converter

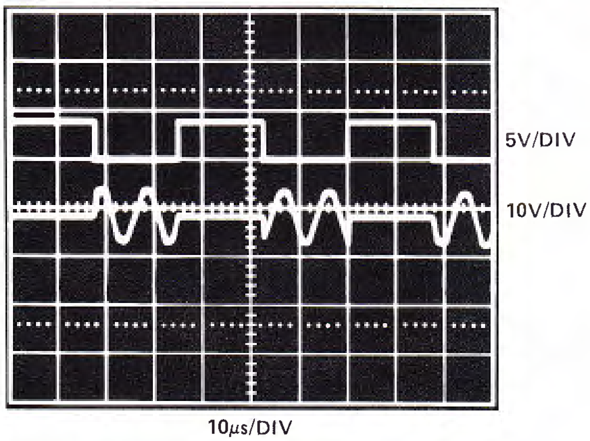


Figure 42a. Response of multiplying DAC to square-wave modulation of sinusoidal input

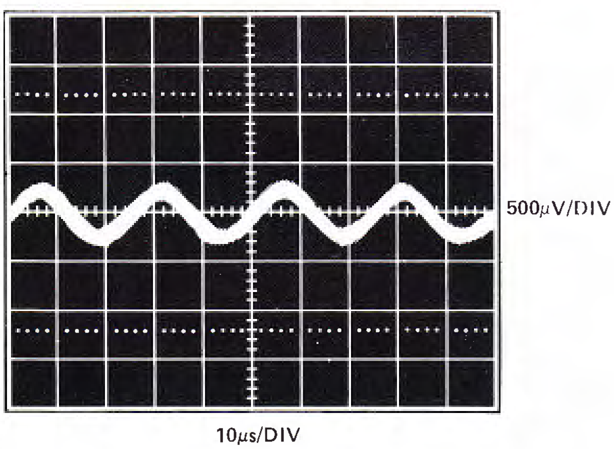


Figure 42b. Feedthrough of multiplying DAC, zero input multiplying ±10V 50kHz sine wave, 500μV/division

Although high-performance multiplying DAC's are evidently better suited to hybrid systems (digital/analog, or human-digital/analog, via keyboards or thumbwheel switches), which abound with applications for multiplying DAC's, it is worthwhile noting that high-performance multiplying DAC's, when coupled with 12-bit-or-better ADC's and a modicum of logic, make possible analog X analog multiplications of transcendent performance at realistic cost.

CURRENT OUTPUTS

The simplest form of output modification in the AD534 is the use of a current – instead of the usual voltage – output. The voltage-controlled oscillator in Figure 19 utilizes this readily achievable configuration. Since multiplication operations are often followed by integrators, the ease of providing a current output, which can charge a grounded capacitor, is attractive. The current-output conversion is shown in Figure 43. Waveforms of an AD534 integrating its current output into a capacitor are shown in Figure 44. Naturally, the capacitor voltage is read with a high-impedance follower.

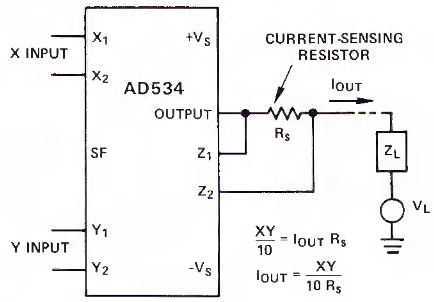


Figure 43. Output-to-current conversion

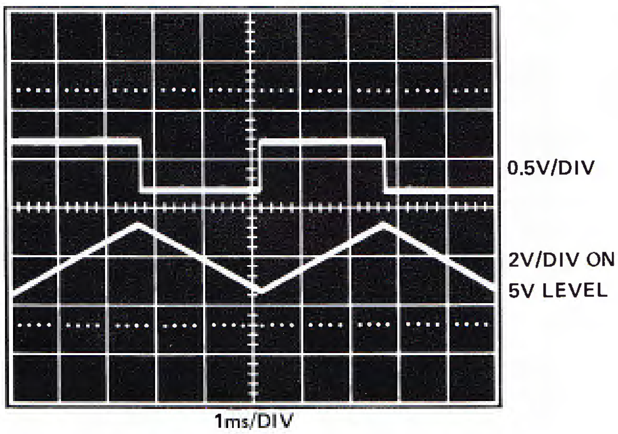


Figure 44. Response of AD534 connected as integrator; I_{OUT} charges a capacitor that is connected to ground. Ramp output for symmetrical square-wave input

CURRENT BOOSTING

The availability of an uncommitted high-gain feedback loop in the AD534, in combination with the device's excellent dynamic characteristics, invites the addition of output current boosters. Since the booster is added inside the loop, it has no effect on the salient parameters of the AD534. Figure 45 shows a current booster that will gladly drive ±12 volts into 50Ω, paralleled by 10,000pF, as fast as the AD534 asks it to. The booster uses standard components, and construction is not critical. The current sources, Q1–Q2 provide adequate biasing under all drive conditions. The 12Ω resistors provide current limiting and output-to-ground short-circuit protection; but it would be unwise to short the output to either supply.

Figure 46 shows the response of a booster, happily ensconced inside the AD534's loop, driving a ±10V, 333Hz square wave into a 50Ω, 10,000pF load.

AUDIO POWER BOOSTER

The configuration of Figure 47 is not just a *tour de force*, but a useful booster as well. The

AD534 is used in association with a high-quality audio power amplifier to produce a 75-watt output into 8 ohms. The amplifier used is a venerable favorite in research labs, due to its clean response and the transformer-isolated output.

Despite the amplifier's age and its vacuum tubes (remember them?), there is no "generation gap" in this family. The amplifier is inserted into the AD534's feedback loop in a similar manner to the current booster. The output is taken across 8Ω and attenuated by the 9kΩ-

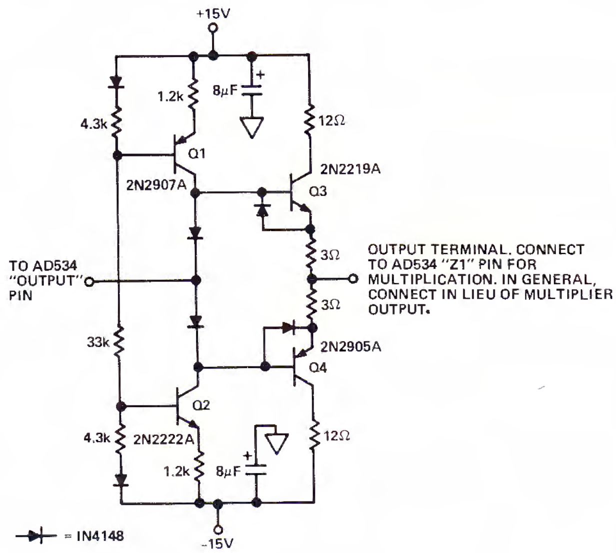


Figure 45. Current booster

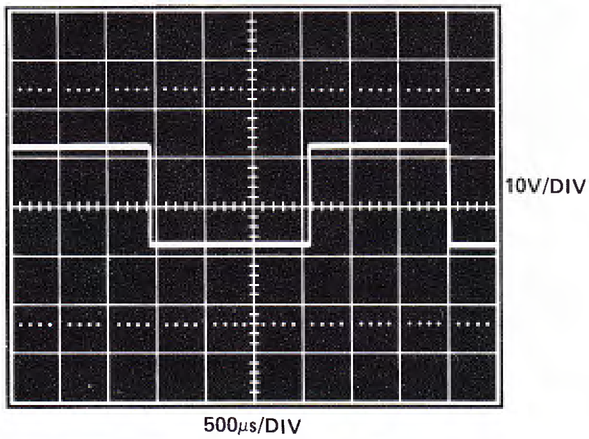


Figure 46. Square-wave response of current booster

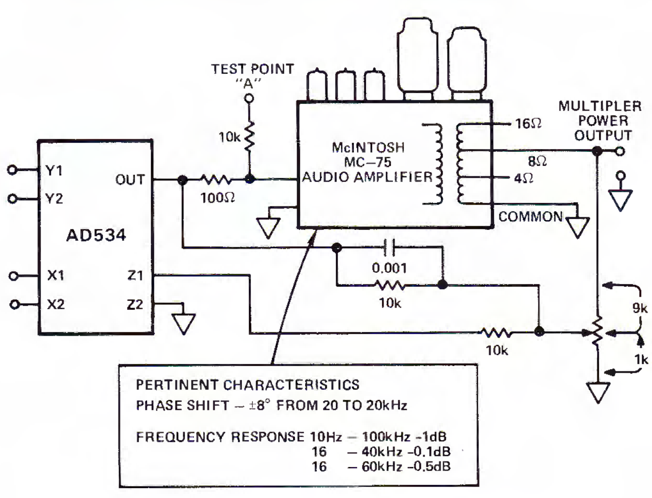


Figure 47. 75-watt inside-the-loop booster (ac only)

1kΩ feedback divider. The 10kΩ resistors provide a dc feedback path for the multiplier (around the transformer-coupled amplifier), and the 0.001μF capacitor provides dynamic stability for the loop.

Figure 48a shows the output of the boosted "multiplier" for X = a 2kHz ramp, Y = an 80kHz, 12V sine wave. In another test, with X = 10V, and Y = 0, a 4kHz, 10V square wave is injected at the test point "A" to test the loop gain. Figure 48b shows the happy result the first time it was tried (with no prior attempt made to optimize the loop response). The loop manages to grab control just 50μs after the disturbance occurs.

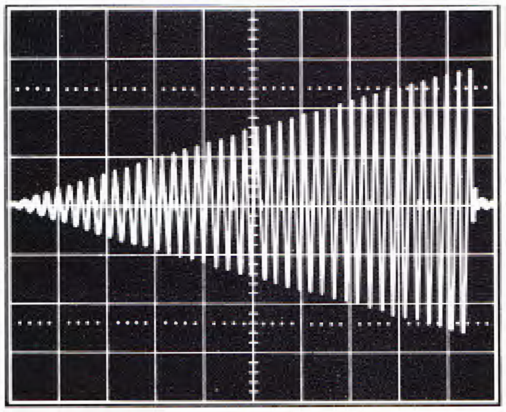


Figure 48a. Ramp-modulated sine-wave output of boosted multiplier

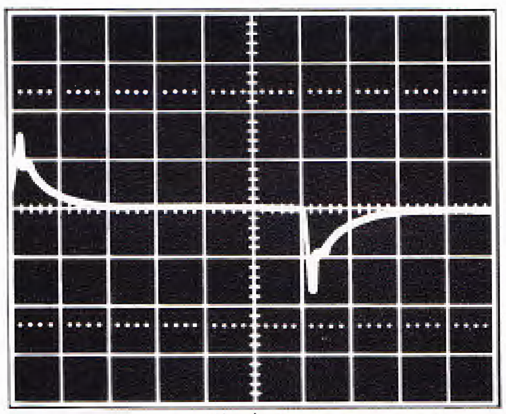


Figure 48b. Transient response of boosted multiplier

HIGH-VOLTAGE BOOSTER – BUT BE CAREFUL!

One of the popular techniques biochemists use in separating cells and proteins is electrophoresis. This involves exciting the sample with high-voltage potentials so that separation-by-charge can take place. It is important that the sample dissipate a constant amount of power during the run, which may take 12 hours or longer. To meet this need, constant *wattage* supplies are used. Figure 49 shows a booster that meets this need by transforming the 300mW output of an AD534 into a 1000V, 300W roaring Goliath.

BEFORE PROCEEDING ANY FURTHER, THE READER IS ADVISED THAT THE CONSTRUCTION, DEBUGGING, AND USE OF THIS CIRCUIT MUST BE UNDERTAKEN WITH EXTREME CAUTION. THE OUTPUT PRODUCED BY THIS CIRCUIT IS MANY TIMES ABOVE THE LEVEL NECESSARY TO KILL AND IS ABSOLUTELY LETHAL. THERE IS NO SUBSTITUTE FOR CAUTION, PRUDENCE, AND CLEAR THINKING WHEN WORKING WITH HIGH-VOLTAGE CIRCUITRY.

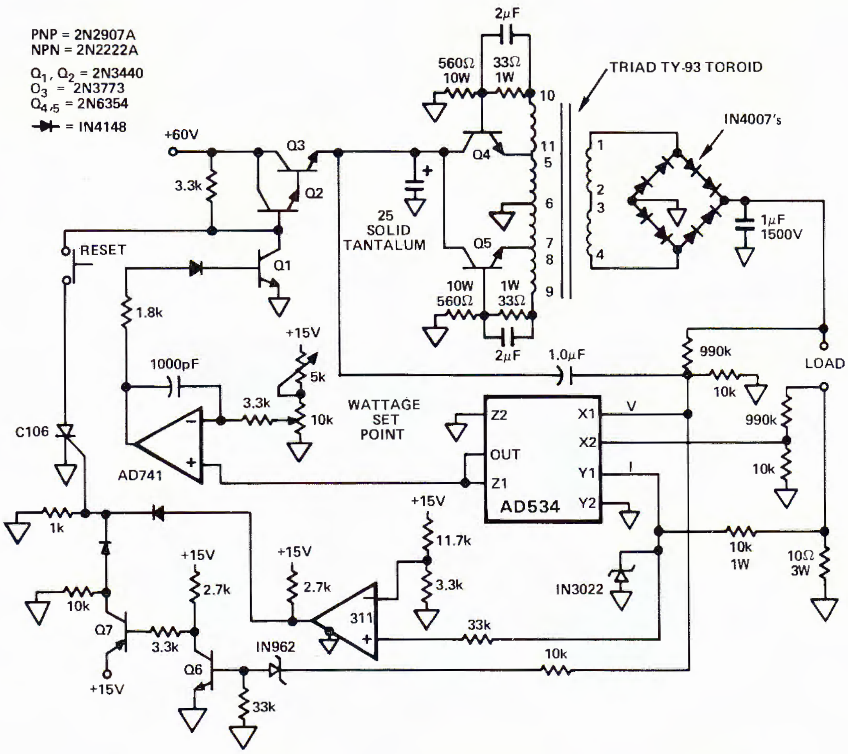


Figure 49. High-voltage booster. DANGER! SEE TEXT. USE CAUTION IN ASSEMBLING, TESTING, AND USING. PROVIDE ADEQUATE WARNING TO PROTECT OTHERS.

This booster, unlike the previous types, is not inserted into the feedback loop. Instead, it functions by using the power-measuring AD534's output to control the output of a toroidal dc-dc converter within the feedback loop of an amplifier. Both the multiplier and the dc-dc converter "hang" inside the AD741's feedback loop. The 741 sees, and is only concerned with, the low-voltage causal relationship between its input and its output . . . it is oblivious to the bustling nature of its complex feedback loop.

When power is applied, the circuit output is zero, the "V" and "I" multiplier inputs are at zero, and the multiplier output is of course zero. Under these conditions, the AD741 output goes negative, cutting off Q1. The Darlington pair, Q2-Q3 come on, and power is delivered to the self-exciting inverter, consisting of Q4, Q5, the transformer, and the associated components. The transformer output is rectified and filtered in the conventional manner. The 1μF capacitor is adequate for filtering the square-wave output of the converter.

Current and voltage are sensed at the load and multiplied together by the AD534. The voltage stolen by the current shunt is factored out using the differential inputs of the multiplier. The AD534 output is fed to the AD741, to be compared with the set-point voltage. The AD741, sensing the difference, provides a drive for Q1 in the appropriate amplitude and polarity to keep the output power servoed at the set-point value.

The 4kHz switching frequency of the inverter determines the ultimate limit on how rapidly information can be transmitted around the loop, but the 1μF capacitor sets the most significant response pole. Circuit stability is assured by rolling off the loop at a few Hz via the 1μF capacitor that is connected between the 2N3773 emitter and the output voltage-divider. The waveforms in Figure 50 were taken at the emitters of Q4 and Q5, at a 300W output level, at 1000V. Despite the high switching currents (7-8A), the combination of good transistors and a well-designed transformer yields clean waveforms, with a minimum of overshoot or ringing.

Overcurrent/short-circuit protection is provided by the AD311 comparator and the C106 SCR. If the current-sensor line rises above 3.3V (330mA), the AD311 output goes high and turns on the C106. The C106 grounds the drive to Q2 and latches *on*, shutting off all power to the dc-dc converter. Power is restored by removing the overload and pushing the "reset" button. Overvoltage protection is necessary for this supply, because a load dropout will cause the servo to "go through the roof" as it seeks to keep a constant wattage across an infinite-impedance load. If the voltage-sense line rises above the 12V Zener diodes' breakdown level, the Q6-Q7 pair conduct and provide the C106 with a short turn-on-spike. The C106 shuts down the supply in the same way as for a current overload.

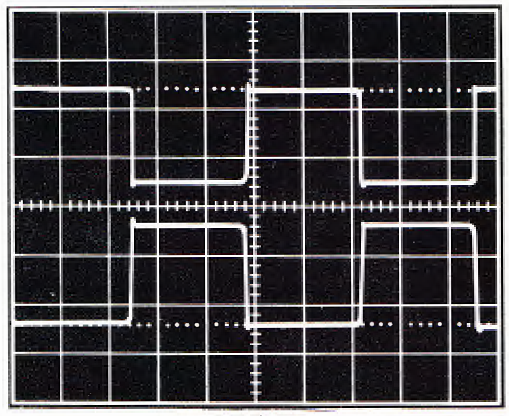


Figure 50. Switching waveforms at emitters of Q4 and Q5 in constant-power supply of Figure 49.

MULTIPLIER MEDLEY

The circuits in the following collection, although presented in somewhat less detail than many of those in the preceding sections, are neverless valuable, despite their brevity. Some will work as described, others are offered in conceptual form. All are useful and should prove catalytic for the alert reader seeking solutions. Further information can be found in the NONLINEAR CIRCUITS HANDBOOK, individual product data sheets, and (where noted) ANALOG DIALOGUE.

DIFFERENCE OF THE SQUARES[1]

The circuit of Figure 51 will compute the difference of the squares of two input signals. This is useful in vector computations and in weighting the difference of two magnitudes to emphasize the greater nonlinearly. This circuit can also be used to determine absolute value if "A" is the input, "B" is connected to E_0 through a diode, and both Z terminals are grounded.

The balance equation is:

$$A^2 - B^2 = 0$$

Therefore the output, B, must be equal to the absolute value of A.

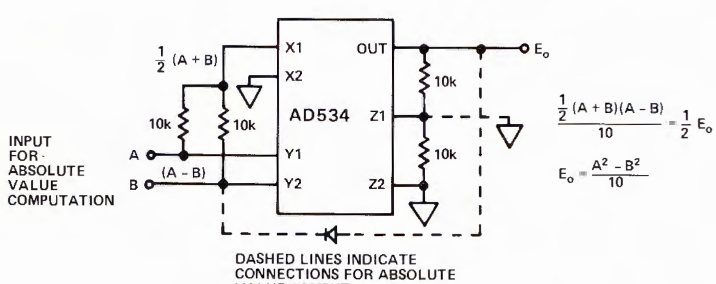


Figure 51. Difference-of-squares circuit

[1] ANALOG DIALOGUE 11-1, 1977, p. 8

AUTOMATIC LEVEL CONTROL

Figure 52 illustrates a simple automatic level control. The AD533 is set up in the *divide* mode. Its output is rectified and compared with a power-supply-derived –15V reference. The net current is integrated by the AD741 and fed into the denominator input of the AD533, maintaining the level of the output at the "ac average" value programmed by the feedback circuitry, 7V, with ±1% stability, in this case. The level can be changed by changing the reference-current resistor. Normally, the output is ac-coupled, and no offset trim is necessary.

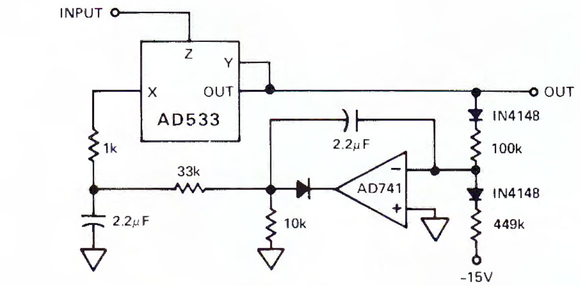


Figure 52. Automatic level control circuit

AUTOMATIC GAIN CONTROL

A more-sophisticated circuit is the automatic gain-control loop of Figure 53. Here, a low-cost AD531 maintains a 3V p-p output for inputs from 0.1V to over 12V, with 2% regulation for the range from 0.4V p-p to 6V p-p. Distortion is less than 1%. Input frequency can range from 30Hz to 400kHz.

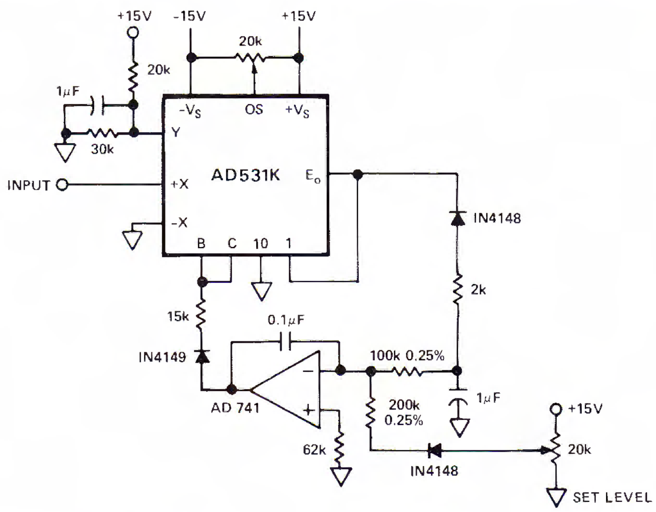


Figure 53. Automatic gain control circuit

If the input signal increases, the output will try to increase. Its negative peaks, caught by the diode and 1μF capacitor, tend to increase, causing the output of the inverting integrator to increase. This causes the denominator input of the AD531 to increase, which reduces the forward gain and tends to keep the output level constant.

AMPLITUDE MODULATOR

If a high-frequency carrier is applied to one input of a multiplier, and a modulation signal to the other input, the multiplier inherently acts as a "balanced modulator", with suppressed-carrier response, i.e., if no modulating signal is present, there will be no carrier output. In true amplitude modulation, the modulation effectively consists of a constant – to provide a continuous carrier – plus the modulation signal itself. Figure 54 shows how the AD534 can be used as a simple amplitude modulator. The continuous carrier signal is provided by summing the carrier at the Z2 input. Note that the modulation input can be differential (for example, the output of a bridge).

If the Model 429 wide-bandwidth multiplier is used as a modulator, the carrier is applied to one input, and the biased modulation signal is applied to the other input. With its 2MHz full-

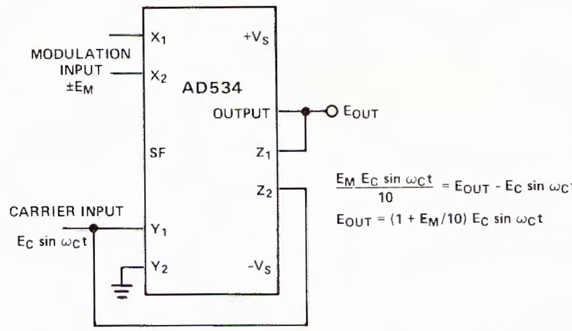


Figure 54. Linear (AM) amplitude modulator

power response, the 429 can put you on the air with your own AM radio station in the broadcast band (but be sure that you comply with F.C.C. licensing requirements!).

VOLTAGE-CONTROLLED AMPLIFIER[2]

In Figure 55, an AD534 is shown functioning as a voltage-controlled amplifier. A constant or varying signal, E_C, applied to the X input, controls the gain for the variable signal, E_{IN}, applied to the Y input. The inputs could be interchanged, but the Y input has the better linearity. For this application, which uses the AD534's "SF" terminal, the "set gain" potentiometer is typically adjusted to provide a calibrated gain of 10V/V per volt of E_C.

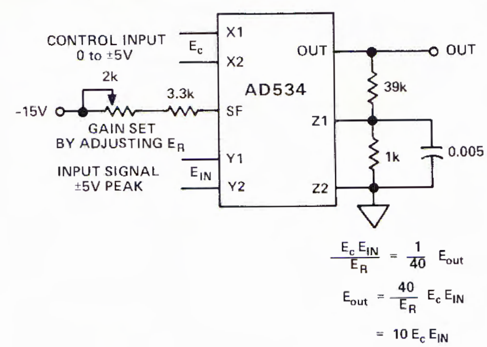


Figure 55. Voltage-controlled amplifier

Bandwidth is dc to 30kHz, independent of gain. The wideband noise (10Hz to 30kHz is 3mV rms, typically, corresponding to full-scale signal-to-noise of 70dB. Noise, referred to the signal input, is of the order of 60μV rms.

POLYNOMIALS – POWER SERIES

Polynomials can be effected with multipliers and summing operational amplifiers. Figure 56 shows the minimum number of multipliers required to accomplish the function for 2nd, 4th, and 8th-degree polynomials or truncated power series. The "X" blocks are multipliers, the "Σ" blocks are adder-subractor circuits.* With feedback, infinite series are possible, but the number of degrees of freedom for adjusting coefficients are limited (Figure 57). Detailed discussion of and mathematical considerations for these circuits are to be found in the NONLINEAR CIRCUITS HANDBOOK.

ARBITRARY (NON-INTEGRAL) POWERS

The Model 433 multifunction module can be used to generate powers and roots, with either continuous adjustment or fixed settings. Figure 58 shows the connections for both modes.

[2] ANALOG DIALOGUE 11-1, 1977, p. 7.

*See "Simple Rules for Choosing Resistor Values in Adder-Subtractor Circuits", ANALOG DIALOGUE 10-1, 1976, p. 14.

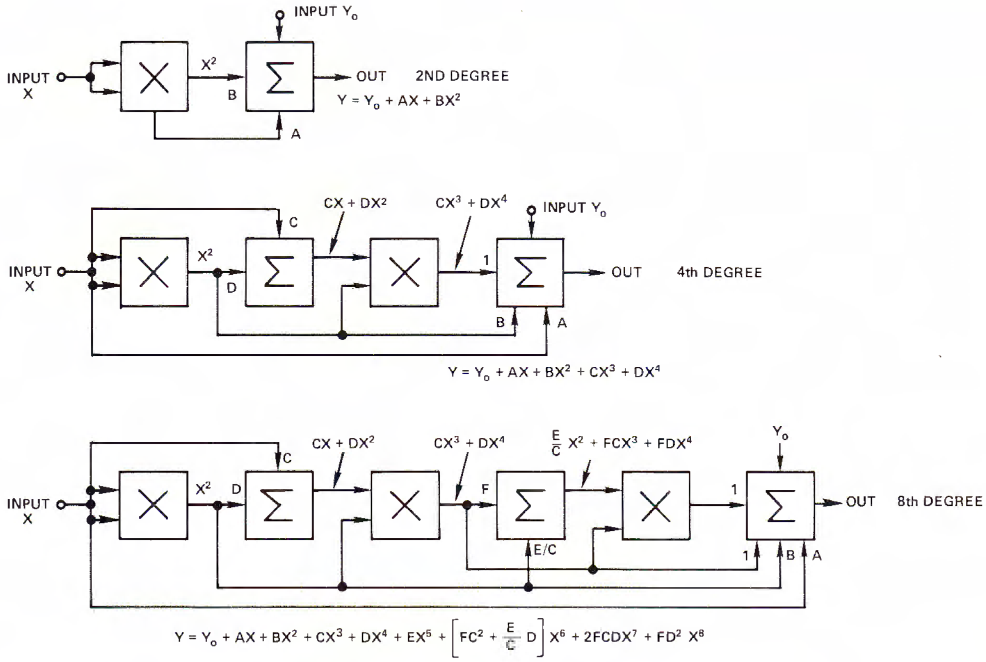


Figure 56. Polynomial (Truncated power series) block diagrams

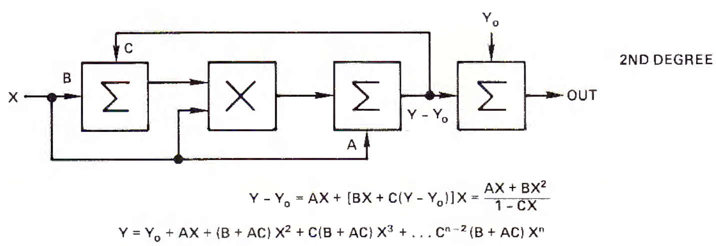
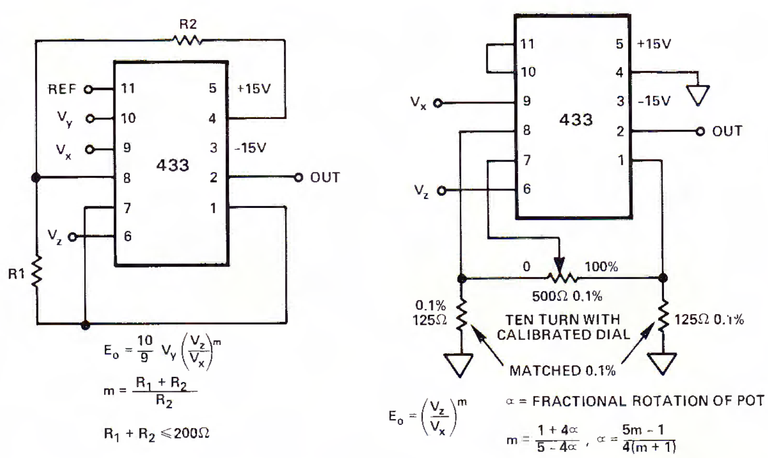


Figure 57. Second-degree polynomial with implicit feedback produces infinite power series



a) Fixed exponent, m>1 *b) Adjustable exponent, 1/5<m<5*

Figure 58. The 433 as an arbitrary fixed power generator or adjustable power/root generator

SINE OF A VOLTAGE[3]

Figure 59 shows how the AD534 can be used to approximate the sine of a voltage in one quadrant. With 0.1% resistors, the accuracy of fit will be to within 0.5% of full scale at all points.

[3] ANALOG DIALOGUE, 11-1, 1977 p. 8

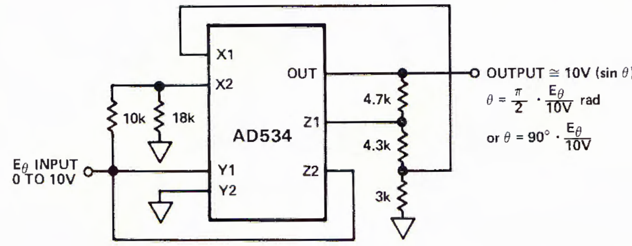


Figure 59. A simple sine-of-a-voltage circuit using the AD534

SQUARE-ROOT OF THE SUM OF SQUARES (VECTOR SUM)[4]

A high-accuracy three-input multiplier-divider (YZ/X), Model 434, is used to perform the vector computation shown in Figure 60. As shown,

$$E_o = V_B + \frac{V_A{}^2}{E_o + V_B}$$

From which,

$$(E_o - V_B)(E_o + V_B) = E_o{}^2 - V_B{}^2 = V_A{}^2$$

and

$$E_o = \sqrt{V_A{}^2 + V_B{}^2}$$

Note that the inputs are the absolute values of V_A and V_B, since the 434 is a single-quadrant device.

The method generalizes for n-dimensional vectors, using (n – 1) 434's, but no additional op amps. Each additional input adds a term, $V_i{}^2/(E_o + V_B)$, to the first equation.

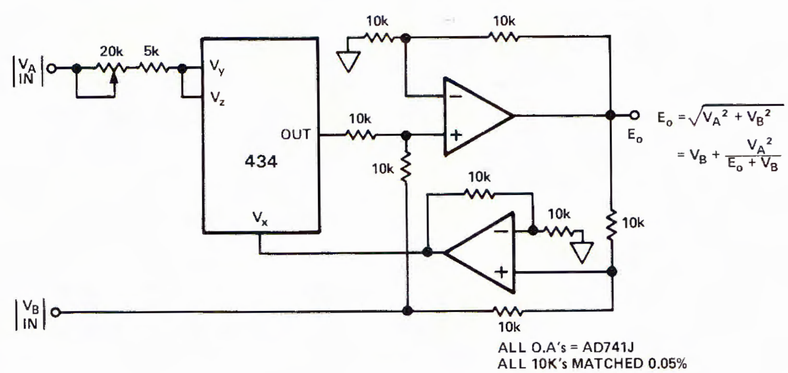
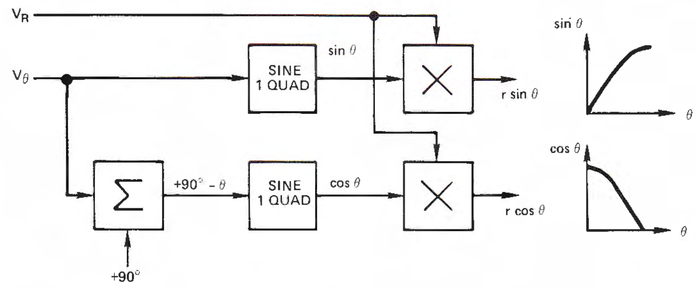


Figure 60. Square-root of the sum-of-the-squares

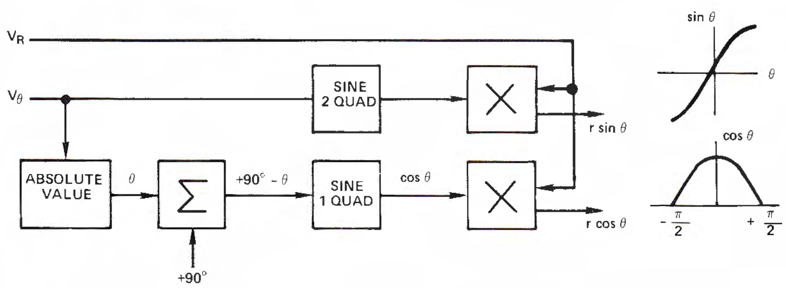
VECTOR OPERATIONS – POLAR-TO-RECTANGULAR

If a good circuit for fitting sin θ is available (a simple one is shown in Figure 59, others with greater accuracy and/or wider angular range can be found in the Appendix to Chapter 2-1 of the NONLINEAR CIRCUITS HANDBOOK), it can be used to perform vector computations of the form r sin θ and r cos θ (Figure 61). The cosine can be either developed as a separate function or by translating a sine 90°.

[4] ANALOG DIALOGUE, 6-3, 1972, p. 3. On page 5 of this issue, one can find an approximation for $\tan^{-1}(V_B/V_A)$, using a 433 set for an exponent of 1.2125. Thus, both magnitude and angle can be determined in a rectangular-to-polar conversion in analog form. The same circuit is shown and discussed in the NONLINEAR CIRCUITS HANDBOOK.



a) Vector Resolution – One Quadrant



b) Vector Resolution – Two Quadrants

Figure 61. Vector operations