Wireless World Circard

Series 4: A.C. measurement

Three main categories of a.c. measurement exist—peak, mean and r.m.s. values of the alternating waveform. Card 1 shows the relation between mean and peak circuits—a moving coil meter simply responding to the mean value and peak values being stored on a capacitor across the circuit load. The simple rectifiers of card 1 suffer from non-linearity at low levels, and it is necessary to use the diode in a negative feedback loop to obtain precise rectification at low levels—see cards 3 and 9, for example.

Another kind of peak detector is worth highlighting if only because of its elegance. The storage capacitor is eliminated by using a high-gain amplifier as a comparator and observing the output change on an oscilloscope or more simply with an l.e.d. Source loading is minimal with this circuit, which can operate at frequencies up to a few megahertz. (Card 6.)

Measurement of r.m.s. values is usually achieved in inexpensive equipment, such as the ubiquitous multimeter, by measuring mean or peak values and calibrating them in r.m.s. values. For sinusoidal waves, the relation $V_{\mathrm{rms}}=1{\cdot}11V_{\mathrm{mean}}$ holds, and some circuits can therefore use one meter scale for both mean and r.m.s. values by using a switched scaling resistor. But for waveforms of non-sinusoidal shape this technique clearly cannot be used unless the appropriate "form factor" is known. The alternatives are to use a thermocouple, frequently used at r.f., a square-mean-square root circuit (card 11), and an analogue multiplier (also card 11).

A.C. Measurements

Measurement of direct voltages is straightforward. A moving-coil meter has good linearity of deflection against direct current in the meter, and the use of parallel and series resistors (shunts and multipliers) allows such meters to give full-scale readings to cope with a wide range of voltages and currents. For very small direct voltages and currents, d.c. amplifiers may be interposed between source and meter, and such amplifiers may also be used to optimize the input resistance of the system, i.e. to minimize loading effects.

For a.c. signals the biggest difficulty can be deciding which parameters of the signal to measure—mean, peak or r.m.s. for example. The issue is further complicated by the need to cope with a range of frequencies so broad that, for example, techniques suitable for high-frequencies result in impossibly long measurement times at very low frequencies.

There is a dearth of sensitive, accurate and low-cost types of meter movement capable of responding directly to a.c.; moving-iron instruments for example require much higher power for a given deflection than moving-coil instruments of comparable quality, while the deflection is a non-linear function of the current being measured. Hence in most cases the a.c. waveform is first processed in such a way that a reading may be obtained on a d.c. meter, which reading is proportional to a desired parameter of the waveform. A basic process employed is that of rectification, where the output voltage (or current) is limited to one polarity regardless of the input.

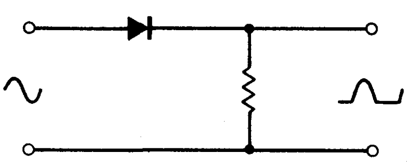
Half-wave rectification (Fig. 1) gives an output which is ideally equal to the input when the latter is positive, and an output which is zero when the input is negative. The ideal diode would pass zero current for all conditions when the anode is negative with respect to the cathode, and have zero p.d. when the polarity is reversed. In practical circuits, while the former ideal is closely approximated to by modern silicon diodes, the diode p.d. in conduction is around 0.5 to 0.8V. The output waveform becomes progressively more distorted as the amplitude of the input voltage is reduced, and for inputs below one volt the output is negligible, i.e. accurate rectification is particularly difficult at low amplitudes. Some improvement is possible by the addition of a second diode biased in such a way that the rectifying diode is brought to the edge of conduction prior to the appearance of a signal.

If a moving-coil milliammeter is placed in series with the load resistance, then the meter current becomes proportional to the average value of the half-wave rectified voltage, provided the frequency is high enough to overcome needle vibration. Such a reading is half that due to a full-wave rectified voltage for symmetrical waveforms such as sine, square and triangular waves. An average reading may also be obtained by feeding the rectified voltage through a low-pass filter to eliminate the a.c. component. Such a modification is necessary where the direct voltage is to be monitored by a digital voltmeter, to provide a digital reading of the mean value of the rectified input.

A direct voltage may be obtained directly as in Fig. 2. The capacitor charges on each positive peak of the input, losing some of that charge between peaks into the resistance of any load. To minimize such losses and make the output a more accurate measure of the repetitive peak input voltage, the time-constant is made much longer than the period of the input signal. Too great a ratio will not allow the capacitor voltage to decay sufficiently rapidly to observe any decay in input peak voltage that may occur during the measurement. Again real diodes introduce a forward-voltage drop that mitigates against accuracy for small inputs.

Full-wave rectification is necessary where the negative and positive portions of the wave may be different. A secondary advantage can be that for symmetrical waves, a full-wave peak detector has its capacitor charge restored twice per cycle, i.e. the time for discharge and hence the ripple is approximately halved. As for half-wave rectifiers, the full-wave circuits could be used for indicating mean or peak values. (The latter would indicate only the largest peaks for an unsymmetrical signal.)

Two methods are available. Bridge rectification as in Fig. 3 requires four diodes to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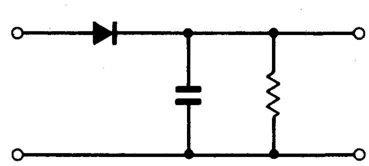
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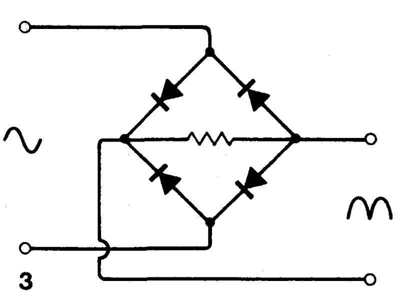
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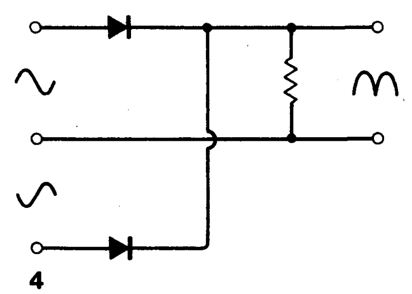
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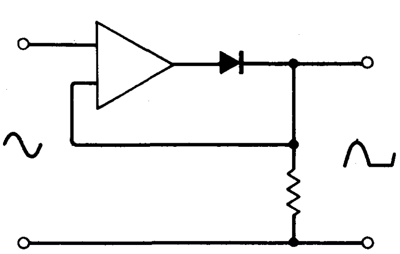
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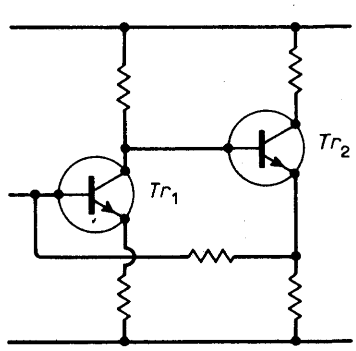

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Fourth set of Circards illustrates techniques of peak, mean and precision rectification. Half-wave, 1 and 2, and full-wave circuits, 3 and 4, can be used to give either mean or peak measurements. Errors due to diode voltage drops can be reduced by putting the diode in a feedback loop, 5, but use of an amplifier limits h.f. accuracy, avoided by using a simple amplifier, 6, with bridge rectifier/meter in the feedback path.

channel current through a load in a given direction regardless of the polarity of the applied potential. Alternatively, the provision of equal but anti-phase drives to a pair of diodes again gives single polarity to the load with each diode contributing on alternate half cycles—Fig. 4. The anti-phase voltage may be provided by a transformer or by an inverting amplifier.

In the above the assumption has been that the rectified waveform would be applied to a measuring device such as a moving-coil meter. Waveform distortion short of that causing significant meter reading error is then unimportant. Where it is required to retain full information on the rectified waveform then a precision rectifier has to be devised, i.e. one in which the rectification process is not burdened by the large errors due to diode voltage drops. Placing the diode(s) in the feedback path of an amplifier allows the effect of the diode p.d. on the output to be reduced by any desired amount.

Fig. 5 shows one version of a precision half-wave rectifier in which, for positive going inputs, the amplifier output is driven positive until it causes the diode to conduct and forces the output voltage to equal the input (or rather to differ from it by a very small p.d. which includes the amplifier offset voltage and a small contribution given by the diode p.d. divided by the amplifier open-loop gain).

The basic circuit shown meets the precision requirements, and in addition minimizes source loading while being capable of supplying normal operational amplifier currents to the load. Many variations are possible leading to: precision half- and full-wave circuits, alternatively known as absolute-value circuits; precision peak detectors and mean-reading circuits.

The use of amplifiers imposes a limit to the upper frequency of operation, which limit is accentuated by the non-linear nature of the circuitry, e.g. the amplifier slew-rate limitation defines the minimum time taken to switch the diode from its non-conducting to conducting state. The precision of the rectification process is more difficult to achieve at higher frequencies and many circuits accurate to a few millivolts at 100Hz are seriously in error at 10kHz. Similar limitations are apparent in any negative feedback system having non-linear elements in the feedback path.

For very high-frequency applications one solution is to construct suitable high-frequency amplifiers of standard design and incorporate a bridge rectifier/meter combination in the feedback path. The simpler designs using the minimum number of transistors are based on circuits such as the d.c. feedback pair of Fig. 6 with the meter circuitry either between Tr_2 collector and Tr_1 emitter, or between Tr_2 emitter and Tr_1 base. Alternating-current coupling of the input signal is then necessary as the direct input voltage cannot be zero in this circuit. The method can be extended to multi-transistor circuits and the feedback network can be located to increase or decrease the input impedance. The lowest frequency of operation is dictated by the largest value of capacitors used, and by the degree of damping of the meter movement.

To extend the frequency downwards, peak detection is usually used, i.e. with a large capacitor to store the peak voltage and minimal discharge current for the period between peaks.

At very low frequencies ($\ll$1Hz) an alternative method is the use of an integrator during a single complete half-cycle or cycle with separate measurement of the time to allow determination of the mean value of the waveform during that cycle.

The amplitude of an a.c. waveform is most frequently quoted in r.m.s. (root mean square) terms, i.e. the instantaneous voltage or current value is squared, the mean value over a complete cycle (or half-cycle) is taken and the square root of that mean value is obtained. It is the r.m.s. value of a voltage that allows calculation of the mean power dissipated in a resistive load, as the power in a resistive load due to an a.c. waveform of V in r.m.s. terms is identical to that due to a direct voltage of V.

It is common for instruments which truly measure the mean rectified or peak values of waveforms to have scales calibrated in terms of the corresponding r.m.s. value for a sine-wave. Hence for non-sinusoidal waveforms the readings fail to give a correct measure of either r.m.s., mean or peak, except where power measurements are concerned, e.g. power fed to a loudspeaker. There is considerable advantage in calibrating the instrument directly in terms of the parameter measured, though this set of Circards includes examples of instruments which incorporate such form factors. True r.m.s. meters are a very different matter. Three common classes depend on

- thermocouples generating an e.m.f. dependent on the power dissipated in a load
- non-linear amplifiers approximating to square-law characteristics where the output can be averaged to give a mean-square reading. A second squaring circuit in the feedback path of a following amplifier gives a square-root action
- multipliers in which the output is proportional to the product of two inputs; if the voltage to be measured is simultaneously fed to both inputs, the output is again proportional to the square of the input.

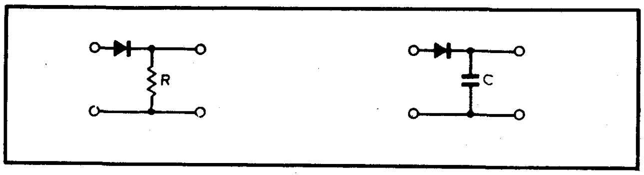
The first method is applied to r.f. signals where the power available is sufficient, and where the use of amplifier/rectifier combinations would introduce errors because of frequency limitations. It is a specialized field and depending as it does largely on the transducer is not covered in this series. The second method requires careful control of the non-linear characteristics for high accuracy to minimize all terms other than second-order; the networks are often obtained as ready-made units from the makers of instrumentation amplifiers. Methods using the square law characteristics of f.e.ts belong to this general class.

The third method can be achieved by using the logarithmic characteristics of semiconductor p-n junctions and by combining several junctions so that their p.ds may be added and/or subtracted functions of the form $(\log V_1 + \log V_2 - \log V_3 - \log V_4)$ may be obtained, i.e. outputs dependent on $V_1 V_2 / V_3 V_4$. These circuits can be made the basis of multipliers, or for $V_1 = V_2 = V_{in}$ and $V_3 = V_4 =$ constant, a square-law circuit results. A practical example is included that allows a meter reading proportional to the mean square of an alternating voltage, i.e. a meter that can be calibrated linearly in terms of the power delivered by that voltage to a given load.

Wireless World Circard

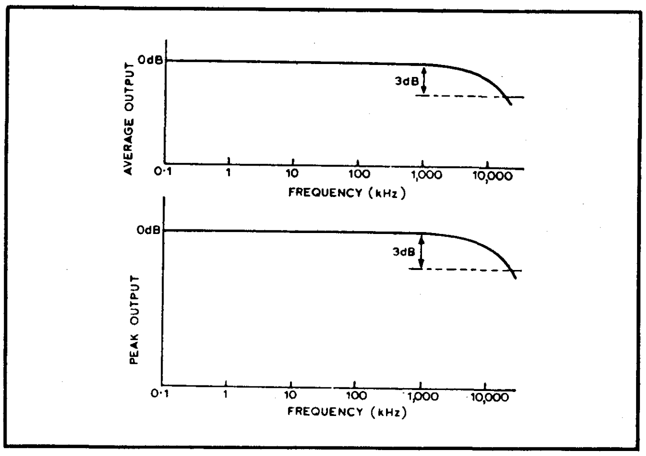
Series 4: A.C. measurement 1

Basic diode rectifiers



Typical performance

Circuit left:
R: 1kΩ; diode: PS101
Input signal level:
3V r.m.s.
Source impedance: 50Ω
Useful frequency range:
up to 18MHz; onset of distortion occurs around 2MHz
Circuit right:
C: 56nF; diode: PS101
Useful frequency range:
up to 25MHz



Circuit description

The basic forms of half-wave diode rectifier are the mean and peak circuits shown above and suffer from the flaw that a minimum potential difference must be developed across the diode itself (0.6V for silicon, 0.4V for Schottky barrier diodes). The only limit to the h.f. response is that of the diode itself, and possibly the source impedance, the graphs above being typical. The transfer function of V_{out}/V_{in} for the resistive load is shown over, being broadly linear, but with no output until V_{in} exceeds 0.6V. If the output is to be read on a moving-coil meter in the upper circuit then the inertia of the coil ensures that the reading is that of the average value of a half-wave rectified signal. At low frequencies the meter needle will vibrate, preventing accurate readings. Typically, readings are adequate to lower audio frequencies. For the peak circuit the continuous d.c. output should be fairly close to the positive peak value of the input, provided the capacitor does not discharge significantly between positive peaks. Hence the time constant comprising C and the effective load resistance (e.g. meter) must be long compared with the period of the input waveform, e.g. a 100-μA meter movement would allow a 1-μF capacitor to decay by approximately 1V in 10ms, corresponding to a 100-Hz mean signal frequency.

Component changes

- Use Schottky diode (HP 2800) to reduce forward voltage drop.
- Buffer peak detector with a voltage-follower to avoid loading the capacitance. This will also mask the effect of moving-coil meter inductance if this becomes a predominant feature (centre, below)
- If frequency performance not important use germanium diode for lower forward voltage drop.

Circuit modifications

Simplest means to eliminate forward voltage drop is to mechanically offset a moving-coil meter, but this does not eliminate non-linearity.

- Linearity at the low level improved slightly with superposed d.c. offset from a power supply or diode connected as shown right. Further improvement by placing D_2 by the collector-base junction of a germanium transistor, but this still does not approach the performance of a diode in the feedback loop of an op-amp.
- For resistive load, connect diode in shunt with load—this is a more suitable arrangement for rectifying a current source.

Further reading

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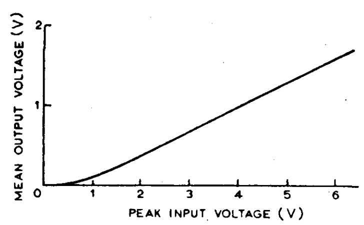
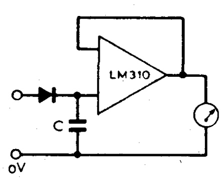
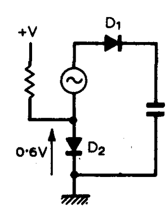
Stewart, H. E. Engineering Electronics, Allyn and Bacon, 1969. pp.130–40 and 738–44.
Hemingway, T. K. Circuit Consultant's Casebook, Business Books, 1970. pp.179–95.

Cross reference

Series 4 card 2, 5, 10



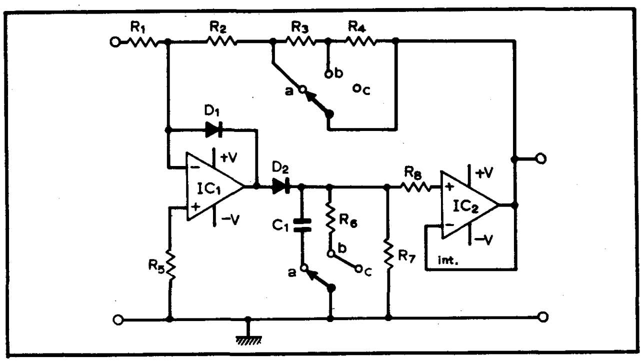




Wireless World Circard

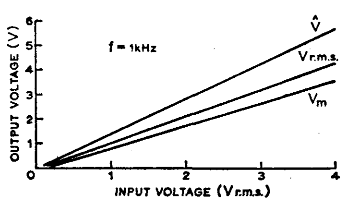
Series 4: A.C. measurement 2

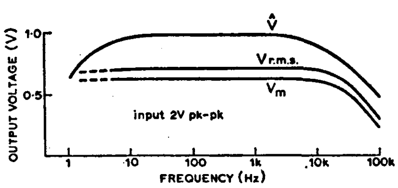
Peak/mean/r.m.s. calibrated rectifier



Components
IC_1: LM301A*
IC_2: LM302
Supplies: ±15V
R_1, R_2, R_3: 100kΩ
R_4: 22kΩ; R_5: 47kΩ
R_6: 6.8kΩ; R_7: 100MΩ
R_8: 2.2kΩ; C_1: 68nF
D_1, D_2: 1N914
* needs 30pF compensating capacitor

Typical performance at 1kHz (all ranges)
Input res.: 100kΩ
Output res.: <5Ω
Load current: 0–10mA
Stability: <0.5%
(V_s ±7 to ±15V, V_{in} >250mV r.m.s.)
Accuracy: ±0.5%, ±errors in R_1 to R_4





Circuit description

The second i.c. is used as a buffer to transfer the rectified output to the load with unity gain and without the load or feedback network presenting any adverse effect to the rectifier. When D_2 conducts, the feedback path is closed and the high gain of IC_1 results in a virtual earth at its inverting input. The output voltage during this period is thus an accurate (inverted) multiple of the input. For switch position (a), the amplifier gain is −1 when diode D_2 is conducting, given that $R_2 = R_1$. Hence the capacitor charges to a positive voltage almost equal to the negative peak input. For all other inputs the output of IC_1 reverse biases diode D_2 and the capacitor stores the peak voltage, which value is transferred to the output by IC_2. The time constant chosen is a compromise between the need for accurate storage of long-period inputs and the need for the circuit to be able to respond to a lower input amplitude in a reasonable period of time. During the period when D_2 is not conducting, D_1 is used to clamp the output of IC_1 by feedback action; this minimizes the recovery time of the circuit prior to the next period when C_1 is to be charged. Mean reading is achieved by removing C_1 and doubling the value of feedback resistance. If the output is fed to a moving-coil meter the reading is twice the mean rectified half-wave value i.e. equal to the mean rectified full-wave input, assuming a symmetrical waveform. Switch position (c) increases the gain of the amplifier in the ratio r.m.s.: mean rectified value for a sine wave. The meter, though deflecting in proportion to the mean rectified value, is now calibrated in terms of the r.m.s. value of the input.

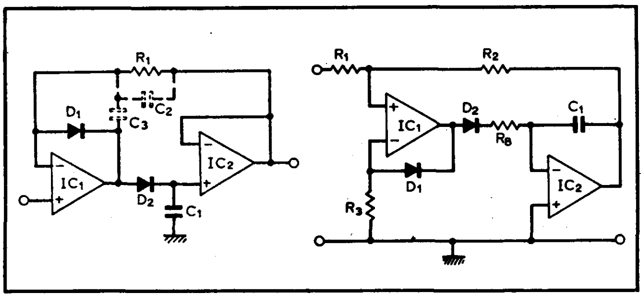
Component changes

Varying R_1 from 1kΩ to 1MΩ gives proportional change in input resistance and input voltage required for given output.

- Increase C_1, R_7 to allow peak detection for lower frequency inputs e.g. for $C_1 > 1\mu F$ peak rectification possible for signals of frequency < 1Hz.
- For true mean-value half-wave rectified let $R_3 = 0$. To retain r.m.s. equivalent increase R_4 to 120kΩ.
- Replace IC_1 by any general-purpose op-amp (741 or 748 with 30-pF compensation capacitor). Replace IC_2 by 310 (improved voltage follower) removing restriction on supply voltage minimum imposed by 302, and further increasing input resistance of stage. For reduced cost, substitute source/emitter follower, checking that reverse breakdown on stage input cannot be exceeded. Direct current offset/drift in follower have no effect on performance provided that any changes are slow i.e. not within one cycle.

Circuit modifications

- Input signal may be applied to the non-inverting input of IC_1. This greatly increases the input impedance at the expense of introducing common-mode input voltages, usually with some worsening of high frequency performance. Addition of capacitors C_2, C_3 in conjuction with R_1 may be necessary with some combinations of amplifiers to avoid risk of high-frequency oscillation. For $R_1 = 100k\Omega$ C_2, C_3 may be around 100pF. (left)
- Using IC_2 as an integrator, the additional inversion provided within the feedback loop must be countered by taking the feedback to the non-inverting input of IC_1. Diode D_1 still provides clamping to avoid saturation of IC_1 and R_8 limits charge rate to C_1. (right)



- In general for all such circuits, corresponding sample-and-hold circuits may be constructed by replacing D_2 by an electronic switch (e.g. f.e.t., c.m.o.s. transmission gate) closed briefly at some desired point on the input cycle.
- In original circuit, using feedforward compensation (see further reading) the upper cut-off frequency could be extended to over 200kHz but with a tendency to unpredictable readings for input amplitudes above 2 or 3V and frequencies above 200kHz.

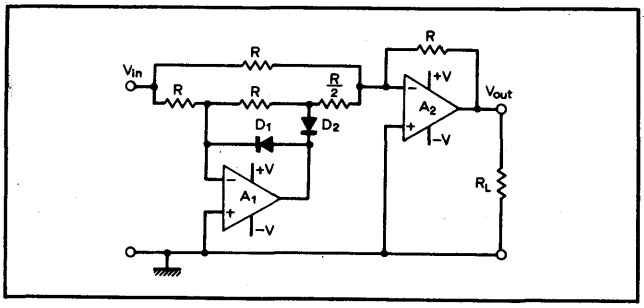
Further reading

National Semiconductor application note AN20, 1969 p.9.
National Semiconductor application note AN31, 1970, p.12.

Cross references

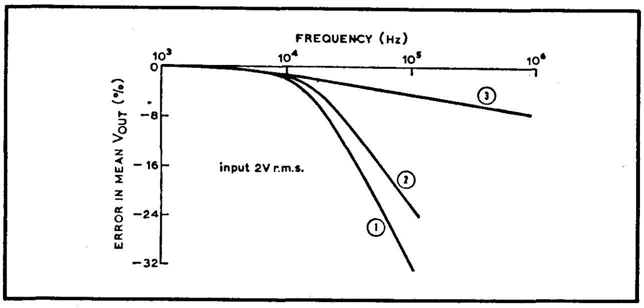
Series 4, cards 1, 3, 5, 7, 9, 10

Absolute-value circuits



Typical performance

A_1, A_2: 741
Supplies*: ±15V, ±5mA
Diodes: PS101
R: 10kΩ±5%
Source res.: 60Ω
L.F. error in V_{out} +4%
*Useful range: ±3 to ±18V



Circuit description

This form of precision rectifier uses two parallel paths feeding currents to summing amplifier A_2. For negative input voltages, the output of the operational amplifier A_1 swings positive causing D_1 to conduct and D_2 to be reverse biased. Thus, for this polarity of input voltage, there is no contribution of current to A_2 through its $R/2$ inverting input path. The only input current to A_2 is therefore $-V_{in}/R$ causing V_{out} to be an inverted (positive) version of V_{in}. For positive input voltages, the signal fed to A_1 causes its output to swing negative which reverse biases D_1 and brings D_2 into conduction. Amplifier A_1 thus acts as a unity-gain inverter causing the voltage at the junction of D_2 and $R/2$ to be $-V_{in}$. Amplifier A_2 therefore receives the sum of two input currents having values of V_{in}/R and $-2V_{in}/R$. The resultant current at the input to A_2, and in its feedback resistor, is $-V_{in}/R$ which therefore makes $V_{out} = V_{in}$. Hence for any input signal V_{out} will be equal to its magnitude or absolute value. Tolerance of the resistors in A_1 are critical if accurate reversal of the gain of the system is to be achieved because for positive input signals the current fed to A_2 represents the difference between those in the two parallel paths. Slew-rate limiting of A_1 for positive-going inputs results in a different amplitude-frequency response to that obtained with negative-going input signals where only the resistive parallel path is relevant. This imposes a separate limit from the slew-rate limitation of A_2 causing the outputs to have different magnitudes for positive and negative inputs.

Component changes

Replacing PS101 diodes with Schottky barrier diodes (e.g. 042–82HP–8211) typically produces amplitude response shown in curve 2. Low-frequency error in V_{out} (mean) is 3.3%. Using Schottky diodes with 741s replaced by 301s with feed-forward compensation as shown below (left) typically produces response shown in curve 3, whose low-frequency error in V_{out} (mean) is +2.2%.

Circuit modifications

● Middle circuit shows a precision rectifier that uses five resistors of the same value which makes their matching somewhat easier than with other circuits. For positive inputs A_1 output goes negative so that D_1 conducts and D_2 is reverse biased. Therefore the junction of R_2 and R_4 is at $-V_{in}$ and as A_2 acts as a unity-gain inverter (non-inverting input of A_2 is virtually grounded), $V_{out} = V_{in}$. For negative inputs A_1 output goes positive, D_2 conducts and D_1 is reverse biased. The input current in R_1 now divides between R_3 and R_2 plus R_4 in the ratio 2:1, so that A_1 output (and non-inverting input of A_2) is at $-2V_{in}/3$. Amplifier A_2 now acts as a follower with a gain of $1 + R_5/(R_2 + R_4) = 3/2$ for $R_2 = R_4 = R_5$, hence $V_{out} = -V_{in}$. Thus V_{out} is a full-wave rectified version of V_{in}. The inverting input of A_1 is a virtual earth and may be used as a summing junction for n inputs from current sources or from voltage sources via n resistors.

● Circuit shown right provides an output that is the absolute value of V_{in} when $R_1 = R_2 = R_3 = R_4/2$ and has a high input impedance since the signal source sees the high common-mode input impedances of A_1 and A_2. For positive inputs A_1 acts as a unity gain follower as D_1 conducts and D_2 is reverse biased. Thus $V_x = V_{in}$ and as $R_4 = 2R_3$, $V_{out} = -2V_x + 3V_{in} = V_{in}$. For negative inputs D_1 is reverse biased and D_2 conducts so that A_1 acts as a follower with a gain of 2 making $V_x = 2V_{in}$ and again as $R_4 = 2R_3$; $V_{out} = -2V_x + 3V_{in} = -4V_{in} + 3V_{in} = -V_{in}$. Hence V_{out} is the absolute value of V_{in}.

Further reading

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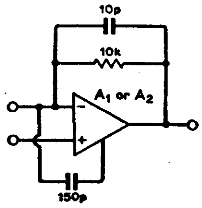
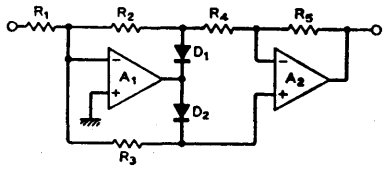
Egan, F. (Ed.), 400 Ideas for Design, vol. 2, Hayden, 1971. pp.152/3.
Graeme, J., Op-amps form self-buffered rectifier, *Electronics* vol. 43, no. 21, 12 Oct. 1970, p.98.
Smith, J., Modern Operational Circuit Design, Wiley, 1972, chapter 6.
Linear Applications Handbook, National Semi-conductor application note AN-31/12, 1972.

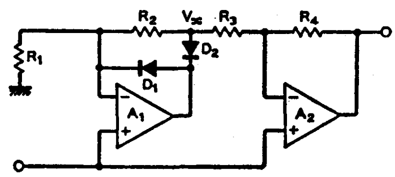
Cross reference

Series 4, cards 2, 7, 9, 12

Circuit modifications

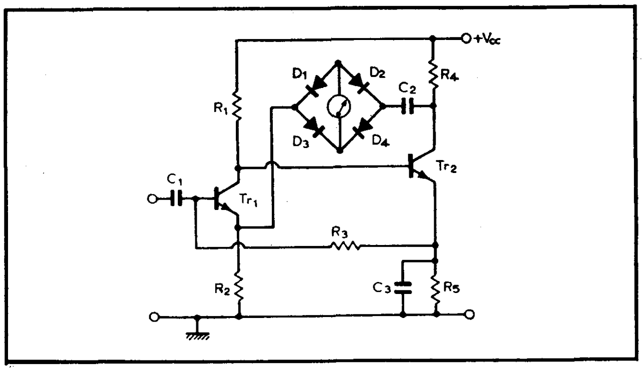






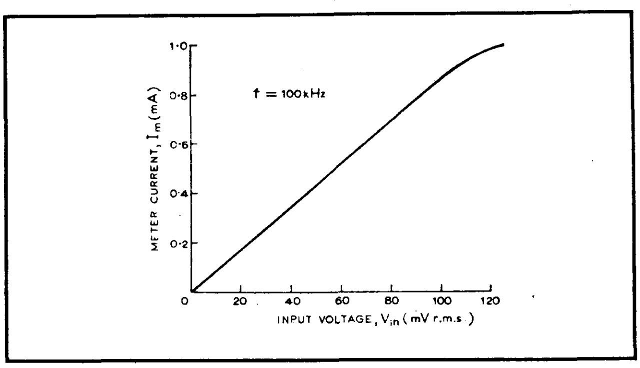
Wireless World Circard

High-frequency voltmeter for a.c.



Typical performance

Supply: +12V, 12mA
Tr_1, Tr_2: 1/5 × CA3046
Diodes: PS101
R_1: 2.7kΩ; R_2: 100Ω
R_3: 39kΩ; R_4: 470Ω
R_5: 270Ω
C_1: 10μF (tantalum)
C_2, C_3: 22μF (tantalum)



Circuit description

The transfer function at this circuit is $G_y = A_y/(1 - \beta_z.A_y)$ where the transadmittance A_y is ideally defined by the desired full-scale meter current for a given value of V_{in} and where the feedback factor β_z is defined by R_2. The input signal is inverted by Tr_1 and again by Tr_2 before being applied to the bridge rectifier through C_2 which removes the d.c. error that would arise from the collector voltage of Tr_2. Meter current flows in R_2 which with perfect follower action would cause the p.d. across it to equal V_{in}. However R_2 also carries the emitter current of Tr_1 which is not a perfect follower. Therefore the choice of R_2 to make the meter read V_{in} (r.m.s.) directly, for a given full-scale meter current, will be less than the value predicted by using $R_2 = V_{in}/1.11\ I_m$. Overall d.c. shunt-derived shunt-applied negative feedback is provided by R_3 and R_5 is decoupled by C_3. Capacitors C_2 and C_3 cause the amplifier to exhibit a lower cut-off in its response.

Component changes

Meters requiring different full-scale deflection currents can be accommodated by a suitable choice of R_2 for a given V_{in}. For R_2 greater than about 100Ω and a full-scale deflection sensitivity of around 100mV for a 1mA movement, the transfer function is defined by R_2 with a typical full-scale error of about 2mV. Careful printed circuit layout, using a single ground point, is necessary to achieve an extended amplitude-frequency response. To prevent instability, the use of ferrite beads on the supply leads and a tantalum bead decoupling capacitor are recommended. It may be necessary to connect a small capacitor between collector and base of Tr_2 and possibly a resistor of around 100Ω in series with the source.

Circuit modifications

- Replacing the input stage by a long-tailed pair, as shown left, decreases the loading on the feedback resistor R_f and allows the transconductance to approach closer to the ideal value $1/R_f$. If the input may contain a d.c. component, capacitive coupling may still be used to remove it from the amplifier input with separate resistors to return the input base to ground potential. The resistors are tapped and driven by a capacitor from R_f, the bootstrapping effect reducing the alternating current in the resistors allowing lower d.c. values for good bias-point stability but without lowering the input impedance.

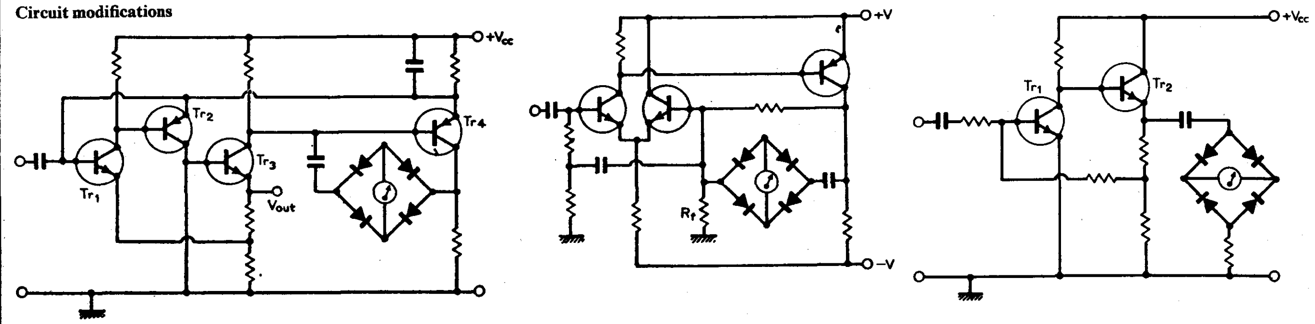
- Shown centre is a three-stage amplifier with overall series-applied shunt-derived negative feedback that raises the input impedance, fixes the voltage gain and provides a well-defined current into the fourth transistor. This has a bridge-rectifier giving shunt-applied feedback for low input impedance ensuring that the a.c. component of the collector current of Tr_3 is diverted into the meter. The emitter of Tr_4 provides a convenient point from which to derive overall shunt-applied d.c. negative feedback, to stabilize the operating conditions. In addition the emitter of Tr_3 provides an amplified voltage output for waveform monitoring. Any loading increases the meter current for a given signal.

- If some non-linearity can be tolerated while maximizing the frequency response of a meter rectifier it is possible to remove the non-linear elements from the feedback loop and drive them from any r.f. amplifier. If the amplifier is designed to have a high Z_{out} rather than low, as shown right the non-linearity is minimized.

Further reading

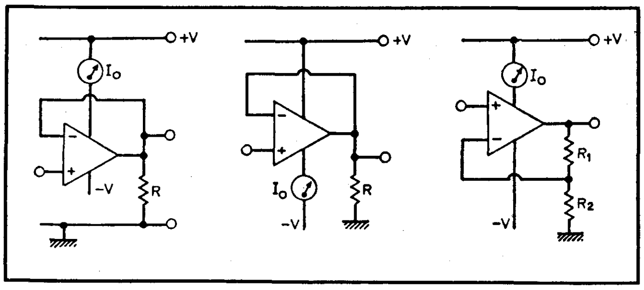
Clayton, G., "Operational amplifiers", *Wireless World*, vol. 75, 1969, pp.482/3.
Application of Linear Microcircuits, vol. 1, SGS-UK Ltd, 1968, p.90.

Circuit modifications



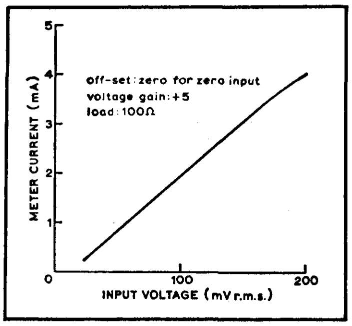
Wireless World Circard

Class-B economy rectifier



Typical performance

IC: 741
Supplies: $\pm 6V$
R: 100Ω
Meter: 5mA f.s.d.
V_{in}: up to 1V r.m.s.
Sensitivity: 4.2mA/V
Quiescent current: 1.3mA
Upper cut-off frequency: 90kHz



Circuit description

An ideal half-wave rectifier conducts for precisely one half-cycle of the input i.e. conduction angle is 180°. Such a conduction angle defines class B operation in an amplifier and suggests a description of the rectifier as a class-B single-ended amplifier of unity voltage and current gains. Conversely the analogy suggests the use of any class-B amplifier to provide an output proportional to the positive or negative parts only of the input i.e. rectification with amplification in which neither voltage nor current gain need be restircted to unity. The example given uses the most widely available operational amplifier, and suffers from a number of disadvantages which are obviated by designing the amplifier or the output stage for this particular purpose. Any standing current in the amplifier affects the reading in two ways: the meter reading for zero input is finite (1 to 2mA for circuit shown over, left) and requires scale-changing or mechanical offset. If the current is in the output stage i.e. it is operating in class AB, then for small input signals the current remains substantially constant. In practice the peak current obtainable is limited to ~ 25mA and the current in the output stage may be ~ 1mA. This latter current ensures that the supply current changes little for signals up to 5–10% of maximum. The circuit has a number of advantages to offset these limitations: the input impedance is very high; the circuit is uncritical of supply voltages, lower values minimizing dissipation and consequent change in meter readings; sensitivity is easily adjusted; amplitude-frequency response up to 500kHz with suitable low-cost amplifier; output of amplifier available for oscilloscope monitoring.

Component changes

Useful range of R: 50 to 500Ω
Useful range of supplies: $\pm 5V$ to $\pm 15V$
There is no advantage to increasing negative supply if meter is in positive line and vice-versa. Use minimum supply voltages possible to minimize heating effect, which changes standing current in class AB stage. Any other feedback configuration can be used such as follower with gain, see-saw amplifier. Reduction of feedback increases sensitivity and allows reduced value of compensating capacitor with op-amps such as 301, 748. Bandwidth may be increased to 500kHz with gain of 10 using compensating capacitor of 3pF for 301 type op-amp. Bandwidth is already improved over some circuits, since small output voltage swing minimizes slew-rate limitations.

Circuit modifications

● The output stage of 741-type op-amps is basically a complementary pair of emitter followers. Driving current into the output uses the transistors instead as common-base stages, i.e. with unity current gain, but developing any desired unipolar output voltage across the meter (circuit left). Effective input resistance → 0; same limitations on minimum signal registered as before.

● Using an existing amplifier, add a separate emitter follower to the output, monitoring the current only in the collector of one transistor. A second transistor may be added to retain both polarities of output and present undistorted waveform to oscilloscope if required, and to maintain closed feedback loop, avoiding saturation on negative-going inputs (middle circuit).

● Alternately add clamping diode if amplifier has suitable access point, normally at base of one output transistor (right circuit). To avoid class A operation, which prevents detection of small signals in this mode of operation, while retaining linearity the novel approach to class B proposed by Blomley is indicated (see Further reading).

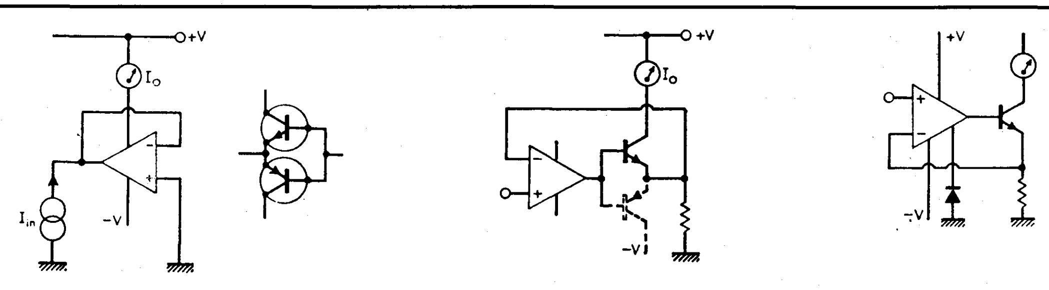
Further reading

National Semiconductor application note AN31-11, 1972.
Blomley, P., New approach to class-B amplifier design, *Wireless World*, vol. 1971, pp.57–61 and 127–131.

Cross references

Circircard series 4, cards 1, 2 & 8.

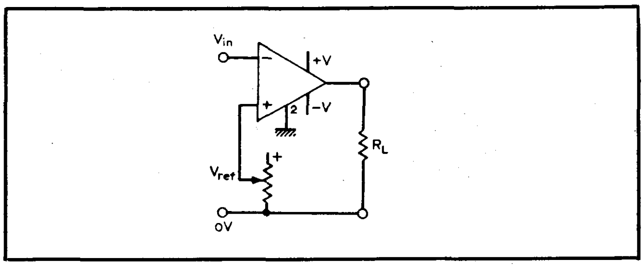
Circuit modifications



Wireless World Circard

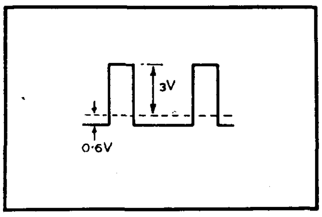
Series 4: A.C. measurement 6

Potentiometric peak-sensing circuit



Typical performance

IC: SN52710
Supplies: + 12V, −6V
P.R.F.: up to 2 MHz
Minimum pulse width: 200ns rise time: 20ns
Nominal voltage: + 4V
Detectable pulse height; ≈ 4V



Circuit description

A comparator is a high-gain amplifier specifically designed for minimum response time. It is therefore a good choice for detecting specific amplitudes of a short-duration signals or pulses. If one input of the comparator is biased to a suitable reference level, then for all input signals below that reference, the output rests in its low state. When the input amplitude exceeds the reference by a small amount, the high gain of the amplifier (typically > 1000) causes the output to change state This change may be observed on an oscilloscope, and input pulses of short duration can therefore be detected. As the input current is low, this method of detection imposes the minimum loading on the source. By reversing polarity of the reference, opposite polarities of input signals can be observed and combinations of such comparators may be used, as with the window comparator of card 11, series 2. As the input current changes at the non-inverting input, is finite, though small, the source resistance of the reference should be low enough that the reference itself does not change during the process. The output voltage change is constrained by the design of the comparator used to be within the range suitable for driving circuitry such as t.t.l. stages.

Component changes

- Useful range of R_L: 10k to 100Ω.
- Same principle can be applied to any other comparator (e.g. 311) to give equal positive or negative output states (over, right).

Circuit modifications

To obtain a square-wave output, drive a divide-by-two t.t.l. bistable. Load the comparator output with 220Ω to ensure a sufficient current sink in the off condition. A moving-coil meter or light-emitting diode may be used to give a visual indication that the pulse has reached the reference level (circuit left).

- Introduce positive feedback to improve switching speed but at the expense of switching level, as shown centre. The diode ensures that the positive feedback only affects one of the switching levels. If it is arranged that the on-level is unaffected, the switch-on will occur at an accurately-controlled amplitude, but the circuit does not switch off for small transients or ripple on the pulse and only when the pulse has fallen by a defined amount.
- For the circuit shown right and with the reference voltage positive an input that is greater than the reference will cause the output voltage to be positive. When the input falls below the reference the output will be negative. Resistor R_1 has to be less than R_2 for small hysteresis, but R_1 should not be so small as to load the source.

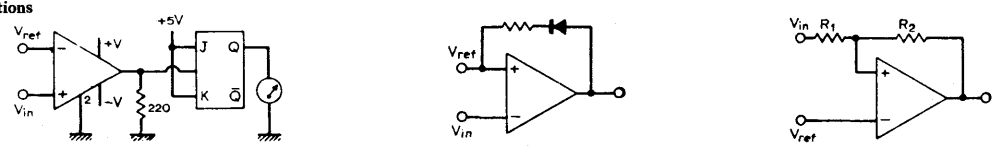
Further reading

Application of linear microcircuits, SGS 1969, p.68.
National Semiconductor application note AN41, 1970.

Cross reference

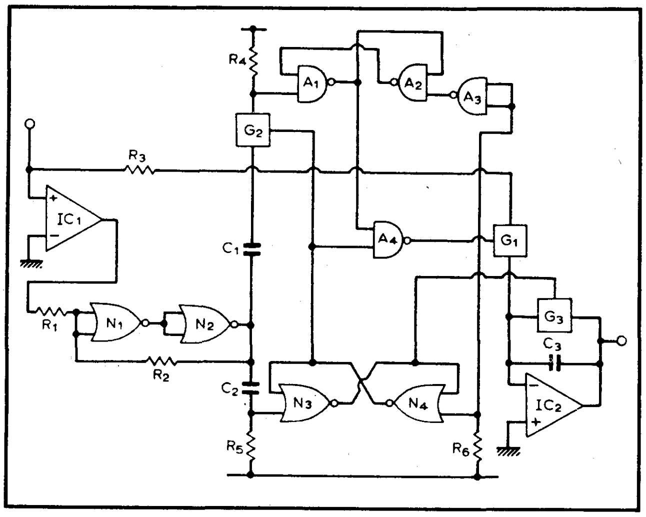
Series 2 card 6.
Series 4, card 10

Circuit modifications



Wireless World Circard

Low-frequency measurement of a.c. waveforms



Typical performance

IC_1, IC_2: 741
A_1 to A_4: CD4011AE
N_1 to N_4: CD4001AE
G_1 to G_3: CD4016AE
C_1, C_2: 1nF; C_3: 4.7nF
R_1: 3.3kΩ; R_2: 120kΩ
R_3: 150kΩ
f: 0.9Hz

V_{in}: 8V pk-pk
Waveform: square, sine. triangle
Output: 1.00V, 0.72V, 0.51V respectively
Stability: <1% change for V_s ± 4 to ± 7.5V

Circuit description

Circuit is used to determine the integral of an input waveform over the duration of the first complete positive period following the receipt of a reset pulse. It does this using IC_1 as a comparator and NOR gates N_1, N_2 as a schmitt trigger circuit to generate pulses via C1, C2 at each zero-crossing of the input. Gate G_1 is opened and G_2 closed, discharging C_3 and leaving the initial output of IC_2 at zero. The first positive-going zero-crossing acts via C_2 and the RS (set-reset) flip-flop composed of NOR gates N_3, N_4 to open G_3 and close G_1. The input is then integrated by IC_2 until a negative-going zero-crossing changes the state of the NAND gate RS flip-flop A_1, A_2. This opens G_1 and no further integration takes place, while leaving G_3 open so that the final integral can continue to be read. Gate G_2 is used to suppress the negative-going step that may occur prior to the start of an integrating period, as when the original reset pulse is fed in during a positive period. The non-zero input current of IC_2 together with leakage effects in G_1, G_2 cause the output voltage to drift and readings should be taken as soon as possible after the end of the positive period normally a single half-cycle of a repetitive waveform. Similar circuits can be produced using all NAND or all NOR elements but with no reduction in package count. Normal gating techniques could be used in place of G_2 but in the present version it avoided the use of a further logic circuit.

Component changes

IC_1: Any operational amplifier/comparator as speed is not critical—low offset voltage an advantage.
IC_2: Minimum input current for lowest drift, e.g. LM308. Alternatively use drift compensation methods.
G_1 to G_3: Any m.o.s. gates; possibly reed switches for minimum drift at low frequencies.
A_1 to A_4, N_1 to N_4: Any RS flipflops, though c.m.o.s. convenient as compatible with gates and op-amps, at V_s ± 5V.
R_2/R_1: 5/1 to 100/1; R_1 > 1kΩ, R_2 < 1MΩ.
R_3: dependent on signal being integrated. Choose for output between 0.5 and 2.5V for given waveform–higher if higher supply available.

Circuit modifications

- Simpler versions of the circuit may be made in which no provision is made for the precautions listed above to avoid integration on succeeding cycles i.e. readings should be taken between cycles or a cumulative answer taken after multiple cycles. Equally, confusion may arise if it is attempted to start during a positive cycle as only a partial integral is achieved. To determine the mean value, time may be measured by any convenient means and the possibility shown is of a second similar integrator fed with a constant voltage. Alternatively the same integrator may be used for a later cycle if the waveform is repetitive.
- Insertion of a good voltage follower to the inverting input of the integrator is an alternative if a low-drift amplifier is not available. Standard drift compensation by feeding a small direct current derived from the positive supply rail.

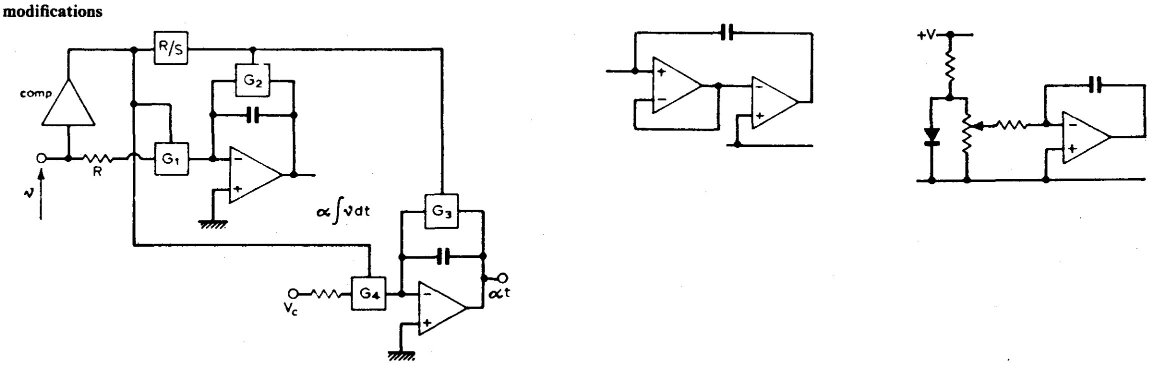
Further reading

Drift Compensation Techniques, National Semiconductor application note AN-3, 1967.
IC Op-amp Beats FETs on Input Current, National Semiconductor application note AN-29, 1969.

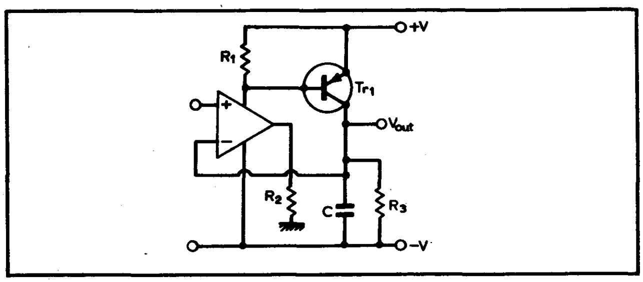
Cross references

Series 4 Cards 2, 3 & 12.

Circuit modifications

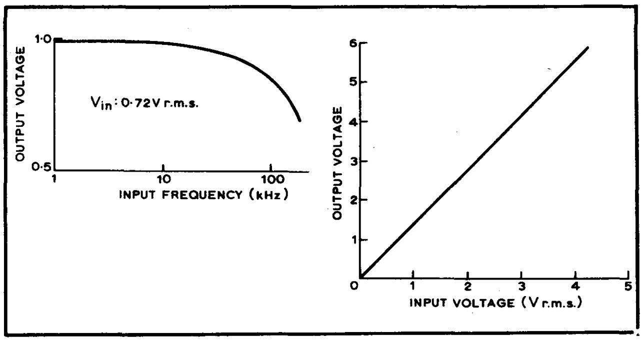


High-current peak/mean rectifier



Typical performance

IC: 741
Tr_1: BFR81
Supplies: ±10V
R_1: 180Ω
R_2: 470Ω
R_3: 1MΩ
C: 0.47μF
V_i: 1.44V r.m.s. at 1kHz
V_0: 2.0V d.c.



Circuit description

The circuit is related to the comparator of card 4, (series 2). In place of a diode to pass or block output signals depending on their polarity a transistor is used which is driven in and out of conduction. depending on the input. This boosts the peak output current available, and the transistor can be driven with relatively low output-voltage swing at the normal op-amp output, minimizing the effect of slew-rate limit in the amplifier. When the input is positive the amplifier output goes positive, the resulting current in R_2 being drawn through R_1. The p.d. developed across R_1 drives Tr_1 into conduction charging C until the potential at the amplifier inverting input increases to match that at the non-inverting input. As the input falls the p.d. between the input terminals reverses its direction and the amplifier output swings negative, the current in R_1 falling to some minimum level insufficient to maintain conduction in Tr_1. For the remainder of the cycle C discharges under the combined action of R_3 and the small input current drawn by the op-amp. Provided the resulting time constant is long compared with the lowest frequency of the input voltage, the peak voltage is accurately retained. Some discharge has to be allowed, so that the capacitor p.d. can decay, when the succeeding measurement is of a signal with lower peak amplitude. Similar circuits can be used for driving low-resistance loads without the shunt capacitor.

Component changes

Replace BFR81 by any general-purpose silicon transistor. For peak current ratings less than 200mA add suitable limiting resistor in series with collector (~ 50 to 100Ω). Op-amp may be replaced by compensated 301, 748 etc. provided resulting p.d. across R_1 falls below level at which Tr_1 conducts when input is zero or negative.
C: 1nF to 1000μF
R_3: 100k to 10MΩ
R_1: 100 to 220Ω
Raising R_1 further is likely to leave Tr_1 conducting permanently; too low a value requires excessive signal drive.
R_2: up to 1kΩ. With s/c the non-linearities are exaggerated; high R_2 reduces overall gain and reduces peak output current.

Circuit modifications

- Replacing the capacitive load by a low-value resistance results in a precision half-wave rectifier. Values of R down to 15Ω may be used with input voltages of around 1V r.m.s. Higher peak currents may be used depending on transistor current gain/peak current rating. A d.c. milliammeter used instead of R reads mean current of half-wave rectified signal i.e. circuit may be used as a.c. mean-reading meter with moving-coil movements of low sensitivity (circuit left).
- For voltage gain R may be replaced by potential divider in usual way. For rectification giving negative-going outputs, an n-p-n transistor can be driven from other amplifier supply line. Voltage gain as shown is $(R_2 + R_1)/R_1$ (middle circuit).
- Inverting action is possible but requires diodes to maintain the output at zero for the input polarity for which an output is not intended. However full-wave rectification is possible with circuits such as that shown right where the input is applied in common to the two inputs with outputs also commoned. For positive-going input IC_1 drives Tr_1 into conduction and any current through resistors R is absorbed. For negative-going input IC_2 similarly drives Tr_2.

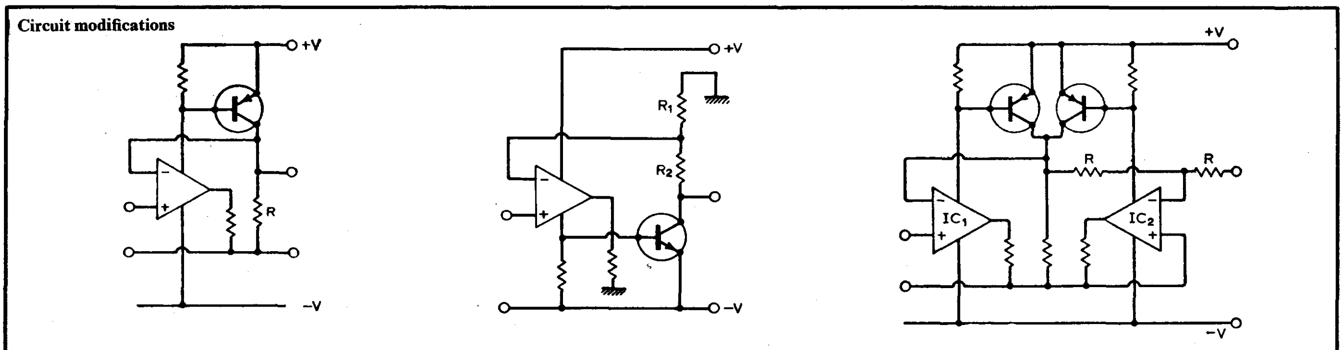
Further reading

Graeme, J. G., Tobey, G. E., & Huelsman, L. P., Operational Amplifiers, McGraw-Hill, p.249.
National Semiconductor Linear Brief LB-8, 1969.

Cross references

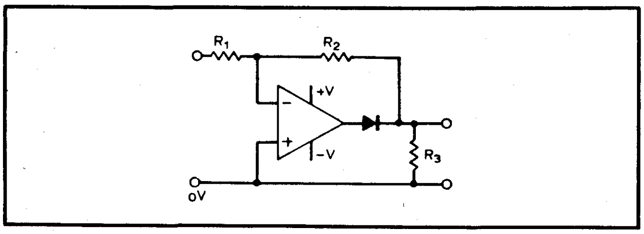
Series 2 card 4.
Series 4 card 5 & 10.

Circuit modifications



Wireless World Circard

Simple precision rectifiers



Typical performance

IC: 741
Diode: PS101
Supplies: ± 15V
R_3: 10kΩ; R_2: 3.3kΩ
R_1: 6.8kΩ
Signal level: 5V pk-pk
Amplitude response: see graphs

Linearity maintained for input signal level down to 0.5V pk-pk. Reduction to 0.2V using compensated op-amp also improving amplitude response.

Useful value of R_2/R_1:

$$V_{out} = \frac{V_{in}.R_3}{R_1 + R_2 + R_3}$$

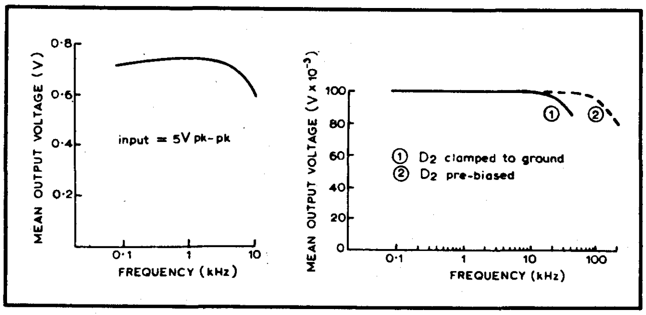
$$= \frac{V_{in}\, R_2}{R_1}$$

$$R_1\, R_3 = R_2\,(R_1 + R_2 + R_3)$$

Let $R_1 + R_2 = R'$

$$R_2 = \frac{R_1\, R_3}{R' + R_3}$$

If $R_3 = R'$
then $R_2 = R_1/2$



Circuit description

This circuit has the advantage that only one op-amp is required, but the load must be maintained constant to preserve the full-wave rectified waveform. When an alternating signal is applied at V_{in}, diode D_1 is alternately forward and reverse biased. When V_{in} is positive with respect to ground the op-amp output is negative, D_1 is non-conducting and V_{in} is developed across R_1, R_2 and R_3 in series. When V_{in} goes negative, the op-amp output is positive, D_1 is forward biased, and V_{out} is then defined by the ratio of R_2: R_1. One ratio of R_2: R_1 to ensure that the alternative positive half-cycles are equal is deduced in the analysis shown. Note that the effect of the diode forward voltage drop is minimized as it is within the feedback loop. As the amplifier supplies the output only during one half-cycle, the amplitude response for this condition and that when only the resistors are in circuit must be different. Further, no mechanism is shown for limiting amplifier saturation for negative outputs i.e. the recovery time from this saturated condition is long, and includes the slew-rate limitation, which may be well below 1V/μs.

Component changes

- Supply voltages can be reduced to ±3V with appropriately reduced signal level.
- Use of Schottky diode (HP 2800) will reduce cross-over peaks.
- R_3 may be altered over a wide range, but the relationship with R_1 and R_2 described above must be maintained.

Circuit modifications

- Reverse diode D_1 to obtain negative voltages.
- Use 301 with feedforward compensation capacitors to improve response; C_1: 15pF, C_2: 150pF.
- Use clamping diode D_2 between pin 8 on 301 and ground to improve low-level performance (circuit left).
- Cross-over troughs on output waveform are minimized by pre-biasing the clamping diode D_2 (middle circuit). About 40% of +15V reduces a trough to 40mV above zero level.
- Variation of load is possible independently of R_2: R_1 using the circuit shown right. Tr_1: BC125, R_1: 10kΩ, R_2: 9.8kΩ. Useful range of R_3: 3 to 10kΩ, R_4: 6.8kΩ. When input is positive diode D_1 conducts, IC acts as amplifier with unity gain When input is negative, D_1 is reverse biased, Tr_1 conducts and signal is applied across the load. Trimming of R_2 necessary to equalize the peaks of V_{out}.

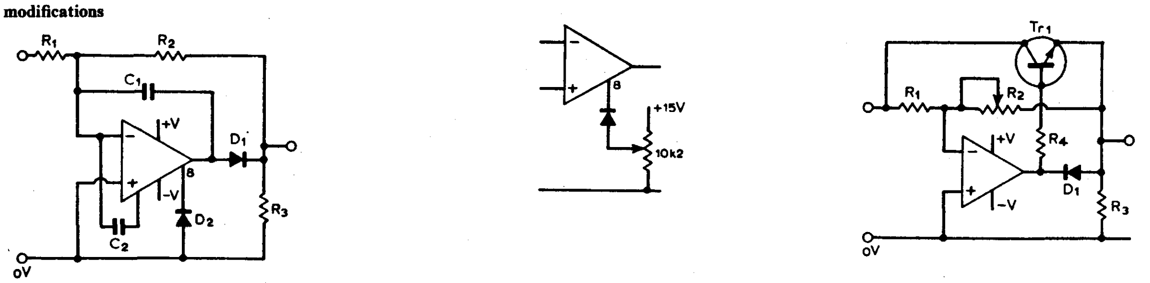
Further reading

Precision full-wave rectifier uses op-amp, in Electronic Circuit Design Handbook. Tab, revised 4th edition, p.302.
Applications Manual for Operational Amplifiers, Philbrick/Nexus Research, 1968, pp.58/9.

Cross reference

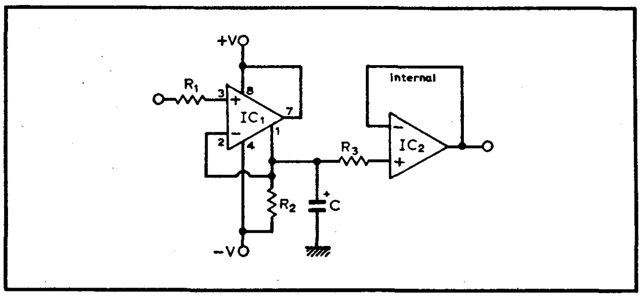
Series 4, cards 2 & 3.

Circuit modifications



Wireless World Circard — Series 4: A.C. measurement 10

Positive/negative peak detector



Typical performance

IC_1: LM311
IC_2: LM310
Supplies: ±15V
R_1: 2.2kΩ
R_2: 1MΩ
C: 1μF (35-V tantalum)
R_3: 10kΩ
Ripple <1% down to 200Hz

Peak detection for inputs <100mV to >20V pk-pk. For supplies of ±10V, max V_{in} for accurate peak detection reduced to 16V pk-pk.
Max frequency >500kHz

Circuit description

The peak detector shown is based on a particular comparator though the method is similar to that described for other peak detectors. The difference lies in the output stage of this comparator which may be considered as equivalent to a switch controlled by the relative potentials of the amplifier input, but where the switch may be effectively floated with respect to the supply lines. While the polarity of the switch p.d. must be defined, it allows either end of the switch to be connected to the appropriate supply point (in the given circuit pin 7 is taken to the positive line with pin 1 as the output; if pin 7 is used as output, pin 1 is taken to the negative line). This change in connection introduces an additional inversion in the loop, equivalent to changing from common-emitter to common-collector output configuration. This necessitates interchange of the input to which the feedback is returned. When the input voltage goes positive the output stage goes into conduction supplying a large current to the capacitor C, which charges until the potential returned to the inverting input matches the signal. As the input signal falls, the output stage cuts off and the capacitor C holds this peak potential until it receives current on any succeeding positive input peak greater than its stored potential. To ensure that the capacitor discharges at a controlled rate and is capable of responding to lower peak inputs within some defined time, resistor R_2 is included. The voltage follower allows normal load resistances to be used, including e.g. moving-coil meters, without changing the time constant of the circuit

Component changes

IC_1: For this particular circuit only comparators having this kind of output stage can be used i.e. equivalents to the LM311 (or the higher specification LM211 and LM111).
IC_2: Any voltage follower including standard op-amps such as 741 connected in voltage-follower mode. The lower input resistance that results reduces the time-constant somewhat but this is often not serious.
R_1: 1.5k to 4.7kΩ.
R_2: 220k to 3.3MΩ; too high a value allows output stage leakage to charge capacitor beyond range at which positive peaks may be sensed; too low a value increases ripple at low frequencies.
C: Determines low-frequency limits in conjunction with R_2. Use of tantalum for higher values ensures that risk of r.f. oscillation on peaks is minimized.
Supply voltage: ±5 to ±18V.

Circuit modifications

- Interchanging the output connections and the output connections allows a negative-peak detector of comparable performance. Note reversal of capacitor polarity. Other buffers such as f.e.ts may be used with the penalty of d.c. offset and drift.
- If other comparators are used which have a ground referred output-voltage swing, they cannot be used directly to charge the capacitor on peaks since they will also discharge the capacitor during the remainder of the cycle. Interposing a t.t.l. inverter with open-collector output adds further gain providing larger charging current than could be provided through a diode. As shown, the input and output are with respect to the ±5V line of the system though this could equally be made the common line of the remaining system, renaming the positive supply as +5V and the most negative supply as −10V. For amplitudes of ±5V pk-pk the circuit provides low-cost peak detector up to several MHz using 710 comparator, one gate from quad-nor t.t.l. package 7401 or inverter type 7404.
- The basic similarity between peak-detectors and sample-and-hold circuits allows an electronic switch to be used to isolate the capacitor at a desired point in the cycle. Location of feedback depends on whether output stage is class A or class B.

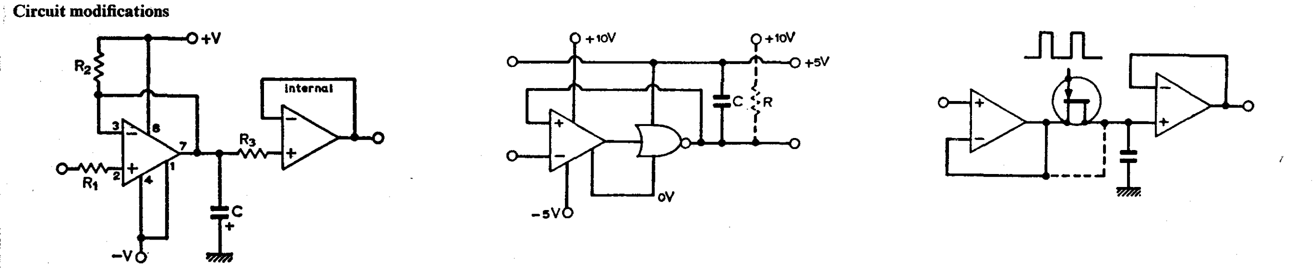
Further reading

National Semiconductor, LM311 voltage comparator data sheet, 1970.
Positive peak-detector for fast pulses, Applications of Linear Microcircuits, SGS, vol. 1, p, 98.

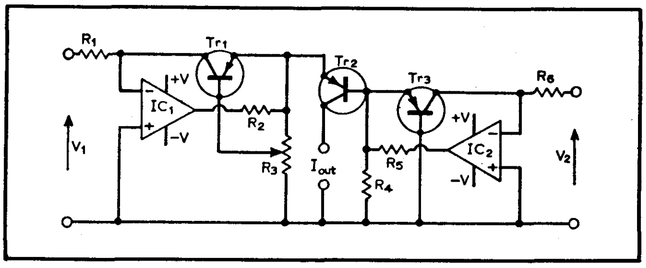
Cross references

Series 4, cards 1, 2, 6 & 8.

Circuit modifications

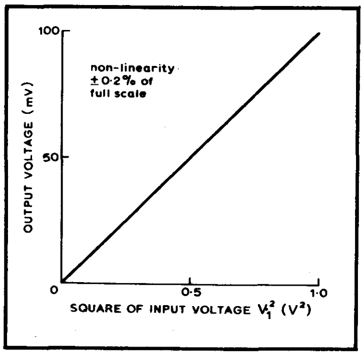


Square-law meter circuit



Typical performance

V_s: ±10V
IC_1: 741
IC_2: 741
Tr_1, Tr_2, Tr_3: matched n-p-n transistors, CA3046 (RCA) using three out of the five transistors in package
R_1: 10kΩ; R_2: 220Ω
R_3, R_4: 100Ω
R_5: 220Ω; R_6: 10kΩ



Circuit description

True r.m.s. and power measurements by purely electronic methods (not depending on devices such as thermocouples, moving-iron meters) can be performed using logarithmic amplifiers. The method is of broad application and it is intended as the subject of a separate series; however an example of a low-cost circuit is included to demonstrate the principle. Devices IC_1 and Tr_1, IC_2 and Tr_3 comprise amplifiers whose outputs are proportional to the logarithm of the input currents. As these currents can be made proportional to voltage sources V_1 and V_2, the p.d. applied between base and emitter of Tr_2 is of the form $\log_e A - \log_e B$ where A and B depend on V_1, V_2 respectively. Collector current of Tr_2 is then proportional to the antilog$_e$ of the p.d. between its base and emitter (see Theory). By tapping the base of Tr_1 onto R_3 the effective output voltage from this first circuit can be made any desired multiple of Tr_1 base-emitter p.d. provided that the base current is much less than the current in R_3. The output voltage fed to the emitter of Tr_2 then becomes proportional to $2\log_e V_1$ or $\log_e V_1^2$ when R_3 tap is set to its centre value. The second logarithmic stage is required for temperature compensation even where V_2 is made a constant reference voltage. All the transistors should be well matched and operate at equal junction temperatures. Circuit is designed to use n-p-n types throughout and these are available in a standard low-cost multi-transistor package. As shown, the circuit deals only with positive-going voltages for V_1 and V_2, but a modification is shown that extends the operation to bipolar form. The second amplifier may also be used to provide power-law action such that the output current becomes proportional to V_1^n/V_2^m with the restriction $(n-m) = 1$ for temperature-compensated operation.

Component changes

IC_1, IC_2: General-purpose op-amps tend to oscillate due to additional gain of transistors in feedback path. Heavier compensation of amplifiers such as 301, 748; shunt capacitance from output to inverting input if speed not important.
R_2, R_5: may be omitted if other means used to avoid oscillation (they reduce loop gain in conjunction with R_3, R_4).
R_4: May be omitted subject to above precautions or may be tapped as with IC_1.
R_1, R_6: 1k to 100kΩ, setting sensitivity of circuit. At both high (≫ 1mA) and low (≪1μA) current transistors depart from log law; op-amp input current limits low-level operation for particular circuit given.

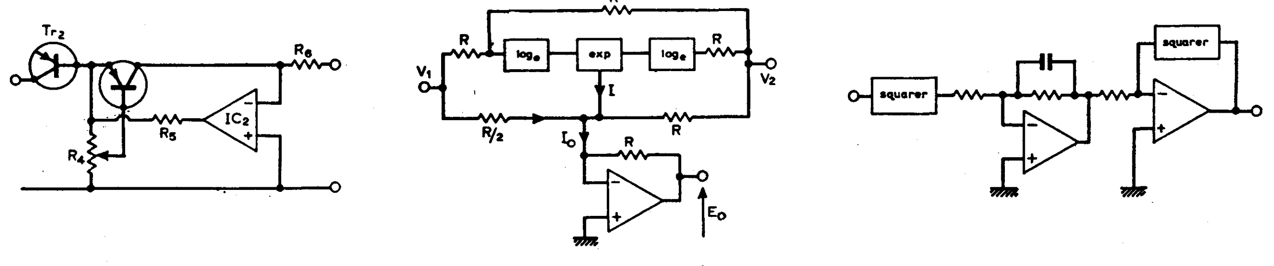
Circuit modifications

- If IC_2 drives the transistor, but with the base of the transistor taken to a tap on R_4, the output of the amplifier is some multiple of the transistor V_{be} i.e. is proportional to a multiple of $\log_e V_2$. This results in Tr_2 receiving a base-emitter p.d. depending on different power laws of V_1 and V_2: $IC_2 \propto V_1^n/V_2^m$. For the single junction of Tr_2 base-emitter to maintain temp. compensation the choice of n, m is restricted by $n = m + 1$.
- The circuit as originally shown cannot accept negative values for V_1. Where it is desired to obtain a true squared output for, say, a sinusoidal input, the modification shown may be used. V_2 is made constant and greater than the highest magnitude of V_1 in the negative sense. Hence both log amplifiers receive positive inputs at all times. By combining various proportions of V_1, V_2 at inputs and output, the output can be made a square-law function of V_1 for both polarities.
- Once basic controlled-function blocks are available, whether as shown or the high-performance blocks available from specialist manufacturers, they may be combined to provide other functions. The input voltage (see right) is squared, its mean value taken and applied to a circuit with a squaring network in its feedback. With feedback proportional to square of output, input signal equalling feedback, the output is proportional to square-root of input. Overall function performed is thus r.m.s. value of input.

Further reading

Ehrsam, B., Transistor Logarithmic Conversion Using an Integrated Operational Amplifier, Motorola application note AN-261.
National Semiconductor application notes AN29-12, AN30, AN31-18 & AN31-20, 1972.

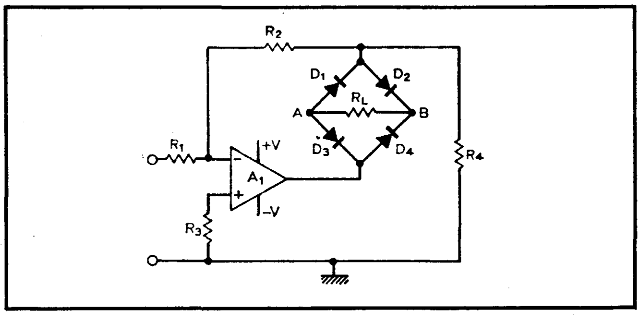
Circuit modifications



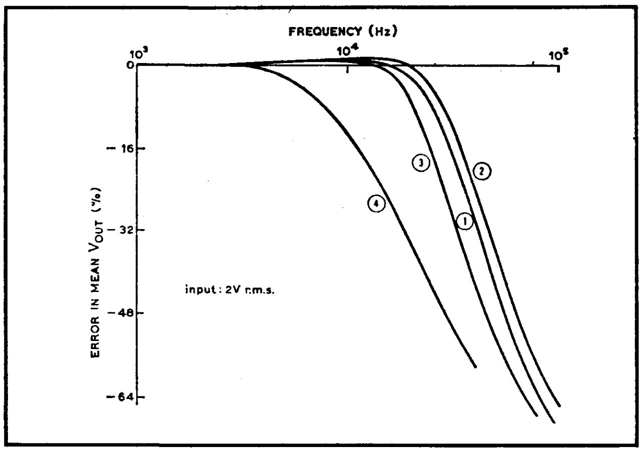
Wireless World Circard

Series 4: A.C. measurement 12

A.C. adaptor for digital voltmeter



Typical performance
Supplies: ±15V, ±2mA
A_1: 741
Diodes: CA3019 (part)
R_1: 20kΩ, ±5%
R_2, R_3, R_4, R_L: 10kΩ ±5%
Source res: 60Ω
Error in mean V_{out}: ≈ +3.7%



Circuit description
This circuit is basically an inverting amplifier having its gain defined by R_2/R_1. All of the amplifier's output current flows in the bridge rectifier and then divides between R_2 and R_4. For negative input voltages the output of the operational amplifier goes positive, producing a current in R_L from node B to A via diodes D_4 and D_1. For positive values of V_{in} the amplifier's output swings negative bringing diodes D_2 and D_3 into conduction, with D_1 and D_4 reverse biased, again producing a unidirectional current in R_L from B to A. The p.d. across R_L is thus a measure of the mean value of V_{in}, its value depending on the choice of the resistors. As the r.m.s. value is 1.11 times the mean value for a sineware input it is possible to scale a moving-coil instrument connected in place of R_L to read r.m.s. values directly. Full-scale current will be determined by R_2 in parallel with R_4. By making R_2 about ten times R_4 the full-scale current can be set by R_4, which allows the r.m.s. scaling factor of the movement to be determined largely by R_2 for a given value of R_1. A digital voltmeter having a differential input may be connected between nodes A and B to measure V_{in} (r.m.s.) directly provided that the ripple component of the p.d. across R_L is sufficiently smoothed This can be achieved by the use of a sufficiently large capacitor across R_L or by replacing R_L with a circuit of the form shown over (left). In either case, the value of R_4 should be chosen to prevent overloading of the operational amplifier during the initial charging of the capacitor. To provide a reasonably small degree of loading on the source, R_1 and hence R_2 and R_3 must be made much larger than the source resistance. The amplitude response of the circuit can be improved by making A_1 an operational amplifier that allows the use of feed-forward compensation, as shown in card 3.

Component changes
Replacing the CA3019 diode bridge with 4 × PS101 silicon diodes typically produces the response shown in curve 2: low-frequency error in V_{out} (mean) ≈ +3.7%.
Using the PS101 diodes with the 741 operational amplifier replaced by a 301 with a 33-pF compensation capacitor typically produces the response shown in curve 3: low-frequency error in V_{out} (mean) ≈ + 3.7%.
Useful range supplies: ±3 to ±18V.
Useful range of V_{in}: 350mV to 4.2V r.m.s. R_4 min for V_{in} min and no significant peak clipping ≈ 15Ω.

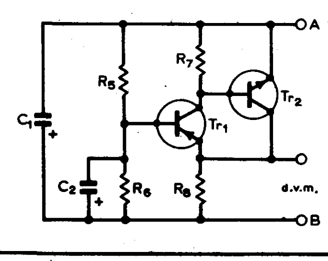
Circuit modifications
- To measure alternating voltages on a differential-input digital voltmeter the resistor R_L should be replaced with a network that is capable of passing the d.c. and which has a long enough time constant to sufficiently smooth the a.c. ripple. A circuit of the form shown left may be used for this purpose with Tr_1: BC126, Tr_2: BC125, R_5, R_6: 1MΩ, R_7, R_8: 10kΩ C_1: 4.7μF, C_2: 22μF, A_1: 741, D_1 to D_4: PS101, the response was typically as shown in curve 4: low-frequency error in V_{out} mean: +5·4%. Lowest useful frequency was approximately 10Hz. Tr_1 may be replaced by an f.e.t. or some other high input impedance circuit such as a follower to allow the use of larger-value bias resistors and a smaller C_2 value. V_{out} can be made equal to the r.m.s. value of V_{in} by scaling $R_2 = (2\sqrt{2}R_1)/\pi$.
- The circuit shown centre may be driven from a grounded source and exhibits a very high input impedance but is subject to a common-mode error. The circuit shown right also has a high input impedance, does not have a common-mode problem but must be supplied from a floating source.

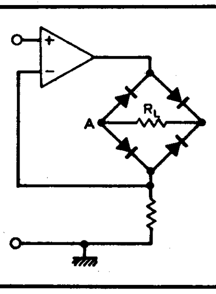
Further reading

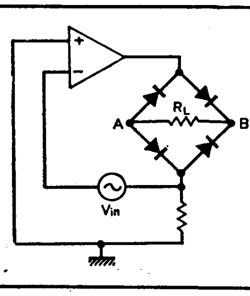
Gellie, R. W. & Klein, A. G., Accurate a.c.-d.c. converter for low frequencies, *Electronic Engineering*, 1967, p.484.
Dromgoole, M. V., Op.amp. a.c. millivoltmeter, *Wireless World*, 1970, p.75.

Cross references
Series 4, cards 3 & 7.

Circuit modifications





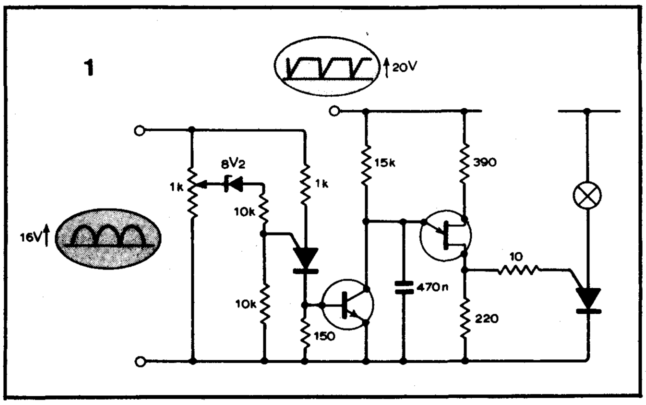


Wireless World Circard

A.C. measurements

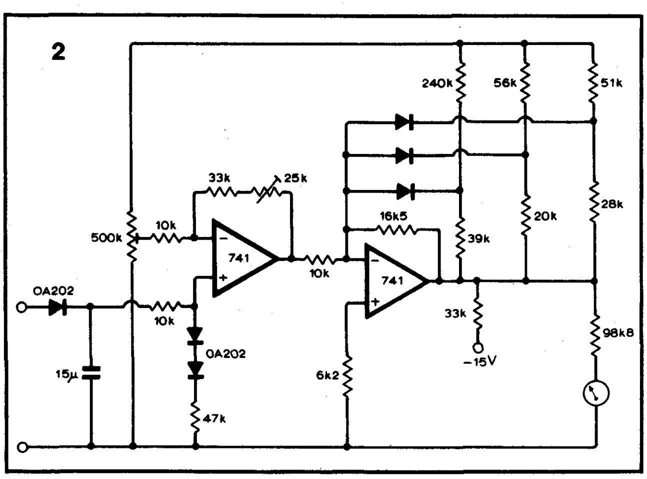
1. Strictly speaking, this circuit detects a particular peak level of an input waveform rather than measures a range of values. A full-wave bridge-rectified sine wave, derived via a transformer from the main supply being monitored, is applied via a 1-k Ω potentiometer and a zener diode to a thyristor. If the peak value of this voltage fails below a critical value the thyristor fails to fire, the transistor remains non-conducting and the unijunction transistor is allowed to fire the second thyristor, illuminating the lamp (an interrupt button would be needed to restore the thyristor to the off-state). The supply for the unijunction is derived from a separate bridge rectifier and a higher voltage winding (100V r.m.s.), zener stabilized so that supply variations cannot inhibit operation of the unijunction transistor.

Spetsakis, A. and Milas, A. Line-voltage fluctuation indicator detects single-cycle variations, *Electronics Design*, vol. 20, 1972, pp.80–2. (Oct. 26).

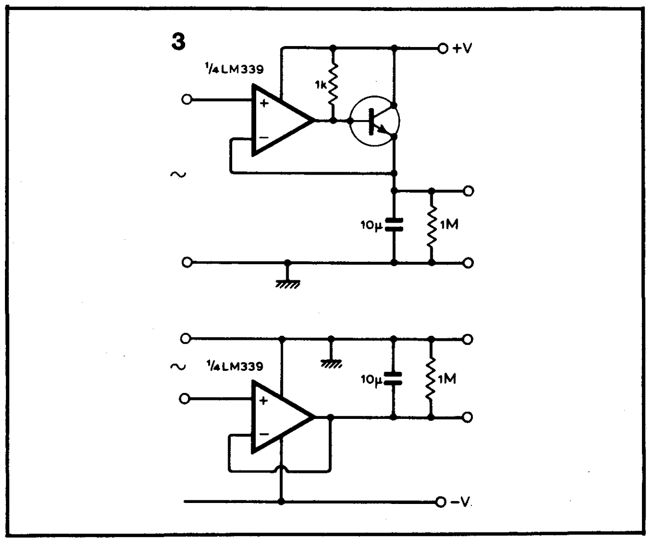


2. For measurement on telephone circuits and other equipment conforming to similar specifications, a meter scaled linearly in decibels (relative to a standard level) is an advantage. The circuit shown illustrates the wave-shaping technique that can be used. The signal is first peak-rectified, and then amplified with the addition of temperature compensating diodes. A second amplifier has non-linear negative feedback that shapes the transfer function. The resistors are non-standard and require to be selected or synthesized by series/parallel combinations. The frequency response of the original circuit was only required to reach 3.5kHz and the 741 amplifiers are more than adequate. Additional stages are shown in the reference article to extend the range of the meter.

Smith, J. H. Linear decibel meter, *Wireless World*, vol. 78, 1972, pp.488–90.



3. Quad-comparators such as the LM339 have considerable advantages in the design of simple peak-detectors. Because the input common-mode range includes ground when operated from a single positive supply, then if 100% negative feedback is applied the output is restricted to positive values. However the active device at the comparator output is an open collector transistor. The external transistor provides positive current to charge the capacitor until its p.d. matches that at the comparator non-inverting input. This continues until the input positive peak after which the capacitor discharges slowly through the 1-M Ω resistor. The second circuit senses negative peaks using a negative supply but the input common-mode range does not then include ground, so the circuit is not suitable for small signals.



Smathers, R. T., Frederiksen, T. M. and Howard, W. M. LM139/229/339, A quad of independently-functioning comparators, National Semiconductor application note AN–74, 1973, p.16.