

circuit designs

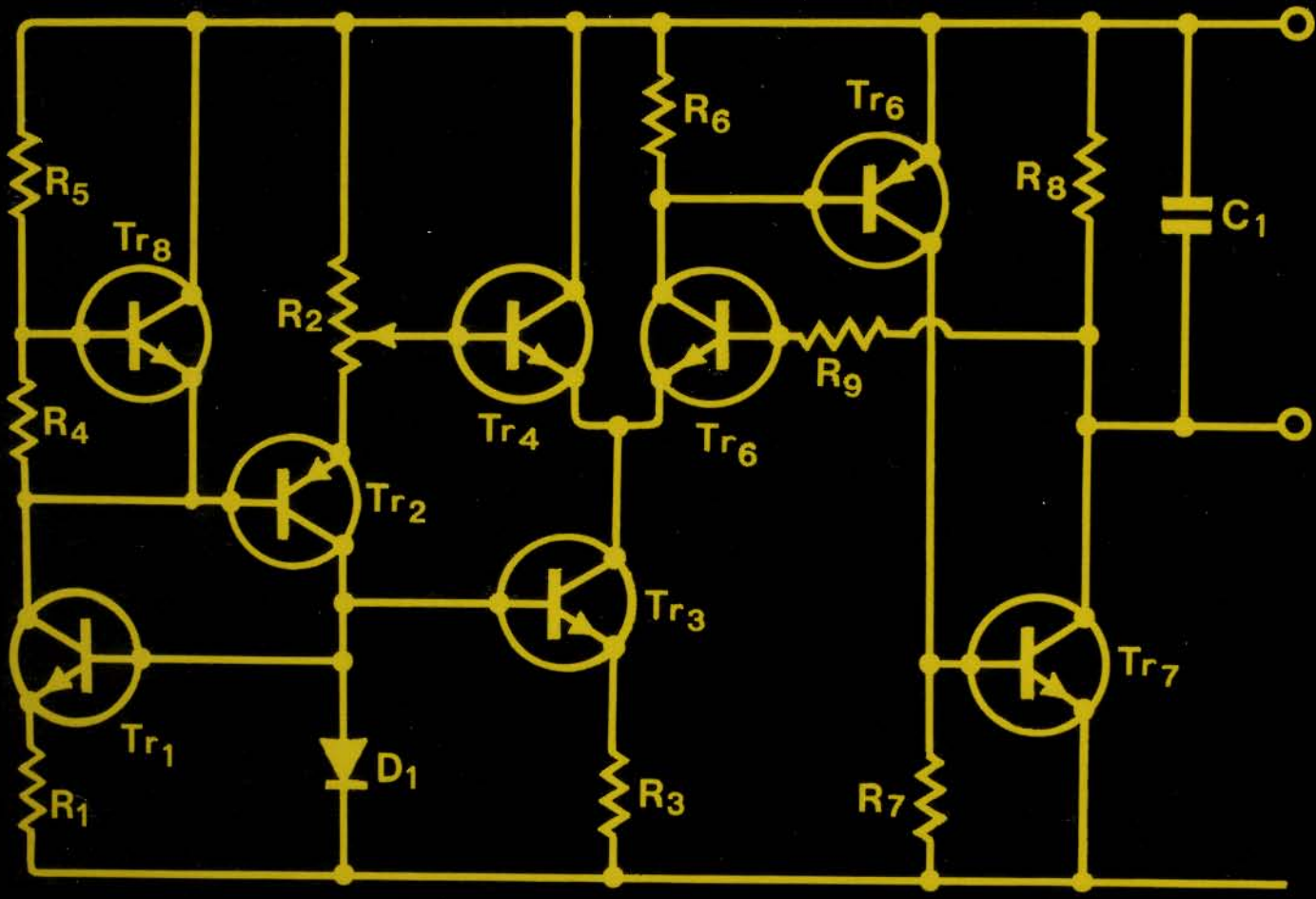
circuit designs

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Collected Circards

PWilliams/J Carruthers/JHEvans/JKinsler

PWilliams/J Carruthers/JHEvans/JKinsler



This book is for those who would design, use or understand electronic circuits. In October 1972 *Wireless World* launched a circuit information system called CIRCARDS.

The system was based on published sets of 8" x 5" cards to complement personal filing systems. Published in conjunction with introductory articles in *Wireless World*, CIRCARDS gave the subscriber *selected* and *tested* circuits, descriptions of circuit operation, component values and ranges, circuit limitations, modifications, performance data and graphs.

This book contains information previously included on the first ten sets of cards. This information has been corrected where necessary and supplemented with 29 additional circuits.

Introductory articles published in *Wireless World* complement each section.

The CIRCARD system won an IPC Business Press journal award for the best innovation.

circuit designs

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Collected Circards

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Preface

This book is for those who would design, use or understand electronic circuits. It is based on Circards—an award-winning* circuit information service that commenced in the October 1972 issue of *Wireless World*. We first thought of this scheme as an offshoot of our teaching, when we saw the problems faced by our students during their industrial training and in the early stages of their career. No matter how good the student (nor how brilliant the teachers!) it is very difficult for him to compete with engineers experienced in that field. In particular they will have acquired an instinct for the faults and foibles of various circuits, and will have at least a mental library of starting points if not complete solutions.

The same difficulties face any designer or user of electronic circuits and systems when tackling an unfamiliar topic. Specialist skills in the design of, say, voltage regulators will not promote a flying start in the race to design the perfect r.f. amplifier against skilled competitors. Technicians are frequently called on to produce a prototype at short notice and with a specification that gives little help in deciding on the best implementation.

For all these people, we considered that there was a common information requirement that could not be easily met by the conventional sources. Our proposals were that available information on circuit design should be sifted by a small team with a varied industrial background; that one circuit function at a time be considered and that a representative sample of circuits be tested and reported on. The format of the original was that of cards (hence Circards) each set containing ten or a dozen cards with one basic circuit per card. In each case the aim was to show not only how that circuit operated, but also to discuss means of extending its operating range by varying component values and using alternative devices. Variants on the circuit together with further reading and cross-references to related cards were also included. We felt that in collating the published information and cross-checking by further measurements, we were doing what the users of Circards could have done for themselves – if they could have found the time.

The chosen format was aimed directly at bench-use, where a single card could be removed after being selected as the most likely solution. On other cases the need is for a source that has a discussion of the general principles of a class of circuits in addition to the detailed study of examples. While such discussions are readily available in journals and textbooks we hope that by including the original articles from *Wireless World*, the reasoning behind the choice of particular

circuits will be clarified. New circuits on each of these topics have appeared in abundance since the cards were published and a selection from these is included together with brief descriptions. These are inevitably a personal choice, which we believe will supplement and complement the original data without supplanting it.

We should like to record our heartfelt thanks to our wives and families for their patience; to the governors, principal and staff of Paisley College for their encouragement and support; and to the editorial staff and management of *Wireless World* for the enthusiasm, initial and sustained, with which they took up our proposals. *Peter Williams*

We believe this book to be unique among books of collected circuits in presenting thoroughly tested circuits together with typical measured data on circuit behaviour. More than this, the effects of changes in component values, supply voltage, and where appropriate load resistance and temperature, will have been investigated (though not necessarily all at the same time), together with the use of alternative transistors or integrated circuits. To extend the circuits further, there is normally a section on modifications, usually to extend performance in some way but sometimes to simplify the circuit.

The Circard project started off in an attempt to break some of Finagle's laws:

- the information you have is not that which you want
- the information you want is not that which you need
- the information you need is not that you can obtain.

We hope the authors have been successful in helping you to break these—if sales are any judge they certainly have. To a large extent the success of such a project depends on feedback from readers. While there is no shortage of ideas for further topics in the on-going Circard project, we do make a point of passing on any ideas from subscribers to the authors—novel ideas for circuits can often be incorporated into a set. And a significant demand for a particular topic can lead to an alteration of priorities in regard to the sequence of topics presented.

We trust you will join us in spirit in thanking the authors for undertaking this ambitious project. *Geoffrey Shorter*, technical editor, *Wireless World*.

* *Circards* gained the IPC Business Press Award for the Best Journal Innovation of 1973.

Wireless World Circard

Series 1: Basic active filters

This set of cards—the best selling one of all—deals mainly with second-order active filters i.e. those with active devices for gain having a second-order transfer function. The most popular nowadays are probably the Sallen & Key types using unity-gain amplifiers (cards 4 and 5). Though widely used, they suffer from the disadvantage that two resistors need to be made variable and ganged for variable frequency applications. The positive feedback Wien-bridge notch filter too needs to have two components made variable to alter its centre frequency. Three cards deal with filters that are tunable by just one potentiometer (cards 3, 8, 10), and card 9 includes a bibliographical reference to a not so well-known twin-T circuit having single-component variation of the notch frequency.

One tunable filter technique that uses digital integrated circuits is given in the *n*-path filter card. Here resonant frequency is varied by changing commutating frequency, leaving bandwidth constant.

Many of the circuits given use integrated-circuits, mostly the 741 op-amp. (Card 6 is an exception, using the TAA960 i.c. with four inverters.) But for many low-frequency, low-Q applications, the op-amps can be replaced by a single transistor or Darlington pair. In the Sallen & Key types, the discrete device would need to be used as an emitter follower.

- Wien-bridge bandpass filter 1
- Wien-bridge all-pass network 2
- Voltage-controlled filters 3
- Low-pass Sallen & Key filter 4
- High-pass Sallen & Key filter 5
- Low, high and band-pass triple amplifier 6
- Op-amp triple with phase compensation 7
- Multi-feedback filter 8
- Adjustable-Q twin-T notch filter 9
- Easily-tuned notch filter 10
- Compound filters 11
- N-path filter 12

Basic Active Filters

Like the dotting parent reproached for a birthday present of an encyclopaedia with the response "It tells me more about things than I want to know", any writer on this topic has to exercise care.

The filters to be described belong to that broad class of circuits using active devices together with passive components, some of which are reactive, to produce a frequency-dependent transfer function. The possibilities range from single CR time-constants in the feedback path of an amplifier to complex multi-gyator circuits. The former have applications in audio frequency work, while the latter are still the subject of continuing research. It is the in-between varieties that we shall consider, in particular second-order filters of various kinds.

These have numerous applications used separately but may in addition be combined to produce more complex filter functions. By second-order is meant that the transfer function is represented by an equation containing ω^2 as well as ω . This is because of the presence of at least two reactive components — capacitors or inductors. The latter are inconvenient and expensive below the radio-frequency range, and active-filters dispense with them by synthesizing the desired function using capacitors.

The common filter functions are

Low-pass: signals below a given cut-off frequency are passed at, or amplified to, a fixed level, falling by 3 dB at the cut-off frequency with a slope that reaches 12 dB per octave for a second-order filter — Fig. 1.

High-pass: the pass-band lies above the cut-off frequency, with 3 dB attenuation at the cut-off frequency, and with attenuation increasing at 12 dB/octave below this frequency — Fig. 2.

Band-pass: signals at both low and high frequencies are progressively attenuated with a maximum response at the centre frequency; the sharpness of the response is measured by the Q of the circuit with $Q = f_0/2\Delta f$, where f_0 is the centre frequency and $2\Delta f$ is the bandwidth for an attenuation of 3 dB relative to that at the centre frequency — Fig. 3.

Band-stop or notch: low and high frequencies are passed with a maximum attenuation (ideally zero transmission) at the centre or notch frequency — Fig. 4.

All-pass: the magnitude of the transfer-function is constant but with a frequency-dependent phase-shift — Fig. 5.

The low- and high-pass filters are sometimes modified so that the output falls to a low but constant value at some other frequency — Figs. 6 and 7. The band-pass and

band-stop filters may require extended frequency ranges within which the pass and stop functions are maintained. One possibility is the use of complex high-order filters. Alternatively a combination of one or more second-hand filters tuned to different frequencies can be used. A perfect null response is possible only at precise specified frequencies — Figs. 8 and 9.

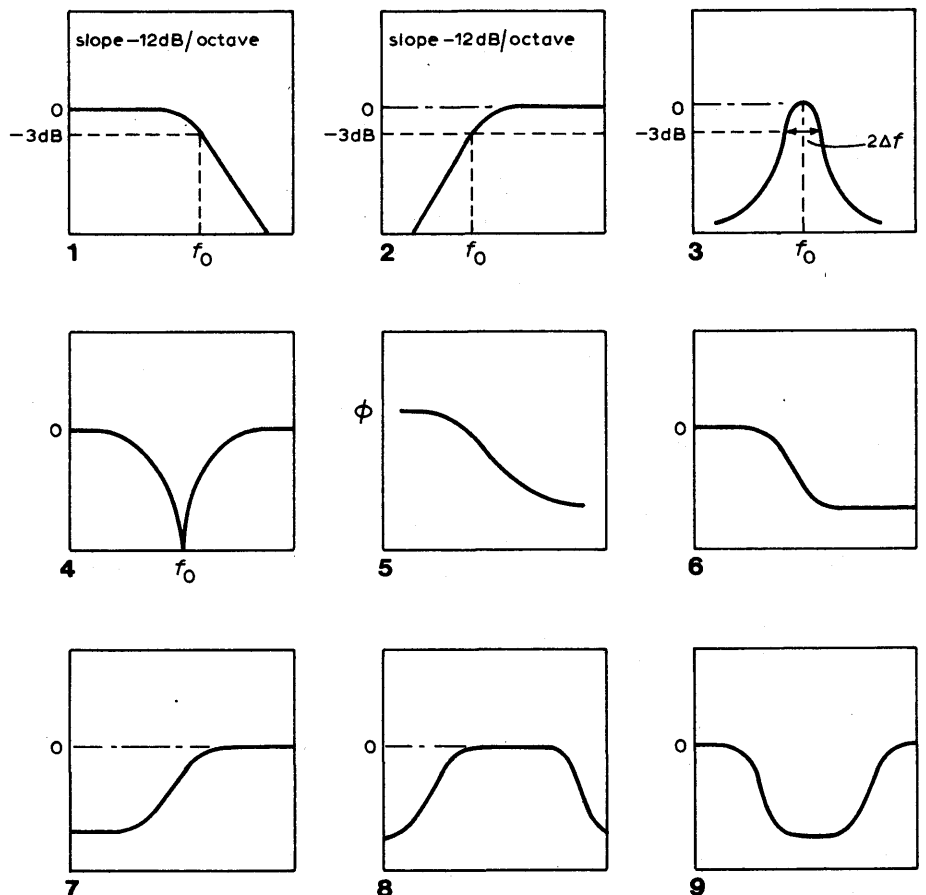
In determining which active circuit configuration most suits a particular filter function, the required Q is a major factor. High Q is often a requirement for band-pass and notch filters. It can be attained by two fundamentally different methods.

The first, requiring a single amplifier in the simplest cases, arranges for Q to depend on the difference between two nearly equal terms in the transfer function. The circuits using this technique are sometimes recognizable as a bridge circuit with the bridge almost at balance, combined negative and positive feedback with the latter almost equalling the former, or a negative-impedance converter in which an imped-

ance in one part of a circuit appears elsewhere with a negative sign.

Any given circuit may be recognizable under more than one of the above headings which might be considered as different manifestations of the same underlying process. While high Q is possible, it becomes over sensitive to component and environmental changes, as it depends on the near-cancellation of comparable terms. The fractional change in Q caused by a change in some component is broadly proportional to the nominal Q aimed for, e.g. a circuit with $Q = 20$ as the design value might have a practical value of $Q = 25$ for a 1 or 2% change in a particular resistor.

The corresponding practical values for nominal Q values of 10 and so might be 11 and 75 respectively. The centre frequency is also dependent on component values but active circuits are comparable to passive networks in this respect. The exception is at high frequencies when the amplifier limitations may be all too apparent. This is shown by the increasing departures at the observed



In these graphs illustrating low-pass, high-pass, band-pass, band-stop and all-pass active filters, transfer function magnitude is plotted as the ordinate with frequency as the abscissa, except in 5 where the ordinate is phase angle.

centre-frequency from the nominal value, coinciding often with a fall in the achievable value of Q .

The second high- Q approach uses multiple active devices. One type, well-known in analogue computing, uses two integrators in a closed loop such that the open-loop gain of the system at the centre frequency, and hence Q , is controlled by a single resistor. The same pattern of passive components reappears in filters based on the gyrator, a device which has an impedance at one port related to that presented to a second port in such a way that a capacitance is "gyrated" into an inductance, a parallel tuned circuit into a series tuned circuit, etc. Gyration may be constructed from various configurations of amplifiers, with the usual limitations on frequency response, for example.

In neither approach is the Q excessively sensitive to component variations. Broadly the sensitivity is comparable to that obtaining with passive circuits, i.e. a 1% change

in any component will not normally change the Q by more than 1% even at high values of Q . With multiple amplifier filters, there is more than one output available, and some circuits have band-pass, low-pass and high-pass functions available simultaneously. The falling cost of operational amplifiers makes such solutions attractive, particularly for larger systems where the flexibility is a considerable advantage. A problem can be the phase-shift in multiple amplifiers, which can result in oscillation when high Q is attempted at frequencies near the upper frequency limit of the amplifier.

For simple low-pass and high-pass applications the single-amplifier methods are suitable, and as an example scratch and rumble filters may use twin-gang potentiometers for widely variable cut-off frequencies. Using special i.c. amplifiers with very high input resistance the cut-off frequency can be readily extended below 1 Hz if required, while the upper limit approaches 1 MHz.

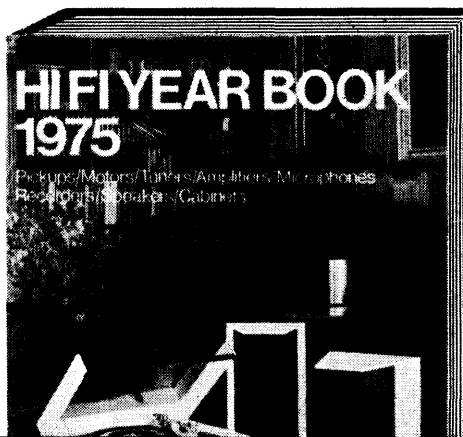
A completely different approach which is certain to become dominant in many fields is that of the n -path filter. There will presumably be as many variants as there are at present with the analogue approach, but the basic technique appearing in some recently published circuits is as follows.

In place of a single filter a number, say n , of identical units are switched into the circuit in succession. These may be, for example, low-pass filters, passive or active, or simply the capacitors of a set of CR circuits using a common R . Switching frequency is set to n times the fundamental frequency to which the filter is to be tuned.

The filter has responses at harmonics of the fundamental frequency and may have a very high Q at each, but simple low-cost filters of the type described earlier can be used to attenuate these responses.

A vital point is that the filter centre-frequency is controlled by the clock-rate driving the switches, and therefore readily variable over a wide range, as is bandwidth.

THE WHO, WHAT AND WHERE OF HI-FI

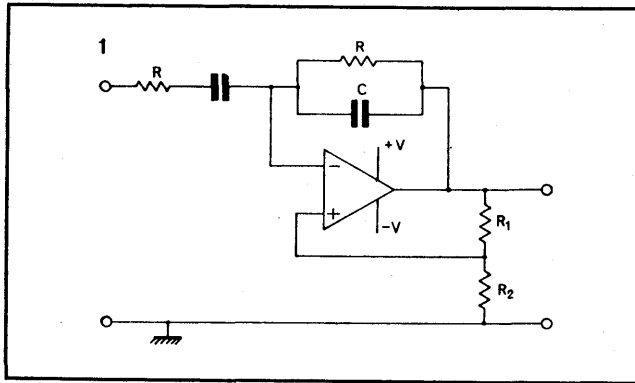


In the search for quality it helps to know where to look. Hi-Fi Year Book tells you everything you need to know, with separate illustrated sections for every major category, giving you prices and specifications of over 2,000 products. And it's got directories of dealers and manufacturers — plus a host of articles on the latest hi-fi developments and their application. So if you want information like you want hi-fi, order your copy today because it sells out pretty quickly! £2.00 inclusive from the publishers, or from newsagents and bookshops.

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Wien-bridge bandpass filter



Typical performance at 1kHz

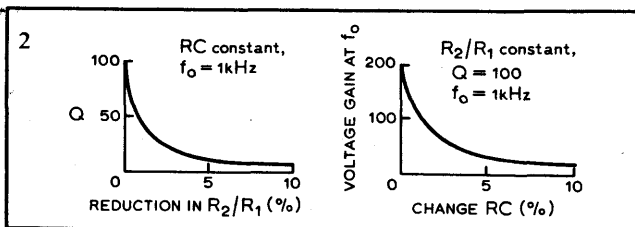
$$f_0 = \frac{1}{2\pi CR} \text{ Hz}$$

$$Q = \frac{1}{2 - (R_2/R_1)}$$

$$\text{Voltage gain, } \frac{V_{out}}{V_{in}} \approx -3Q$$

To find Q measure voltage gain and divide by -3

IC : 741; supplies $\pm 15V$
 $C = 1nF \pm 1\%$
 $R = 150k\Omega \pm 5\%$
 $(R_2 + R_1) = 10k\Omega$
 Source resistance: 60Ω
 $V_{out}(\text{max.}) = 9.2V \text{ r.m.s.}$
 Max. load current: $12mA$
 Q is constant for supplies between $\pm 3V$ and $\pm 18V$ if clipping is avoided.
 $Z_{in} \propto 1/Q$; $Z_{out} \propto Q$.



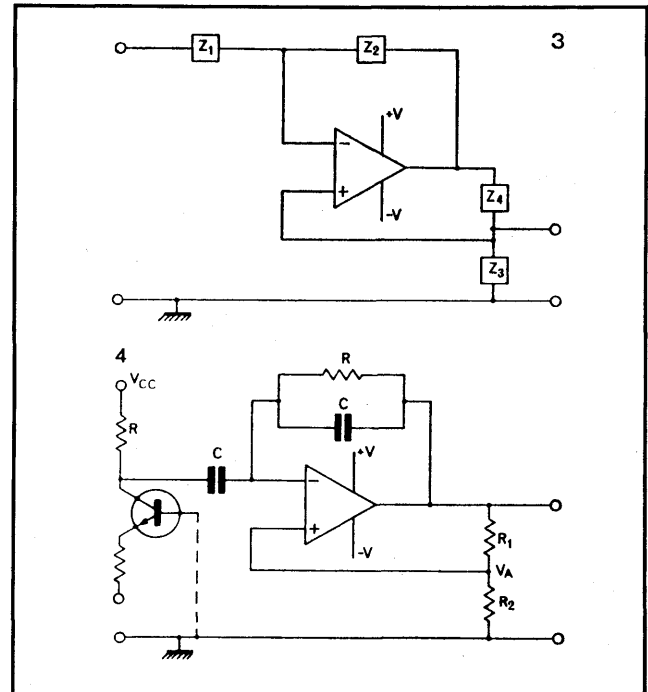
Circuit description

The Wien network—Fig. 1—gives a maximum response at $f_0 = 1/2\pi RC$ Hz. Positive feedback through R_1 , R_2 sharpens the response, e.g. with R_1 and R_2 replaced by a potentiometer, continuous control of circuit Q is obtained without change in centre frequency. Changing both R s or both C s, maintaining equality, varies f_0 without changing Q . Sensitivity to component changes is proportional to Q for large Q . At high frequencies, amplifier phase shift causes Q and f_0 to depart from nominal values. Input impedance falls as Q increases. Satisfactory for moderate Q values (5–25) at frequencies in the audio range. High Q is obtainable if only short-term stability is required.

Component changes

● Using $\pm 1\%$ capacitors and $\pm 5\%$ resistors, f_0 will typically be within $\pm 5\%$ of theoretical value up to approximately $15kHz$ and within $\pm 10\%$ up to about $22kHz$.

- Large R -values for low f_0 produces an output d.c. level up to about $1V$ which can be reduced with offset null adjustment.
- Source resistance should be $\ll R$ for predictable f_0 and $< R/15Q$ for predictable Q within -5% .
- Circuit will oscillate when $R_2/R_1 \geq 2$.
- 741 op-amp may be replaced by a 748 or 301 using a $30-pF$ compensation capacitor.
- Reducing compensation capacitor of 748 or 301 to $3.3pF$ typically makes f_0 predictable to within $\pm 5\%$ of theoretical value up to about $22kHz$.



Circuit modifications

- Fig. 3 shows the general form of the circuit, one version of which has been discussed. Other configurations are possible by interchanging Z_1 and Z_4 and/or Z_2 and Z_3 . Note that the output may not then be taken from the op-amp output.
- Since $Z_{in} \propto 1/Q$ for high Q a buffer amplifier may be added at the input. Buffer at the output is only required for alternative versions (except at very high Q).
- R at the input of the circuit may be taken to ground and from a current source e.g. from collector of a common-base stage—Fig. 4—or a cascode stage.

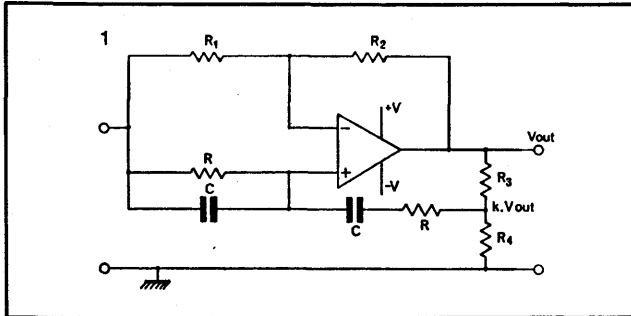
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Cross references

Series 1, cards 3, 6, 7, 8, 11 & 12.

Wien-bridge all-pass network



Typical performance at 1kHz

180° phase shift occurs at f_0 , where

$$f_0 = \frac{1}{2\pi RC} \text{ Hz}$$

$$\lambda = R_1/R_2$$

For unity voltage gain

$$k = \frac{5\lambda - 1}{\lambda + 1}$$

IC : 741; supplies $\pm 15V$

$C = 1nF \pm 1\%$

$R = 150k\Omega \pm 5\%$

$(R_3 + R_4) = 1k\Omega$

$(R_1 + R_2) = 10k\Omega$

Source resistance: 60 Ω

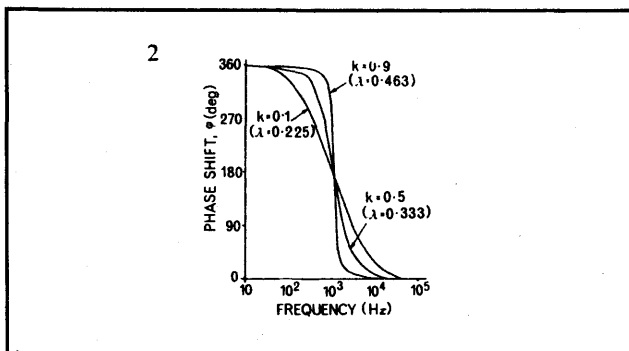
With unity voltage gain

$V_{out(max.)} = 26V$ pk-pk.

No significant change in

performance for supplies in the range ± 3 to

$\pm 18V$.



Circuit description

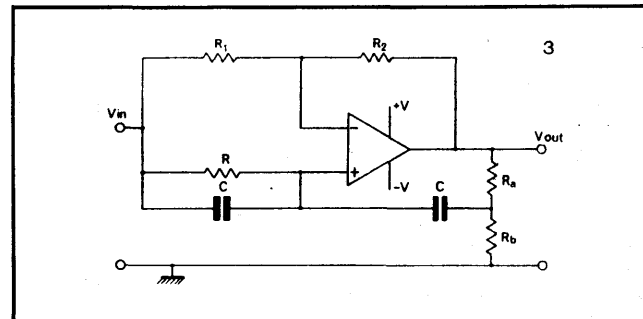
When λ has the correct value for a given value of k the network has unity voltage gain but the phase difference (ϕ) between output and input is frequency dependent. At $f_0 = 1/2\pi RC$ Hz, $\phi = 180^\circ$. Increasing k and adjusting λ to the appropriate value increases the magnitude of $d\phi/df$ in the region of f_0 . Suitable for use at audio frequencies.

Component changes

● 741 i.c. may be replaced with a 748 or a 301 using a 30-pF compensation capacitor without change in performance e.g. $d\phi/df$ characteristic predictable up to $f_{0(max)}$ of about 10kHz, with unity voltage gain.

● Reducing compensation capacitor to 3.3pF extends range of predictable performance up to $f_{0(max)}$ of about 20kHz.

● With $R = 150k\Omega$ and $(R_1 + R_2) = 10k\Omega$, $(R_3 + R_4)$ may be increased to approximately 22k Ω without significant change in performance.



Circuit modifications

● As in the bandpass filter (card 1) the arms of the bridge may be interchanged to produce alternative configurations having differing $d\phi/df$ effects.

● For fixed $d\phi/df$ and f_0 , the arrangement in Fig. 3 may be used where R_a and R_b in parallel equals R and their ratio determines the value of K .

● For variable $d\phi/df$ characteristic, without having to incorporate output potential divider resistance in CR calculations, insert a voltage follower between the potential divider and the series network.

● A buffer amplifier may be used at the input, though the input impedance is not likely to be as low as in the high- Q bandpass filter-card 1.

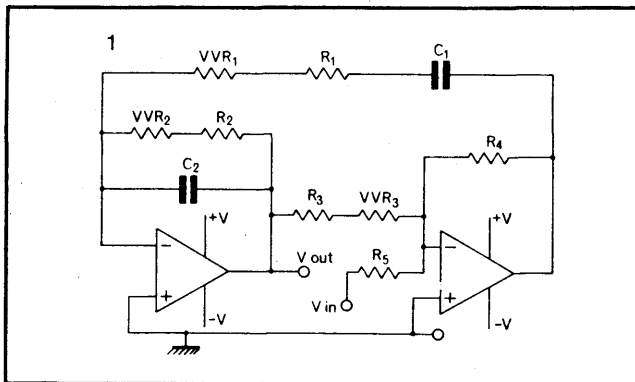
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2. As 1: Alternative allpass networks using Wien's bridge, p.188.
3. Mitra, S. K., Analysis and synthesis of linear active networks, Wiley, 1969 pp.479-81.

Cross references

Series 1, card 1.

Voltage-controlled filters



Typical performance at 1kHz

$$f_0 = \frac{1}{2\pi RC} \text{ Hz}$$

where $R = (R_1 + VVR_1) = (R_2 + VVR_2)$

ICs 741; supplies $\pm 6V$

VVRs: $1/4 \times CD4016AE$

$C_1 = C_2 = C = 0.1 \mu\text{F} \pm 2\%$

$R_1 = R_2 = 1\text{k}\Omega \pm 5\%$

$R_3 = 1.2\text{k}\Omega \pm 5\%$

$R_4 = 3.3\text{k}\Omega \pm 5\%$

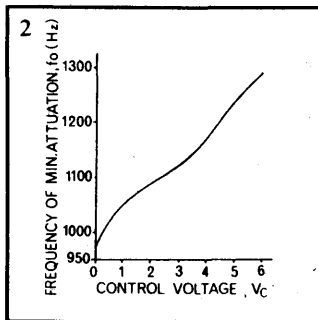
$R_5 = 10\text{k}\Omega \pm 5\%$

$V_{DD} = +3V, V_{SS} = -3V$

$V_C = 0$ to $+6V$

VVRs variable within the range approximately 680 to 280 Ω .

$V_{C3} = 4.32V$ to produce $Q \approx 10$ when $V_{C1} = V_{C2} = 0V$.



Circuit description

This bandpass circuit is related to that of card 1. Separation of the frequency-sensitive and resistive networks allows each section to have one terminal as a virtual earth. This simplifies the injection of signals (defined input resistance e.g. through R_5) and allows devices such as m.o.s. transistors to be used as frequency and/or Q control elements, with bias voltages referred to ground. The control elements used were complementary m.o.s. transmission gates with resistances controlled by the applied bias voltages. VVR_1 and VVR_2 control f_0 and VVR_3 controls Q by varying the ratio $R_4/(R_3 + VVR_3)$.

Component changes

● Supplies may be varied within the range ± 3 to $\pm 18V$ and are conveniently chosen to be compatible with supplies required for the VVR elements. (N.B. CD4016AE max $V_{DD} - V_{SS} = 15V$)

● The greater the ratio VVR_n/R_n , $n = 1$ or 2 , the wider the range of f_0 variation for given values of C within the audio band.

● For a given value of R_4 , Q variation is greatest when $R_3 \rightarrow 0$.

Circuit modifications

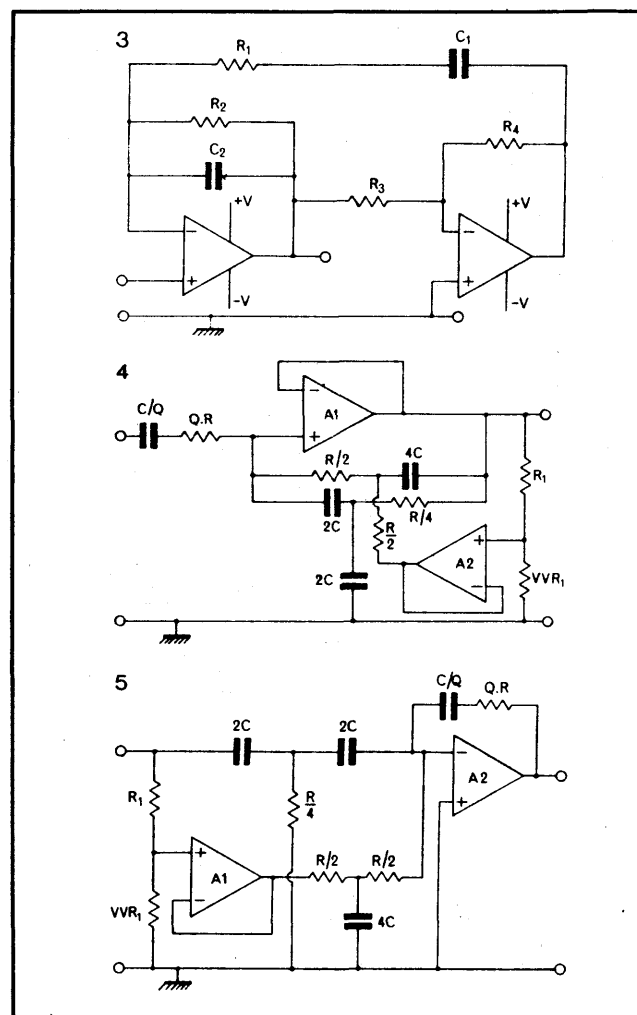
● By scaling $R_1 = R_2/2$, $C_1 = 2C_2$ and setting $R_3 \rightarrow R_4$ the bandpass filter will provide outputs from the op-amps that are in antiphase and of almost equal amplitude.

● Any device with a fairly linear resistance, variable by an external parameter may replace the c.m.o.s. gates in this and other filters, (easiest where R_s have a common point at or near ground potential, permitting this point to be connected to sources of f.e.t.s, for example). Possibilities include CdS

photocells, temperature-sensitive resistances like copper, platinum, and thermistors where signal swings do not change resistance, i.e. relatively high-power types.

● As in all virtual-earth amplifiers, the signal may be injected in voltage form into an otherwise-grounded, non-inverting terminal, as in Fig. 3, providing a high input resistance may alter filter characteristic.

● Figs. 4 and 5 show the voltage control principle applied to a parallel-T bandpass and notch filter respectively. In each circuit f_0 is controlled by the VVR.



$$\text{In Fig. 4: } f_0 = \frac{1}{2\pi RC} \cdot \sqrt{\frac{R_1}{R_1 + VVR_1}} \text{ Hz.}$$

$$\text{In Fig. 5: } f_0 = \frac{1}{2\pi RC} \cdot \sqrt{\frac{VVR_1}{R_1 + VVR_1}} \text{ Hz.}$$

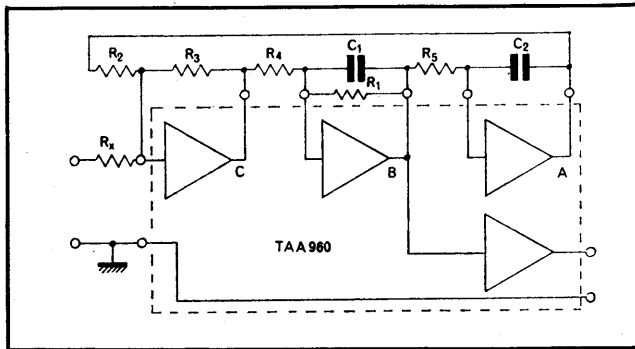
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Cross references

Series 1, cards 1, 6-10.

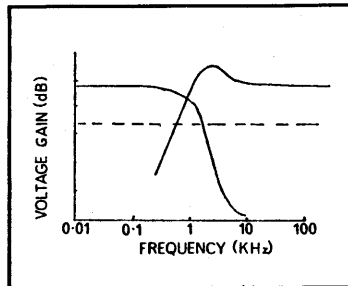
Low, high & band-pass triple amplifier



Typical performance

$$f_0 = \frac{1}{2\pi \sqrt{C_1 C_2 R_4 R_5}}$$

Drive: 400mV pk-pk.
 Supply: +6V.
 $R_2, R_3, R_4, R_5 = 10k\Omega \pm 5\%$
 $C_1, C_2 = 10nF \pm 1\%$
 $R_X = 470k\Omega \pm 5\%$
 $R_1 = \infty$
 $f_0 = 1,583Hz$ $Q = 49.5$



Circuit description

Three inverting amplifiers form the filter. A fourth amplifier buffers the output. There are two integrators, one damped by resistor R_1 , and a unity-gain inverting amplifier. A different filter function is obtained at the output of each amplifier; low-pass at A, bandpass at B and modified high-pass at C. Where all four amplifiers are included in a single i.c. as above, a restricted performance is offered by each amplifier individually, e.g. lower voltage gain, and the output resistance of a value that requires buffering if load resistances are not to change the filter characteristic. With R_1 absent, the Q of the band-pass filter is approximately 50. As R_1 is decreased, the response at A and C approach that usually required for second order, low-pass and high-pass responses respectively. Single-supply operation and direct coupling place a lower

limit on R_x of approximately $10k\Omega$, with $R_x = 50R_2$ for correct biasing. In addition, all outputs have a d.c. content.

Component changes

Supply voltage +6V: $f_0 = 1583Hz$, gain = 1.15, $Q = 49.5$.
 Supply voltage +5V: $f_0 = 1562Hz$, gain = 0.95, $Q = 40$.
 For best Q values, capacitors should be matched within 1%.
 For low-pass and modified high-pass filters $R_1 = 6.8k\Omega$,
 $Q = R_1/R_4 = 0.68$
 Drive signal 2V pk-pk.
 Output approx. 50mV pk-pk.

Circuit modifications

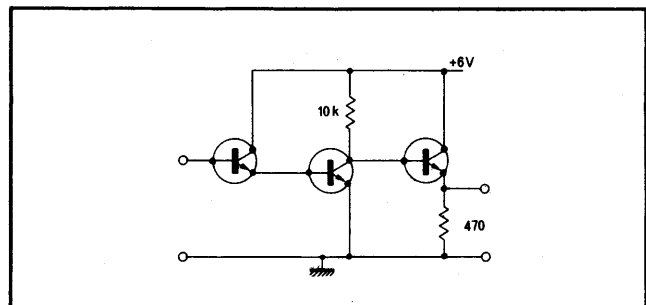
- As the three amplifiers are identical, the buffered amplifier may be used as the output for the low-pass or modified high-pass filter.
- For higher gain without disturbing d.c. conditions, input and output would have to be a.c.-coupled, then the input resistor could be reduced.
- The simple structure of the amplifiers can be duplicated by discrete circuits such as the amplifier with emitter-follower output shown.

Further reading

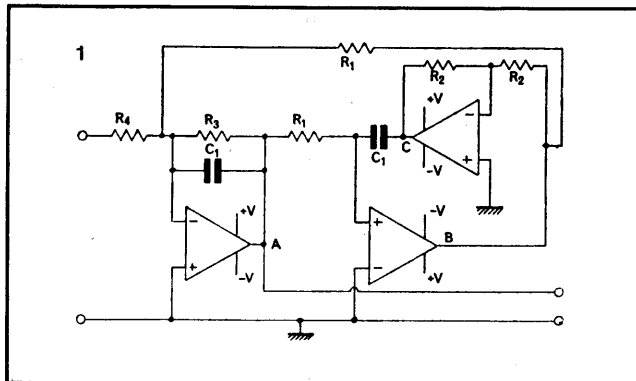
Mullard Ltd.: Triple amplifier for active filters, TAA 960.

Cross references

Series 1, cards 1, 3, 7, 8, 11 & 12.



Op-amp triple with phase compensation



Typical data

$$R_1 = 3.3\text{k}\Omega \pm 5\%$$

$$C_1 = 0.1\mu\text{F} \pm 10\%$$

$$R_3 = 330\text{k}\Omega \pm 5\%$$

$$f_0 = \frac{1}{2\pi R_1 C_1} = 482\text{Hz}$$

$$Q_0 = R_3/R_1 = 100$$

$$f_0(\text{obs}) = 470\text{Hz}$$

$$Q(\text{obs}) = 105$$

$$V_{\text{out}}/V_{\text{in}} \approx Q \text{ at } f = f_0$$

Circuit description

The same passive components as in card 6 may be used with separate op-amps having much higher gain and more suitable input and output characteristics. With the same amplifier configuration high Q is possible with excellent stability. At high frequencies the observed Q rises because of amplifier phase shifts. The alternative configuration shown provides some degree of cancellation of phase shifts allowing high- Q designs up to 100kHz with low-cost amplifiers. Precautions against over driving under these conditions are necessary as slew-rate limiting changes the amplifier response leading to the possibility of sustained oscillation.

Component changes

$$R_1 = 470\Omega \text{ to } 1\text{M}\Omega$$

$$R_3 = 470\Omega \text{ to } \infty$$

$$C_1 = 330\text{pF} \text{ to } 10\mu\text{F} \text{ (non-polarized)}$$

Amplifiers: N5741V, SN72741P, MC1741, etc.

Supply ± 5 to $\pm 18\text{V}$ ($\pm 3\text{V}$ in some cases).

Alternative amplifiers: any op-amp compensated for integrator operation.

$$f_0 = < 1\text{Hz} \text{ to } > 100\text{kHz}$$

$$Q < 1 \text{ to } > 100$$

Circuit modifications

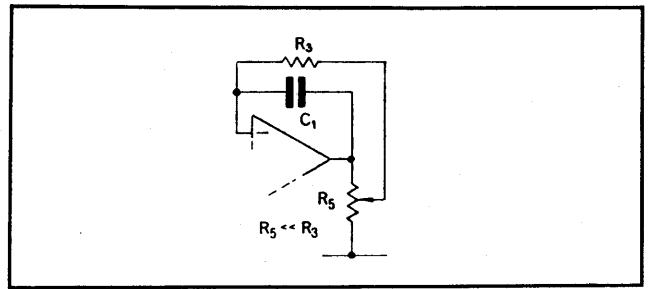
- Variable damping at high Q inconvenient because of large value for R_3 . Replace by fixed resistor fed from tapping across amplifier as shown e.g. $R_5 = 10\text{k}\Omega$ pot., $R_3 = 100\text{k}\Omega$.

- Summing outputs at three amplifiers with conventional virtual earth summer gives more general transfer function.

- Low-pass output available at C (normally used with $Q = 0.7$ for simple low-pass filters as in audio applications—higher Q gives peak in response just below cut-off frequency).

- Modified high-pass output at B—comments as above.

- Where configuration as in card 6 is used phase lead may be introduced by a small capacitor across input resistor of inverter—alternative method of neutralizing phase-shift.



Further reading

Kerwin, W. J., Huelsman, L. P., and Newcomb, R. W. State-variable synthesis for insensitive integrated circuit transfer functions, *IEEE J. Solid-State Circuits*, vol. SC-2, 1967 pp.87-92.

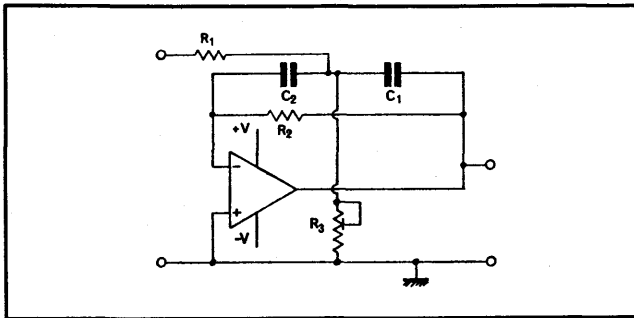
Vogel, P. W. Method for phase correction in active RC circuits using two integrators, *Electronic Letters* 1971, pp.273-5.

Girling, F. E. J. and Good, E. F., Active filters-7 The two-integrator loop, *Wireless World*, vol. 77, 1970 pp.76-80.

Cross references

Series 1, cards 1, 3, 6, 8, 11 & 12.

Multi-feedback filter



Typical performance

For $C_1 = C_2 = C$

$$f_0 = \frac{1}{2\pi C} \sqrt{\frac{R_1 + R_3}{R_1 R_2 R_3}}$$

$$Q = \frac{1}{2} \sqrt{\frac{R_2 (R_1 + R_3)}{R_1 R_3}}$$

$A_0 = R_2/2R_1$
 $C = 0.01 \mu\text{F} \pm 10\%$
 $R_1 = R_3 = 1.5\text{k}\Omega \pm 5\%$
 $R_2 = 470\text{k}\Omega \pm 5\%$
 Amplifier: 741
 $f_0(\text{obs}) = 845\text{Hz}$
 $Q(\text{obs}) = 11.3$

Circuit description

The inverting amplifier has multipath feedback which has a minimum at a single frequency. The overall response is then band-pass peaking at that frequency, which may be changed by varying C_1 , C_2 together, the circuit Q remaining constant. Changing R_3 also varies the centre frequency keeping the bandwidth constant.

Component changes

R_2 : 1.5k Ω to 1M Ω

R_1 , R_3 : 1k Ω to 100k Ω

C_1 , C_2 : 150pF to 10 μF

$Q = 1$ to 50 (At high Q values, sensitivity to component

variations is excessive)
 $f_0 < 1\text{Hz}$ to $> 100\text{kHz}$

Circuit modifications

● Varying R_3 changes the centre frequency, leaving bandwidth and centre-frequency gain unchanged.

● Feeding from a current source, R_1 can be omitted, readjusting R_3 to give required characteristic. Alternatively if preceding stage has specified output resistance it can be incorporated into R_1 .

● For low-gain low- Q applications the amplifier may be replaced by a single transistor in the common-emitter mode. R_2 provides base-current. Typical collector load resistances in range 1 to 10k Ω for passive network given. Alternatives include Darlington-pair amplifiers.

● Alternative passive networks for low and high-pass characteristics given in the reference.

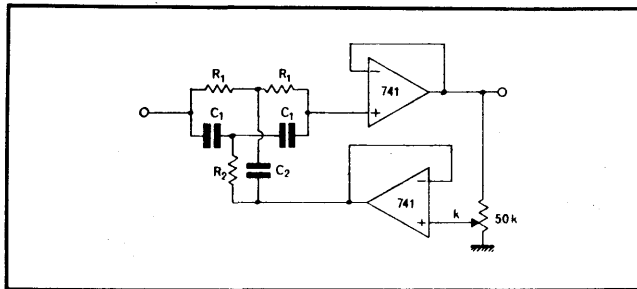
Further reading

Graeme, J. G., Tobey, G. E., Huelsman, L. P., Operational amplifiers: design and applications, McGraw-Hill, pp.287-95.

Cross references

Series 1, cards 1, 3, 6, 7, 8, 11 & 12.

Adjustable-Q twin-T notch filter



Typical performance

$$f_0 = \frac{\sqrt{n}}{2\pi R_1 C_1}$$

where $n = 2C_1/C_2 =$

$R_1/2R_2 (=1 \text{ usually})$

$f_0 = 1000\text{Hz}$ with

$C_1 = 10\text{nF}$,

$C_2 = 20\text{nF}$ (trimmed)

$R_1 = 16.2\text{k}\Omega$

$R_2 = 8.1\text{k}\Omega$ (trimmed)

$f_1 = 680\text{Hz}$ $f_2 = 1650\text{Hz}$

$f_0 = 50\text{Hz}$ with

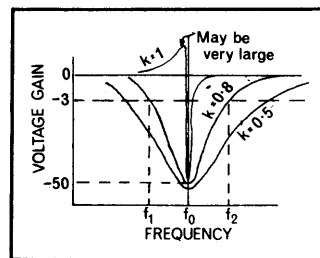
$C_1 = 10\text{nF}$, $C_2 = 20\text{nF}$

(trimmed)

$R_1 = 324\text{k}\Omega$

$R_2 = 162\text{k}\Omega$ (trimmed)

$f_1 = 33\text{Hz}$ $f_2 = 76\text{Hz}$



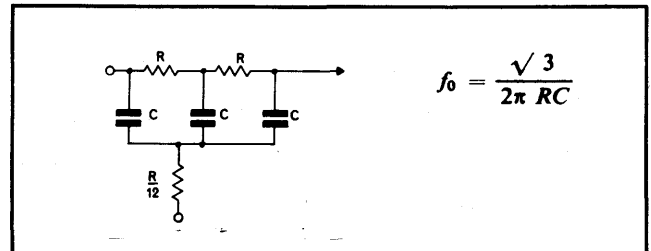
Circuit description

The passive network has a perfect null at a defined frequency if the components are accurately matched. The rate of approach to that notch is sharpened by positive feedback, the buffer amplifier having a negligible output resistance and hence theoretically not disturbing the depth of the notch. Variation in notch frequency requires simultaneous variation e.g. of all three capacitors or of R_2 and C_2 and the circuit is most suitable for fixed frequency operation. In addition too much positive feedback ($k \rightarrow 1$) may give unsatisfactory results.

Component changes

With low values of C_1 gain at high frequencies may not be

0dB. A 12% deviation from the nominal 0dB was observed with $C_1 = 270\text{pF}$, $R_1 = 600\text{k}\Omega$, $n = 1$, $f_0 = 1\text{kHz}$. Very large resistors ($> 10\text{M}\Omega$) should be avoided.



Circuit modifications

● Any circuit with a notch e.g. above may be used in place of the twin-T network e.g. ref. 1. Trimming of $R/12$ may be necessary if k is varied over a wide range.

● Notch filters having a narrow notch width can be unsatisfactory with signals whose frequency stability is not good. This can be overcome by cascading two notch filters, having slightly different notch frequencies e.g. two filters tuned to 48 and 52Hz respectively will effectively remove mains pick-up². It may be possible to simply cascade the passive networks with due attention to loading of the first by the second, and still only use two op-amps.

● Single component variation of the notch frequency is possible with a twin-T network (card 3 and ref. 3) and is also possible with other networks (card 10). Buffer amplifier may be omitted if potentiometer value is low.

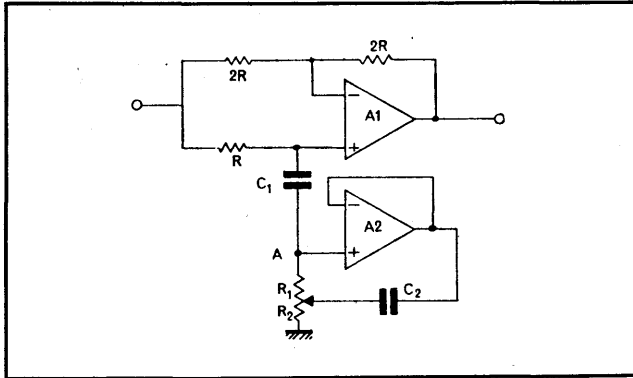
Further reading

1. Graf. Electronic Design Data Book, Van Nostrand Reinhold.
2. N. B. Rowe, Designing a low frequency active notch filter, *Electronic Engineering*, April 1972.
3. Douce and Edwards, *Electronic Engineering*, July 1964.
4. National Semiconductor Linear Brief LB-5.

Cross references

Series 1, cards 3 & 10.

Easily-tuned notch filter



Typical performance

$$Q \text{ of resonant branch is } \frac{1}{R_1 + R_2} \sqrt{\frac{C_2 R_1 R_2}{C_1}}$$

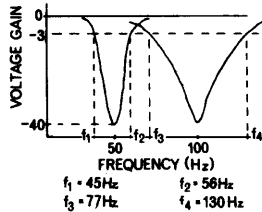
$$f_0 = 1/2\pi \sqrt{C_1 C_2 R_1 R_2}$$

$$f_0 = 50\text{Hz with}$$

$$R_1, R_2 : 50\text{k}\Omega, C_2 : 1\mu\text{F}$$

$$C_1 : 4\text{nF} \quad R : 100\text{k}\Omega$$

Results shown are for these values, with some trimming of one of the $2R$ resistors for good notch depth. 741 op-amps were used. Considerable trimming is necessary at higher frequencies.



Circuit description

The impedance between A and ground is equivalent to an inductor, $C_2 R_1 R_2$, in series with a resistor, $R_1 + R_2$. At a specific frequency C_1 and the equivalent inductance series resonate, leaving the equivalent resistance to bring the bridge into balance provided $(R_1 + R_2) = R$ i.e. an overall notch characteristic. Variation in the equivalent inductance or in C_1 changes the notch frequency with no theoretical change in notch depth. In practice component imperfections, circuit strays etc. may require significant departures from bridge nominal resistances to reach balance, particularly at frequencies above 100Hz. If the circuit Q is large, A2 will saturate at the notch frequency unless V_{in} is kept low. At frequencies of the order of 1kHz or more the notch depth and Q are sensitive to resistance in series with C_2 . A low-loss capacitor and low-contact-resistance potentiometer should then be used (avoid potentiometers with non-metallic contacts).

Component variations

R: $50\text{k}\Omega$ to $1\text{M}\Omega$

C_2 : $10\mu\text{F}$ to $0.1\mu\text{F}$

C_1 : 40 nF to 100pF .

Circuit modification

R_1 and R_2 may be kept as fixed resistors if a variable capacitor for C_1 is available.

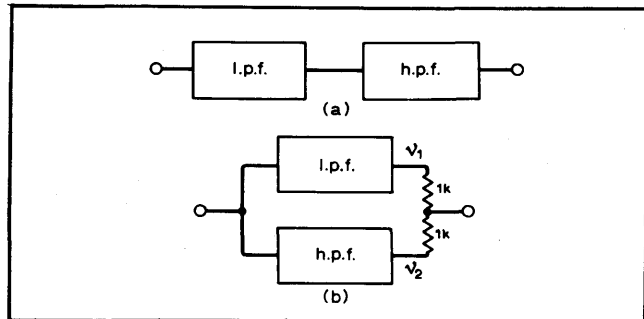
Further reading

National Semiconductor LM307 Data Sheet.

Cross references

Series 1, card 9.

Compound filters



(a) provides the band-pass characteristic shown. Low and high-pass filters used are described in cards 4 and 5 with cut-off frequencies 10kHz and 2.35kHz respectively. $f_1 = 5\text{kHz}$, $f_2 = 1.9\text{kHz}$ and $f_3 = 13\text{kHz}$.

(b) provides the band-stop characteristic shown. In this case the low and high-pass filters had cut-off frequencies 0.76kHz and 12kHz respectively.

Circuit description

The basic filter functions described may be combined in a variety of ways to produce alternative functions or to improve on existing ones. The two examples shown illustrate ways of

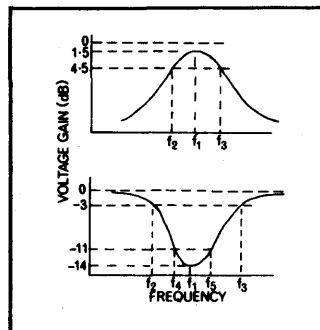
$$v_0 = (v_1 + v_2)/2,$$

$$f_1 = 3\text{kHz}, f_2 = 0.76\text{kHz},$$

$$f_3 = 12\text{kHz},$$

$$f_4 = 2.1\text{kHz} \text{ and}$$

$$f_5 = 4.5\text{kHz}.$$



producing simple band-pass and band-stop filters from low-pass and high-pass sections. Cascading l.p. and h.p. filters with the l.p. filter cut-off frequency above that of the h.p. filter gives a pass-band lying between these two frequencies, and for second-order filters, an attenuation approaching 12dB/octave outside this range. Multi-order l.p. and h.p. filters may be used to increase this attenuation.

Conversely if the filter responses are added with the l.p. filter cut-off frequency being the lower the response is minimum between the cut-off frequencies with the depth of the attenuation dependent on the separation. Simple resistor summing, or a virtual earth summing amplifier may be used.

Circuit modifications

One can cascade second-order low-pass and high-pass sections to achieve steeper sides to the stop and passbands. This only produces even-order filters with poor roll-off characteristics. Improved higher-order filters can easily be used (see ref.).

One can also cascade the bandstop characteristic with a notch filter to achieve precise nulling of a fixed frequency with reasonable attenuation on either side of the notch. Several notch filters tuned to slightly different frequencies may be used to obtain a bandstop characteristic (card 9). Several band-pass filters (card 1) can be similarly cascaded to give a good bandpass characteristic.

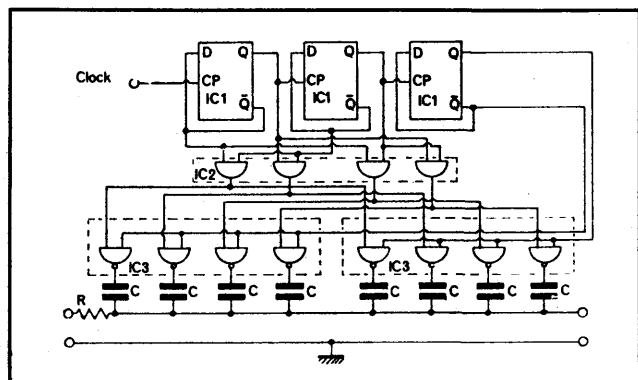
Component changes

Any low-pass filter or high-pass filter may be used in the above configurations to provide the band-stop and band-pass characteristics.

Further reading

Graeme, J. G., Tobey, G. E. & Huelsman, L. P. Operational amplifiers. McGraw-Hill p.320.

N-path filter



Typical performance

IC₁ 3 SN7474 D-type flip flop

IC₂ 2 off SN7400 connected as two-input AND gates

IC₃ SN7401 quadruple two-input NAND gates with open collector output.

R: 1kΩ ±1%

C: 0.1μF ±10%

$f_0 = 14.9\text{kHz}$ $Q = 30$

Drive signal 600mV pk-pk.

Resonant frequency

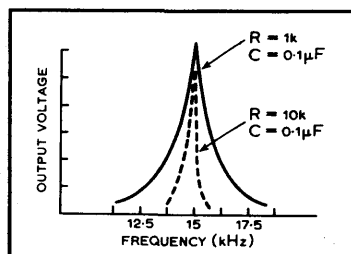
$f_0 = \text{clock frequency}/N$

Bandwidth: $2/NRC$ Hz,

$N = \text{number of low-pass filter sections} = 8.$

Peak on graph:

460mV pk-pk.



Circuit description

The input signal is effectively switched between eight low-pass filter sections at a clock rate, eight times the required filter centre frequency. The output waveform comprises discrete levels approximating the input waveform. This stepped format may be removed by a low- Q bandpass filter.

The resonant frequency can be varied simply by changing the clock frequency thus giving a tunable bandpass filter with constant bandwidth.

Component changes

Increasing R for same C values, increases Q e.g. $R = 10\text{k}\Omega$ ±10%, $Q = 200.$

IC₁ 3 MC7479; IC₂ MC3001; IC₃ MC7403.

Circuit modifications

Due to the sampling action of this filter, responses are obtained at multiples of the clock frequency. If a 4-path filter with a fundamental resonant frequency f_0 is cascaded with a 3-path filter with a fundamental resonant frequency $2f_0$, a single bandpass characteristic is obtained, centred on $2f_0$ (ref. 2).

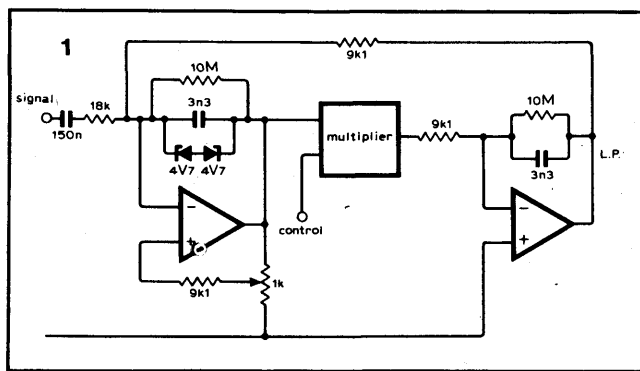
Further reading

1. Broeker, B. Commutating filter techniques *Design Electronics*, 1971, pp.52-5.
2. Roberts, J. N-path filter techniques *Electron*, June 1972, pp.18-9.
3. Bruton, L. T. and Pederson, R. T. Time-multiplexed active filters, *IEEE Journal of Solid-State Circuits*, vol. SC-7, June 1972, pp.259-65.

Basic active filters

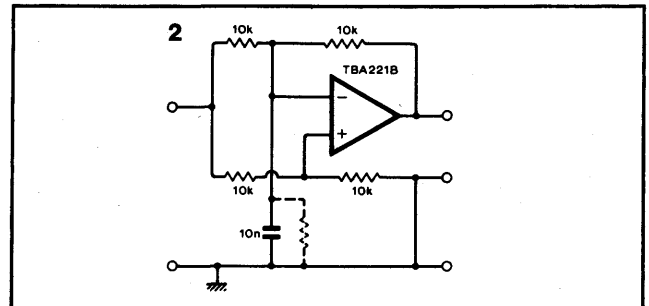
1. Voltage control of active filters is achieved using an analogue multiplier (see Audio circuits, set 5, for details). A loop consisting of a bandpass filter and an integrator has the signal coupled between them by the multiplier acting as a voltage controlled amplifier. The reference article also shows a square-law generator for shaping the frequency/control voltage characteristics. In this configuration the 1-k Ω potentiometer sets the circuit Q, and the direct control voltage the centre frequency. A low-pass output is also available. The zener diodes prevent overdrive from latching the circuit into a permanently switched state. The article also indicates how the circuit can be modified into an RC oscillator with stabilized amplitude, and a notch filter.

Orr, T. and Thomas, D. W. Electronic sound synthesizer, part 2. *Wireless World*, vol. 79, 1973, pp.429-34.



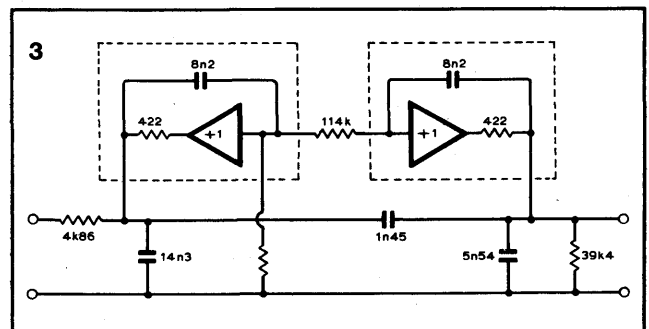
2. The limited amplitude-frequency response of operational amplifiers restricts the upper frequency at which defined filter characteristics can be obtained; it may result in instability/oscillation in some filters. As many op-amps, such as that shown here (and also 741, 307 etc), have a frequency response defined by a single dominant lag internal to the amplifier, then some second-order filters are possible using just one external capacitor. The reference article shows that the resulting transfer function is that of a band-pass filter with centre-frequency $\approx 31\text{kHz}$, centre-frequency gain ≈ 115 and $Q \approx 11.5$. Adding a resistor across the capacitor controls Q and gain without varying the centre frequency. The filter characteristics are supply and temperature dependent - the former can be turned to advantage in trimming centre frequency if variation in the other parameters is acceptable.

Radhakrishna Rao, K and Srinivasan, S. "Bandpass filter using the operational amplifier pole", *IEEE Journal of Solid-State Circuits*, June 1973, pp.245-6.



3. The design of higher-order filters can be based on cascaded first/second-order sections. Alternative approaches, economical of active and passive elements, use interconnected networks of impedance convertors (n.i.cs, gyrators etc) and passive networks. All the parameter values interact but the approach is very efficient and capable of very low sensitivity to component variation. The example shown uses blocks within dotted lines that act as impedance convertors to simulate lossy inductances, and results in a third-order low-pass elliptic filter with 1.0dB ripple in the passband, 30dB loss in the stopband and with a passband edge frequency of 3.4kHz. Other networks are shown in the article generating higher-order filters.

Wise, D. R. Active simulation of floating lossy inductances, *Proc. IEEE*, vol. 121, 1974 pp.85-7.



Wireless World Circard

Series 2: Comparators & Schmitts

Though the introductory article was previously called "Switching circuits", this set of cards deals with comparators—exhibiting no hysteresis or backlash—Schmitt trigger circuits—having a finite hysteresis—and level detectors, which can belong to either class. Switching circuits of the astable kind are dealt with in Series 8 in this book; logic gates, digital counters, pulse modulators, and monostable circuits have been covered in later cards, available separately.

One of the two comparator circuits is interesting (card 4), not only in that it can be simply modified to act as a Schmitt trigger circuit, but because it shows how to use op-amp supply leads as signal outputs which may be used to drive push-pull circuits (see also series 7 cards 2 and 12). The other circuit (card 1) shows how to clamp the output of an op-amp at low current levels for op-amps with access to the output stage drive point.

Most of the other cards show variants of the Schmitt circuit, including three with low hysteresis (cards 2, 9 and 11).

- Op-amp comparator/Schmitt (bipolar clamping) 1
- Basic Schmitt circuit 2
- Complementary m.o.s. Schmitt 3
- High-power comparator/Schmitt 4
- Unijunction-equivalent Schmitt 5
- Variable-hysteresis level detector 6
- High-speed Schmitt circuit 7
- TTL Schmitt circuit 8
- Low-voltage level sensor 9
- Reference-controlled hysteresis circuit 10
- Window detector 11
- Complementary Schmitt 12

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Comparators & Schmitts

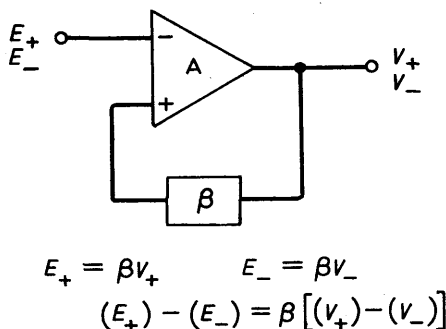
There is a need for circuits whose output changes by a large amount as the input passes through a particular level. There are four cases:

1. The change may be reversible without hysteresis, and those high-gain amplifiers, called comparators belong to this class.
2. The change may be reversed at a different value of input, i.e. exhibit hysteresis, and Schmitt trigger circuits are examples. The output may fail to return to its original value when hysteresis is present, because the input is constrained within limits preventing such reversal, and such circuits are said to "latch".
3. The output, because of some intervening unidirectional device or circuit, may remain in its second state indefinitely regardless of any further excursions of the input unless returned to its original state by some alternative process, and bistable circuits such as the classic Eccles-Jordan belong to this category.
4. Finally, the circuit may respond to the second or succeeding excursions of the input through a given level in a given sense, and such circuits are used for dividing and counting in logic systems.

In types 3 and 4 above, the normal applications are such that precision of switching level is unimportant, as the systems in which they are used are digital, generally binary. Hence the input and output levels, need only be controlled within broad limits and can still be identified unambiguously. While precision in the switching levels may be combined with these other functions it is more usual to separate the functions as design constraints are so different.

Switching circuits to be considered here belong to types 1 and 2 above and can constitute an interface between analogue and digital systems — discrete changes in output are obtained at specified amplitudes of input.

Type 1 may be further sub-divided according to whether the input is differential or single-ended with respect to some prescribed level, often ground potential. The former is readily available in the form of integrated-circuit comparators with excellent performance. Early versions were designed for high-speed operation to be used in conjunction with particular logic families. Supply voltages are required to be within close limits of specific values not always compatible with those in common use for other purposes, e.g. +12/-6V as



against $\pm 15V$ for many analogue systems. For this reason newer versions have appeared capable of working from a wide range of supply voltages, and having lower input current requirements. Yet others are being produced with higher switching speeds.

Voltage gain of these comparators is high, with the full output swing being achieved for an input change of a few millivolts. Thus if one input terminal is taken to some constant reference voltage with the other input fed from the signal source, a sharp output transition occurs when the signal voltage exceeds that of the reference. Finite voltage gain together with offset (unbalance) effects limit the precision achievable at low cost to a few millivolts.

The indeterminate value of output for inputs close to the reference level, makes the use of low-gain amplifiers inappropriate as comparators, and discrete circuits with two or three transistors are less commonly found as type 1.

If positive feedback is applied to any amplifier then under the right conditions the output can be made to switch between two distinct states with little or no further change in output regardless of further variation of the input. This applies to low- and high-gain amplifiers alike, though with the former the magnitude of the feedback factor must be larger to ensure a complete switching action. The margin must be large enough to guarantee the switching action in the presence of parameter variations due to supply and environmental changes for a particular unit, and to cover component tolerances.

A further property resulting from the use of positive feedback is that the output transition in the positive and negative inputs occur at different values of input — the effect being called hysteresis of the circuit. This is illustrated in the figure where

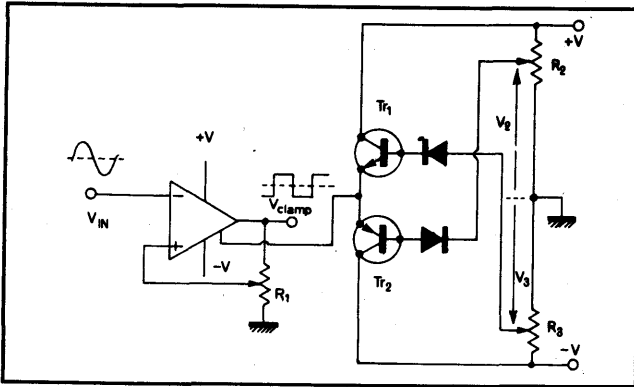
V_+ and V_- are the alternative output voltages of a comparator (or operational amplifier where high speeds are not critical) when fully switched. The input voltages E_+ and E_- are the voltages at which transitions occur. The output will switch from V_+ to V_- when the input exceeds βV_+ by a small amount of the order (V_+/A) , where β is the fraction of the output voltage fed back and A is the voltage gain of the comparator in its active region.

Hence the hysteresis is about $\beta \Delta V$, where ΔV is the change in the output, and provided A is large.

In various applications it may be necessary either to minimize the hysteresis or to increase it and define it. The former requirement indicates the need for a high value of A ; the latter for an accurately defined product $\beta \Delta V$ and either large A or at least a defined value of A . Deciding on the precise value of A to be used in such calculations is difficult — the gain continually changes as the critical point is approached. Using i.c. comparators (or operational amplifiers at lower frequencies) the value of A is sufficiently high that the hysteresis can be defined by the resistive feedback network.

Discrete amplifiers, of which the Schmitt trigger is the classic version, are capable of a very wide range of characteristics with complementary versions increasing the choice of characteristics. A problem frequently encountered in such designs is that the switching levels, hysteresis etc, are often interdependent and have to be pre-selected using suitable algebraic equations. It may not be easy to achieve independent control of these parameters.

Op-amp comparator/Schmitt (bipolar clamping)



Typical performance

Supplies: $\pm 5V$; IC: 748
 Tr₁: BC125; Tr₂: BC126
 Diodes: 1N914
 R₁, R₂, R₃: 10k Ω
 With $V_{in} = 1V$ pk-pk
 at 10kHz, V_{out} adjust-
 able ± 1.6 to $\pm 4.4V$.
 Rise and fall times
 $\approx 700ns$.

Supply currents
 $\pm 2.5mA$
 Max $V_{in} = \pm V$.
 For $0 < V_2 < 3V$
 $V_{out}^+ = V_2 + 3V_{be}$
 $V_{clamp}^+ = V_{out}^+ - V_{be}$
 For $-3V < V_3 < 0$
 $V_{out}^- = V_3 - 3V_{be}$
 $V_{clamp}^- = V_{out}^- + V_{be}$

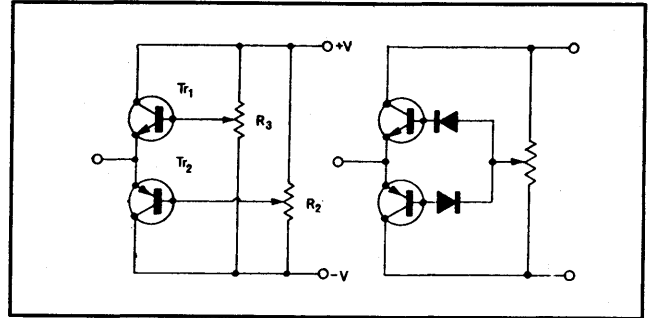
Circuit description

Operational amplifiers used as comparators (or as level-sensing circuits when positive feedback is used for hysteresis) have output swings which vary with temperature and from unit to unit. Some amplifiers have access to the drive point of the output stage and this point may be clamped at selected positive and negative potentials by zener diodes or suitably biased transistors as shown. Clipping is at much lower current levels than if attempted directly at the output. Variable resistors R₂ and R₃ set the positive and negative clamping levels, R₁ determines hysteresis at the clamping levels. Diodes provide base-emitter breakdown protection.

Component changes

Useful range of supplies: ± 5 to $\pm 18V$.
 Transistors: general-purpose silicon types.
 Useful range of R₁: 1 to 100k Ω .
 Useful range of R₂, R₃: 1k Ω (increases supply current drawn) to 100k Ω (produces error in V₂ and V₃ unless base current loading is reduced by use of higher-gain transistors).

Useful frequency range: d.c. to approx. 160kHz. If a higher-speed operational amplifier is used, transistors may limit the frequency response unless high-speed versions are used. Diodes may be omitted for low supply voltages and/or high reverse base-emitter breakdown transistors. Base-emitter junctions may receive supply voltage at extreme settings of R₂ and R₃.



Circuit modifications

- Diodes could be placed in series with the emitters of the transistors. This still provides base-emitter breakdown protection but the diodes then carry the larger emitter currents producing larger diode p.d.s.
- R₂ and R₃ could be connected as shown on left allowing the base potentials of both Tr₁ and Tr₂ to be set positive or negative independently. It would then be possible for Tr₁ base to be positive and Tr₂ base to be negative, which would allow excessive conduction in Tr₁, Tr₂.
- Fig. on right shows a modification which allows the mean level of V_{out} to be set positive or negative by the potentiometer with its peak-to-peak value still determined by $\pm 3V_{be}$.

Further reading

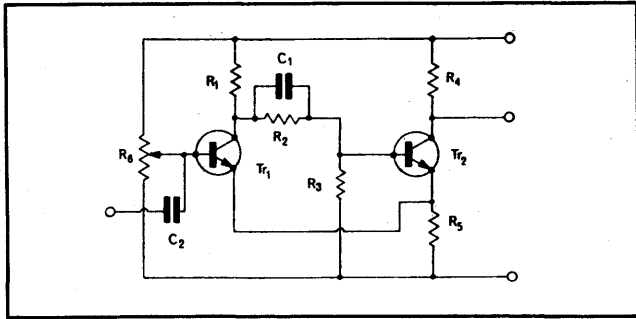
IC op.amp beats fets on input current, National Semiconductor application note AN-29, 1969, p.15.
 Clayton, G. B., Operational Amplifiers, Butterworth 1971, pp.145-9.
 Applications Manual for Operational Amplifiers Philbrick/Nexus Research, 1968, pp.59 & 101.

Cross references

Series 2, cards 4 & 6.

Wireless World Circard Series 2: Comparators & Schmitts 2

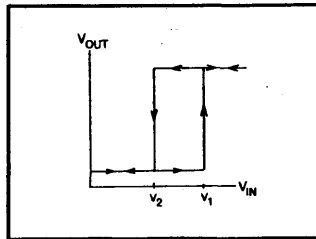
Basic Schmitt circuit



Typical performance

Transistors: BC125
 Supply: +5V
 R_1, R_4 : 4.7k Ω
 R_2, R_3 : 2.2k Ω
 R_5 : 1k Ω \pm 5%
 R_6 : 4.7k Ω pot.
 C_1 : 100pF (speed-up)
 C_2 : 2.2 μ F
 Signal level from 500- Ω
 source: 2V pk-pk.

Output swing: 1.5 to 5V
 up to 100kHz.



Circuit description

Emitter coupling between Tr_1 and Tr_2 introduces positive feedback causing a regenerative switching action into one of two states. When V_{in} is below threshold level V_1 , Tr_1 is non-conducting and Tr_2 conducts, the base voltage being determined by R_1 , R_2 and R_3 . The emitter potential is then well-defined. As input voltage exceeds V_1 , Tr_1 begins to conduct, reducing its collector potential, and hence that of the base and emitter of Tr_2 . This drop in potential is fed back to the emitter of Tr_1 , thus further increasing the conduction of Tr_1 until Tr_1 is on and Tr_2 is off. A similar regenerative action occurs when the input voltage is reduced below the threshold level V_2 , returning the circuit to its original condition. A typical input-output voltage characteristic is shown above, where $V_1 - V_2$ is termed the hysteresis or backlash of the circuit.

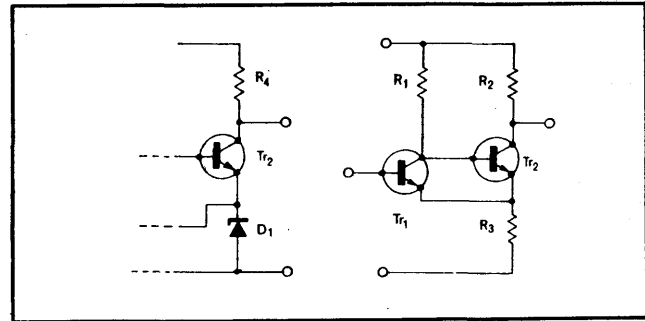
Component changes

Resistor R_6 permits adjustment of threshold level V_1 . Useful ratio R_2/R_3 is in the range 0.5 to 2.0, giving control of trip level and hysteresis range of 0.2 to 0.8V.

Useful C_1 range: 1nF to 100pF. Optimum 100pF at a source frequency of 1MHz.

Rise time: 150 to 100ns using oscilloscope probe.

For $R_4 = 1k\Omega$, $R_1 = 100\Omega$, $R_2/R_3 = 0.5$, output swing at 1MHz is 3 to 5V, for load capacitance up to 33pF.



Circuit modifications

The emitter resistor may be replaced by a zener diode (left). This means the emitter potential variation is less dependent on current flow through each transistor. Typical performance:

$V_{CC} = 10V$, D: 5.1V zener diode. R_1 : 680 Ω , R_2, R_4 : 1.5k Ω , R_3 : 3.3k Ω . Drive signal: 2V pk-pk; output swing: 5.2 to 10V. Hysteresis: 110mV.

On right, useful range of R_3 : 10 to 500 Ω . Useful range of R_2 1 to 2k Ω . R_1 : 1k Ω . Typical performance:

$V_{CC} = 5V$, $R_E = 10\Omega$, $R_1 = R_2 = 1k\Omega$. Minimum sinusoidal drive signal at 100kHz: 2V pk-pk. Output swing: 0.8 to 5V.

Frequency may be increased to 300kHz if drive voltage is increased to 4V pk-pk.

Reference

1. Zero-hysteresis Schmitt trigger, in *Electronic Circuit Design Handbook*, 4th edition. 1971 p.108.

Further reading

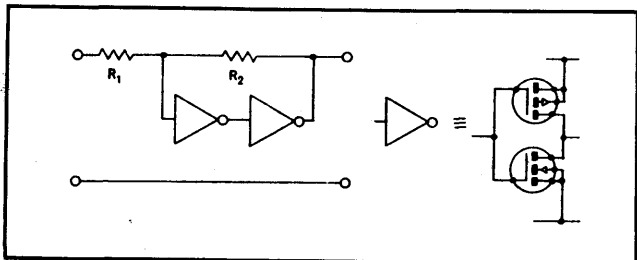
Crump, A. E., Design of Schmitt trigger circuits, *Wireless World*, vol. 73, 1967, pp.122-7 and 175-7.

Circuit Consultant's Casebook, Hemingway, T. K., Business Books, 1970, pp.129-37.

Cross references

Series 2, cards 3, 7 & 8.

Complementary m.o.s. Schmitt



Typical performance

IC: CD4007AE (connected as triple inverter)	V_1 : 5.9V
Supply: 10V	V_2 : 5.1V
R_1 : 1M Ω	Output swing: 10V
R_2 : 10M Ω	Input current: $\pm 0.5\mu\text{A}$

Circuit description

Two c.m.o.s. inverters are cascaded with positive feedback defined by ratio R_2/R_1 . Provided this ratio is less than the forward gain in the inverters' linear region, switching follows the appropriate input changes. Output swing approaches supply lines and current from source is small as very high input resistance of inverter allows R_1 , R_2 to be large. With small hysteresis switching levels are near supply mid-point.

Component changes

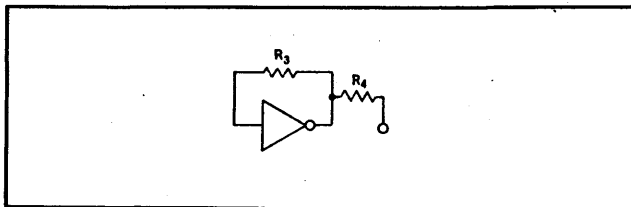
- Any combination of inverters, gates or buffers giving voltage gain $> +1$ may be used. Examples: RCA CD4001AE, Motorola MC14001 quad 2-input NOR gates; CD4049AE hex buffer inverters.
- Supply voltage +3 to +15V (special versions down to 1.5V).
- R_2/R_1 may be varied between 1 and 100. At upper end of range positive feedback may be too little to guarantee switch-

ing. At lower end hysteresis is comparable to supply voltage.

- To minimize capacitive effects/hum pickup reduce R_1 , R_2 to $\sim 10\text{k}\Omega$. Lower values reduce output swing and accuracy of hysteresis, while increasing current from source.

Circuit modifications

- Buffer input with third inverter/gate increasing input resistance (typical input current $\sim 10\text{pA}$). Resistor R_1 may be dispensed with, the output resistance of buffer taking its place, with R_2 reduced to range 1 to 30k Ω . Resulting hysteresis in range 2.5 to 0.2V.
- Use spare inverter self-biased by large resistor ($\sim 10\text{M}\Omega$) (see Fig.) to bias input terminal of first inverter via second resistor ($\sim 10\text{M}\Omega$). This sets mean potential near to centre of linear region, assuming well-matched inverters. Signals may now be a.c. coupled and 200mV pk-pk typically triggers circuit over a range of supply voltage and temperature with no adjustment of bias level.



Further reading

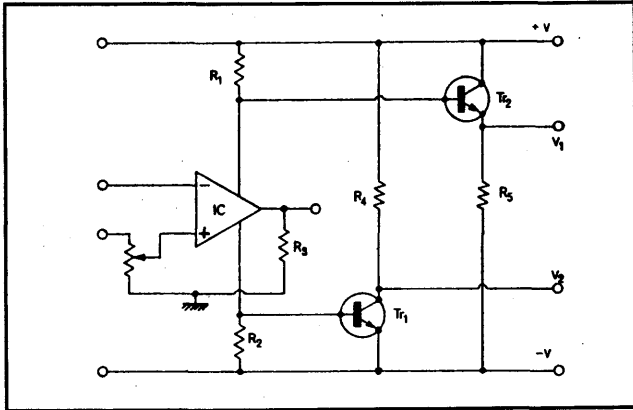
Dean, J. A. & Rupley, J. P., Astable and monostable oscillators using RCA COSMOS digital integrated circuits, RCA application note ICAN-6267.
Schmidt, B., Schmitt trigger design uses CMOS logic, *Electronic Design*, Vol. 20, 27 April 1972, p.72.

Cross reference

Series 2, cards 1, 2 & 8.

Wireless World Circard Series 2: Comparators & Schmitts 4

High-power comparator/Schmitt



Performance data

IC: 741; Supplies $\pm 5V$
 T_{r1} : BFR41, T_{r2} : BFR81
 R_1, R_2 : 220 Ω , R_3 : 270 Ω
 R_4, R_5 : 68 Ω , 3W

Minimum $V_{in} = 800mV$
 pk-pk.
 Max. $\pm V = \pm 7V$,
 for full output swing into
 68 Ω loads

With maximum permissible sinusoidal input of 8V pk-pk, V_1 and V_2 are both square waves swinging between -5 and +4.8V, and -5 and +5V respectively. V_1 and V_2 are in-phase while the currents in R_4 and R_5 are in anti-phase. Max. frequency 1kHz—waveform squareness is lost at higher frequencies.

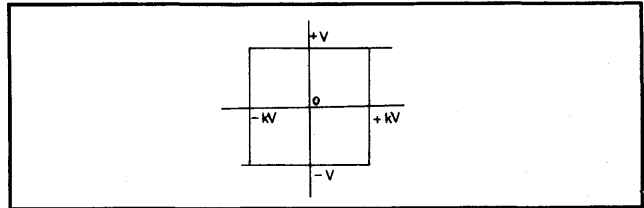
Circuit description

The output swing of standard op-amps is significantly less than the supply voltage, particularly when the latter is low. The current available is also low. With some op-amps the current in the positive supply lead is large when the output voltage is positive and the output current is large but is small when the output voltage is negative. The negative supply current behaves similarly. The change in supply current as the input signal varies can be used to drive following transis-

tors which may supply currents of several hundred milliamperes at a voltage very close to the supply voltage.

Component changes

- To maximize output voltage swings T_{r1} and T_{r2} must be driven into saturation i.e. base currents of 5 to 10% of load current are required. Reducing R_3 will increase base current.
- Resistors R_1 and R_2 may need to be reduced for some op-amps having larger off-load currents (180 Ω was found satisfactory for a 748).



Circuit modifications

The circuit can be altered to give a Schmitt characteristic with controllable hysteresis by connecting either V_1 or V_2 via the pot. shown to the non-inverting input of the op-amp. Using V_2 we obtain the characteristic shown right in which k is the pot tapping and V the supply voltage (5V). This hysteresis is not dependent on the saturation level of the amplifier as it would be if the amplifier output were fed back. Hysteresis width is controllable up to about 0.1V of the pk-pk value of the input provided the input is kept below about 5V pk-pk.

Further reading

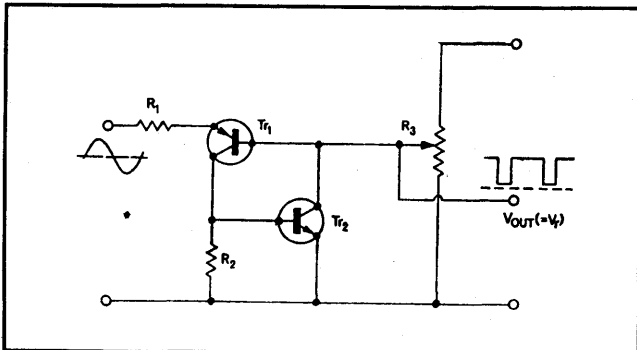
Campbell, D. L. and Westlake, R. T. Build a high current servoamplifier with i.cs. *Control Engineering*, December 1969, p.91.

Widlar, R. J., IC op-amp beats f.e.ts on input current, National Semiconductor application note AN-29, 1969, p.15.

Cross references

Series 2, card 9.

Unijunction-equivalent Schmitt



Typical performance

Supply: +5V	$V_{in}(\text{off}) = 3.0\text{V}$
Tr ₁ : BC126 Tr ₂ : BC125	Supply current: 8mA
R ₁ , R ₂ : 1kΩ R ₃ : 4.7kΩ	(on), 1mA (off)
$V_{out}(\text{on}) = 0.04\text{V}$	$V_{in}(\text{on}) =$
$V_{out}(\text{off}) = V_r = 2.5\text{V}$	$V_r + (n+1)V_{be}(\text{on})$
$V_{in}(\text{on}) = 3.8\text{V}$	where $n = R_1/R_2$

Circuit description

The transistors together have properties similar to those of a unijunction transistor. When V_{in} is low Tr₁ and Tr₂ do not conduct and V_r is defined by R₃. For $V_{in} \approx V_r + 1.3\text{V}$, Tr₁ and Tr₂ begin to conduct, regenerative switching via Tr₂ clamping V_{out} close to 0V. Reversal of switching occurs when V_{in} falls. Significant current is drawn from the source unless a limiting resistor (R₁) is included.

Component changes

Tr₁: any general purpose p-n-p silicon transistor.

Tr₂: any general purpose n-p-n silicon transistor.

Maximum useful frequency $\approx 100\text{kHz}$.

Range of V_r source resistance (seen by Tr₁ base): about 2.2kΩ to 33kΩ.

$R_1(\text{max}) \approx 3.3\text{k}\Omega$. For large R₁ values V_r source resistance must be increased for rapid switching action. The output can only be lightly loaded with large V_r source resistance.

Circuit modifications

If the input voltage is fed directly to Tr₁ emitter the circuit may be used to clamp it to a low level (about 0.7V with a 5V supply) when it exceeds some maximum permissible level. For example, V_r could be the output from a voltage regulator and V_{in} its input voltage. If V_{in} (regulator input) rises excessively the circuit will rapidly clamp the regulator input to a low value protecting the regulator and the circuitry it supplies during the time taken for the supply fuse to blow. The transistors require a current rating greater than the supply peak current on s.c. loading. R₂ and R₃ may then need to be reduced.

Further reading

General Electric transistor manual, 1964, Chapter 13.

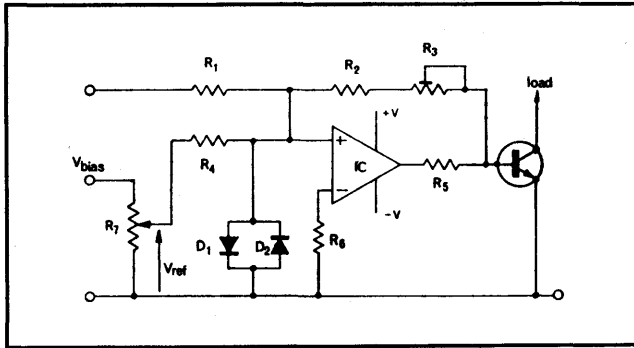
Unijunction transistor timers and oscillators, Motorola application note AN-294 (appendix), 1967.

Cross references

Series 2, cards 2 & 12.

Wireless World Circard Series 2: Comparators & Schmitts 6

Variable hysteresis level detector



Component data

Supplies: $\pm 15V$
 R_1, R_4 : $2.2k\Omega$
 R_2 : $100k\Omega$
 R_3 : $100M\Omega$
 R_5 : $3.3k\Omega$
 R_6 : $1k\Omega$
 R_7 : $1k\Omega$

D_1, D_2 : general-purpose diodes.
 IC: 741
 Tr: ME4103 (in general determined by load current requirements)
 All resistors $\pm 5\%$.

Circuit description

V_{ref} adjusts the level at which the output switches without affecting hysteresis. Positive feedback path R_2 and R_3 provides hysteresis controllable by R_3 . Sensitivity can be modified by changing input resistance R_1 . Positive output swing is determined by the base-emitter voltage of the transistor and the negative output by the particular operational amplifier used. Diodes on the input provide breakdown protection of the op-amp against excessive input voltages.

Component changes

IC: 748 or LM301A.

For V_{ref} of $-1V$ to $-14V$, $R_3 = 0$, supplies: $\pm 15V$;
 hysteresis: $180mV \pm 2\%$;
 trip level: $V_{ref} + 200mV$.

For V_{ref} of $-1V$ to $-4V$, $R_3 = 0$, supplies: $\pm 5V$;
 hysteresis: $700mV \pm 5\%$;
 trip level: $V_{ref} + 100mV$;

Hysteresis: $10mV$, $R_3 = 1M\Omega$, supplies: $\pm 10 \rightarrow \pm 15V$; trip level: $V_{ref} + 100mV$.

In general, hysteresis may be further increased by reducing $R_2 + R_3$.

Circuit modifications

- If output voltage swing required at lower currents, Tr may be omitted and R_5 reduced to zero. Hysteresis is then controlled by op-amp output swing.
- Alternative methods of defining output swing and hence hysteresis include series back-to-back zener diode or diode limiting circuits.
- For higher speed operation, IC may be any comparator.
- For higher output currents, Tr may be replaced by a Darlington pair. If only an indication of output state is required, most op-amps can deliver sufficient current to drive small light-emitting diodes.

Further reading

Linear Integrated Circuits Handbook, Marconi-Elliott Microelectronics, pp.167/8.

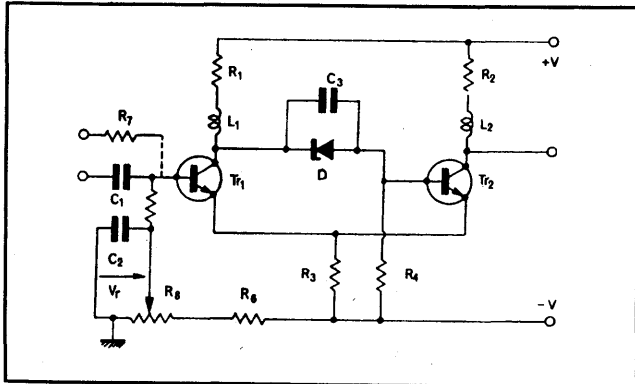
Smith, J. I., Modern Operational Circuit Design, Wiley, 1971, pp.186/7.

Cross references

Series 2, cards 9-11.

Wireless World Circard Series 2: Comparators & Schmitts 7

High-speed Schmitt circuit



Typical performance

Supplies: $\pm 5V$
 $V_r = -1.0V$; $V_{in}(Tr_{2on}) = -2.02V$; $V_{in}(Tr_{2off}) = -0.31V$
 $Hysteresis: 1.71V$
 $V_{out(on)} = 0.2V$
 Supply current
 $+40mA, -50mA$ Tr_{2off}
 $+43mA, -51.5mA$ Tr_{2on}

$Tr_1, Tr_2: BSX20$
 $D_1 = BZX55, C3V9$
 $R_1, R_2, R_3, R_5, R_7: 100\Omega$
 $R_4: 10k\Omega$
 $R_6: 3.3k\Omega; R_8: 1k\Omega$
 $C_1, C_2: 0.1\mu F; C_3: 27pF$
 $L_1, L_2: 0$

Circuit description

For maximum speed, Schmitt trigger circuits should operate with the transistors out of saturation at all times. A zener diode in the bias network can assist this. In this circuit, current levels are higher than in the basic Schmitt to maximize gain-bandwidth product. The inductors compensate for capacitive loading to optimize rise time. The upper and lower

thresholds are negative and the hysteresis is variable but is not independent of the threshold levels.

Component changes

Useful range of V_r : 0 to $-1.47V$

Corresponding hysteresis range: 2.9 to 1.64V.

L_1 and L_2 can be adjusted to produce a required rise time with a defined overshoot for given capacitive loading. The same principle applies to the complementary Schmitt. With $L_1 = L_2 = 0.11\mu H$, rise time $< 8ns$ with 5% overshoot at low switching rates.

The circuit functions to at least 40MHz with defined output levels although the waveform is rounded at high frequencies. Careful printed circuit layout is necessary for good high-frequency operation.

Circuit modifications

Precise adjustment of the negative rail voltage allows the output to be made truly t.t.l.-compatible with levels of 0V and +5V. The output from Tr_2 may be used to feed a high-speed t.t.l. gate or an e.c.l. gate to "square up" the waveform at high frequencies. Circuits of this type may be useful in conjunction with t.t.l. or e.c.l. circuitry as they provide alternative options of switching levels and hysteresis. To assist supply decoupling at high frequencies, ferrite beads can be added to the supply line wiring.

Further reading

E-Line Transistors Applications, Ferranti, 1969. p.20.

MECL Integrated Circuit Schmitt Triggers, Motorola application note AN-239.

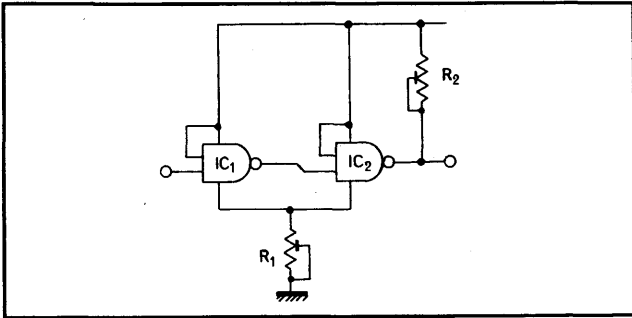
Cross references

Series 2, card 2 & 8.

Wireless World Circard

Series 2: Comparators & Schmitts 8

TTL Schmitt circuit



Performance data

Graph obtained with
 $R_2: 250\Omega$, $R_1: 30\Omega$
 Supply: 5V
 ICs: 7400

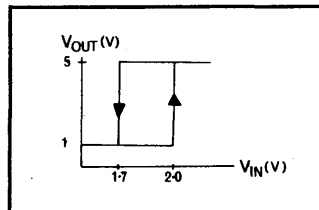
Frequency 0 to 1MHz.
 Threshold values and hence hysteresis may be changed slightly by varying R_1 and R_2 .

Lower limit (1V as shown) is affected by R_1 and R_2 .

Circuit description

Each NAND gate with one input gate disabled behaves as an inverter. The circuit with positive feedback via R_1 is very similar to the basic Schmitt trigger as each gate is essentially identical. This results in the potential across R_1 being constant

With $R_1 = 0$ there is no positive feedback and switching is not clean. With $R_2 = \infty$ upper limit is reduced from 5V.



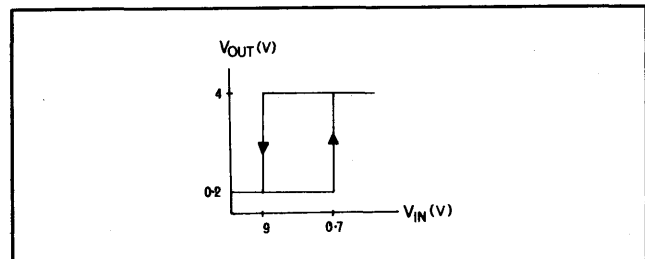
and independent of which inverter is enabled. This results in an offset voltage compensated by R_2 .

Component changes

Any t.t.l. inverter may be used.

Circuit modifications

An alternative t.t.l. Schmitt is SN7413, produced by Texas, and has two in a single package. Typical characteristics are shown below. Frequencies up to several MHz can be handled, but ringing may occur beyond 100kHz if the circuit layout is poor.



Further reading

Electronic Circuit Design Handbook, Tab Books, 4th edition, p.129.

Designing with t.t.l. Integrated Circuits, McGraw-Hill, 1971, pp.53-7.

MECL Integrated Circuit Schmitt Triggers, Motorola application note AN-239, 1967.

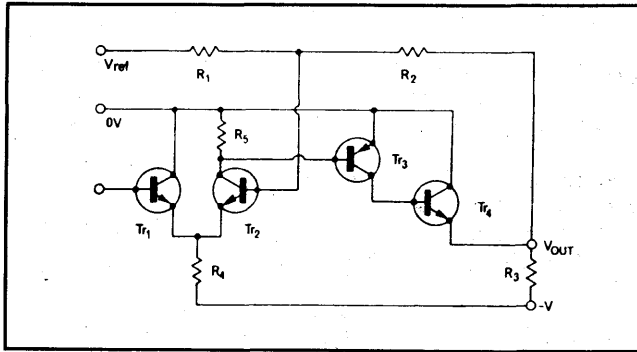
Cross references

Series 2, cards 2, 3 & 7.

Wireless World Circard

Series 2: Comparators & Schmitts 9

Low-voltage level sensor



Typical performance

Supply: -12V	R ₁ : 3.3kΩ
V _{ref} : -1V	R ₂ : 100kΩ
Tr ₁ , Tr ₂ , Tr ₄ : BC125	R ₃ : 470Ω
Tr ₃ : BC126	R ₄ : 82kΩ
Switching levels	R ₅ : 10kΩ
on: -1.35V	
off: -1.03V	

Circuit description

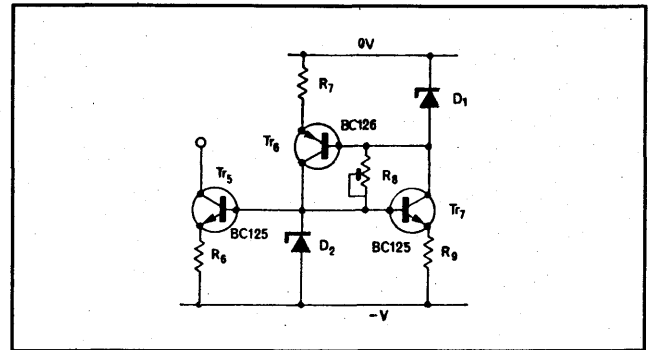
Operation from single-ended supplies makes level-sensing of low voltages difficult (lower limit usually set by transistor V_{be}). Taking signal and reference voltages with respect to opposite side of supply as shown allows much reduced triggering voltages. A long-tailed pair drives an inverting stage with positive feedback from the output to the non-inverting input. Input current is small, reducing to zero after switching. For positive-going signals, a complementary version using a positive supply voltage gives comparable results.

Component changes

- Supply voltage -5 to -25V, upper value depending on transistor breakdown. At lower voltages, switching levels become more supply sensitive. Reduce R_4 at lower supply voltages to keep current in it to $\sim 120\mu\text{A}$.
- Reference voltage -200mV to -5V.
- Load currents up to 100mA possible with no change in

circuit. Replacing Tr_4 by higher rating transistor, and scaling all resistors down by factor of 5 allows load currents of up to 0.5A (BFR41, BFR50 etc).

- Tr_1 — Tr_4 replaced by any general-purpose silicon planar transistors results in comparable performance: matched pair at input reduces drift.



Circuit modifications

- Reference and signal inputs may be interchanged if minimum current drain from reference is required.
- Replacing R_4 by constant-current circuit minimizes shift of switching levels with varying supply voltage. Fig. shows a ring-of-two reference circuit biasing constant-current stage, and providing stable voltage across R_7 to act as switching-level reference. Replaced by potentiometer for variable reference.
- Tapping R_2 with a zener diode to 0-V line stabilizes hysteresis without limiting output voltage swing.
- For light loading Tr_4 may be omitted.

Further reading

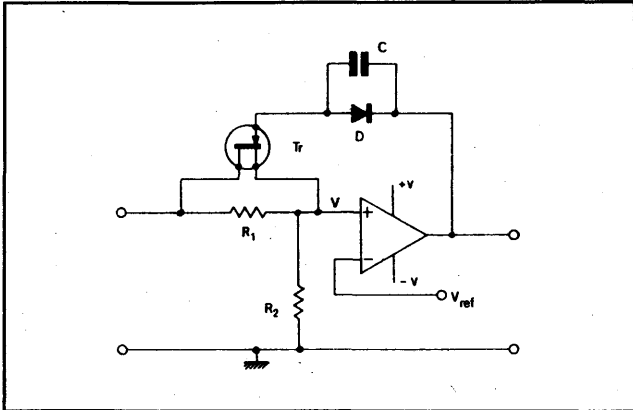
Williams, P., Low-voltage level-sensing circuit, *Electronic Engineering*, 1968, pp.517-9.
Callahan, M. J., Integrated level detector, *IEEE Journal of Solid-State Circuits*, vol. SC-7, 1972, pp.185-8.

Cross reference

Series 2, cards 6, 10 & 11.

Wireless World Circard Series 2: Comparators & Schmitts 10

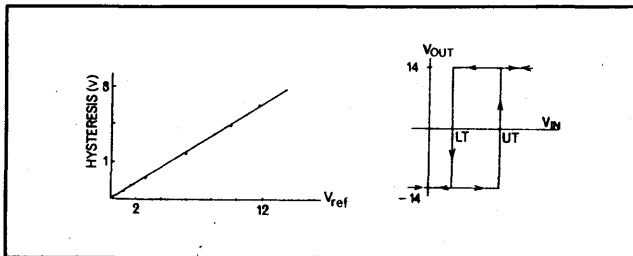
Reference-controlled hysteresis circuit



Typical performance

Graphs obtained with
Supplies: $\pm 15V$
Tr: Motorola 2N4092
D: 1N914, IC: 741
 R_1 : $5.6k\Omega \pm 5\%$
 R_2 : $27k\Omega \pm 5\%$
C: 100pF

Lower threshold (l.t.):
 V_{ref}
Upper (u.t.): $V_{ref} \cdot (R_1 + R_2)/R_2$
Hysteresis: $V_{ref} R_1/R_2$
Max. frequency: 300Hz
 V_{ref} must remain positive.



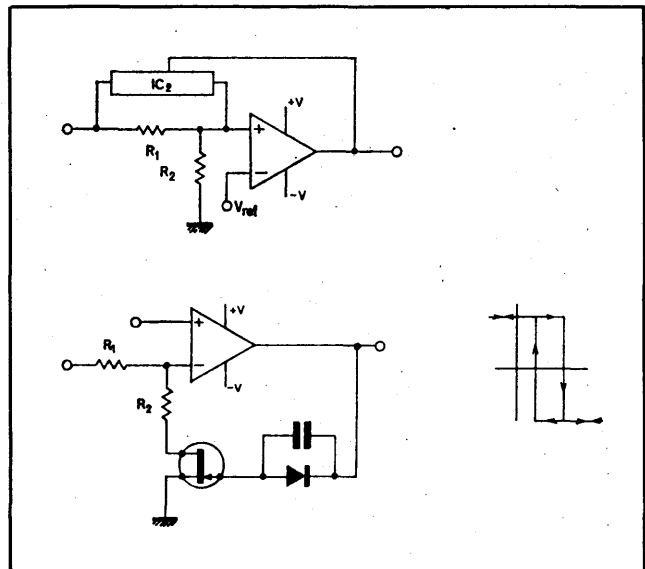
Circuit description

With a low V_{in} , V_{out} is initially negative and the f.e.t. switch is off. V' is then given by $V_{in} \cdot R_2 / (R_1 + R_2)$. Increasing V_{in} until V' is just greater than V_{ref} causes V_{out} to change sign, the f.e.t. then conducts and shorts out R_1 making V' equal to V_{in} and forcing V_{out} to become even more positive. V_{out} will only become negative again when V_{in} is reduced below V_{ref} . The positive feedback does not come into action immediately V_{out} starts to leave its saturated condition, so the output may lie between the saturated levels.

Component changes

Using a 748 op-amp the maximum frequency can be extended to 4kHz. National Semiconductor f.e.t. 2N3819/7127 may

be used. R_1 is chosen such that the f.e.t. on-resistance is much lower than R_1 and the off-resistance is much higher than R_1 . Varying R_1 and R_2 hysteresis of $0.1V_{ref}$ and $10V_{ref}$ can easily be obtained. Choice of diode and capacitor is not critical.



Circuit modifications

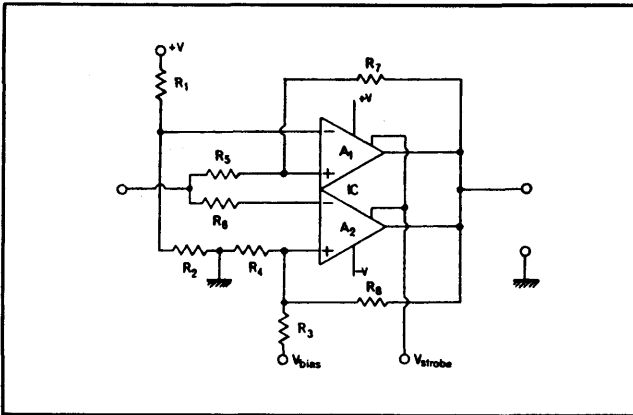
- With a negative V_{ref} , V' should be connected to the inverting input and V_{ref} to the non-inverting input to obtain the positive feedback switching action.
- With a low reference voltage and low supply voltage (e.g. $<1V$ with $\pm 5V$ supply), f.e.t. pinch-off voltage causes unsatisfactory switching. The f.e.t. and its associated diode and capacitor may be replaced by a c.m.o.s. switch, top. The switch used was CD4016AE, the minimum R_1 in this case being about $10k\Omega$.
- For applications where V_0 is required to be positive, for positive V_{ref} and small V_{in} , one may use the circuit at bottom, the resulting characteristic being as shown on right. The formulae for the upper and lower thresholds and the hysteresis are the same as those for the original circuit.

Further reading

G. S. Oshiro, *Electronic Design*, June 1972.

Cross reference

Series 2, cards 4 & 6.

'Window' detector**Typical performance**

IC: 711

Supplies +12V, -6V

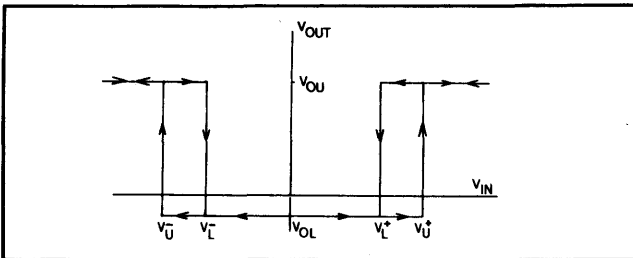
 $V_{\text{strobe}} = +5\text{V}; V_{\text{bias}} = -12\text{V}$ $R_1, R_3: 5.6\text{k}\Omega$ $R_2, R_4, R_5, R_6: 470\Omega$ $R_7, R_8: 22\text{k}\Omega$ $V_U = 953\text{mV}; V_L =$ 886mV (i.e. V_{OU} hys-
teresis = 67mV) $V_U = -946\text{mV}; V_L =$
-879mV (i.e. V_{OL} hys-
teresis = 67mV) $V_{\text{OU}} = +4.4\text{V}; V_{\text{OL}} =$
-0.4V (inputs
commoned)

Supply current:

+11.5mA, -5mA

Strobe current: 76 μ A

Bias current: 2.2mA

**Circuit description**A₁ is a non-inverting comparator having a positive reference

level (V_U) set by R_1 and R_2 . Amplifier A_2 is an inverting comparator having a negative reference (V_L) set by R_3 and R_4 . V_{out} remains at a low level when $V_{r_2} < V_{\text{in}} < V_{r_1}$, and A_2 is capable of switching its output to $+V_{\text{OU}}$ when $V_{\text{in}} < V_{r_2}$. As the outputs of A_1 and A_2 are common, $V_{\text{out}} = +V_{\text{OU}}$ when $V_{\text{in}} < V_{r_2}$, or when $V_{\text{in}} > V_{r_1}$. Hysteresis is introduced by the positive feedback on A_1 (by R_7 and R_5) and A_2 (by R_8 and R_4). See transfer characteristic above. From an output level viewpoint the circuit is t.t.l. compatible.

Component changesMaximum useful frequency $\approx 1\text{MHz}$.

V_{OU} may be varied over the range +0.4 to 4.4V by setting V_{strobe} in the range +1 to +5V. V_{OL} remains fixed at -0.4V. Variation of R_1 and/or R_3 provides independent control of the positive and negative threshold levels. Minimum useful value of R_1 and/or $R_3 \approx 700\Omega$.

Minimum load resistance (for 10% reduction of V_{OU}) $\approx 680\Omega$.**Circuit modifications**

- Where hysteresis is not required the positive feedback resistors may be omitted.

- A visible-light-emitting diode connected to output terminal through a limiting resistor gives visual indication when the input signal is outside the limits set by V_U and V_L . Typically a resistance of 470 Ω provides 5mA which is sufficient to illuminate the l.e.d. without excessively loading the IC.

Further reading

Application of linear microcircuits, SGS, 1969. pp.105-6.

Op-amp circuit collection, National Semiconductor application note AN-31, 1970, p.3.

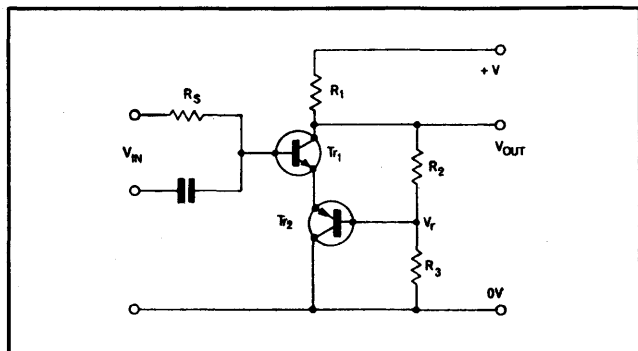
Measurement of burst ("popcorn") noise in linear integrated circuits, RCA application note ICAN-6732, 1971. p.6.

Cole, H. A., Differential discriminator circuits, *Wireless World*, 1971, pp.603-4.**Cross references**

Series 2, card 6.

Wireless World Circard Series 2: Comparators & Schmitts 12

Complementary Schmitt



Typical performance

$V_{in(on)} = V_r + 2V_{be}$	$V_{in(off)} = 1.59V$
Supply: +12V	hysteresis = 0.57V
Tr ₁ : BC125; Tr ₂ : BC126	$V_O(on) = 1.0V$
R ₁ , R ₃ : 1k Ω ; R ₂ : 10k Ω	$V_O(off) = 11V$
R _s : 100 Ω ; C ₁ : 0.1 μF	Supply current:
$V_{in(on)} = 2.16V$	10.5mA (on), 1mA (off)

Circuit description

This is a complementary form of circuit using emitter coupling, as in the classic Schmitt. Neither transistor conducts for low input voltages. When V_{in} exceeds $V_r + 2V_{be}$, both transistors conduct causing the output voltage to fall regeneratively. With suitable resistance values, the supply current in the off state can be made much less than in the on state.

Component changes

● Varying R_2 in the range 5 to 20k Ω allows the hysteresis to be adjusted within the range 1.16 to 0.21V, without significantly changing $V_{in(off)}$.

$V_{in(on)}$ correspondingly varies in the range 2.81 to 1.74V.

● If Tr₂ is a high-current-gain transistor $V_{in(off)} \approx 2V_{be}$; with a lower gain transistor $V_{in(off)}$ will be increased due to the significant p.d. across R₃ produced by Tr₂ base current.

● A speed-up capacitor of about 27pF across R₂ improves the turn-on time from about 90ns to 30ns. Turn-off time is typically 90ns. Maximum useful frequency is typically 2MHz.

Circuit modifications

A 47- Ω resistor included in series with the 'free collector' of Tr₂ provides a complementary pulse output. These pulses typically have an amplitude of 0.6V (with a 12-V supply) i.e. sufficient to drive a following transistor or thyristor. Tr₁ will still saturate and Tr₂ will remain unsaturated. The value of this resistor may be considerably increased if it is returned to a separate negative supply. With a value $\leq R_1$ a second output is then available without significantly changing the circuit action.

Further reading

Hemingway, T. K., *Electronic Designer's Handbook*, Business Books, 1970, 2nd edition, pp.177-181.

Feinberg, R., *Handbook of Electronic Circuits*, Chapman and Hall, 1966, p.52.

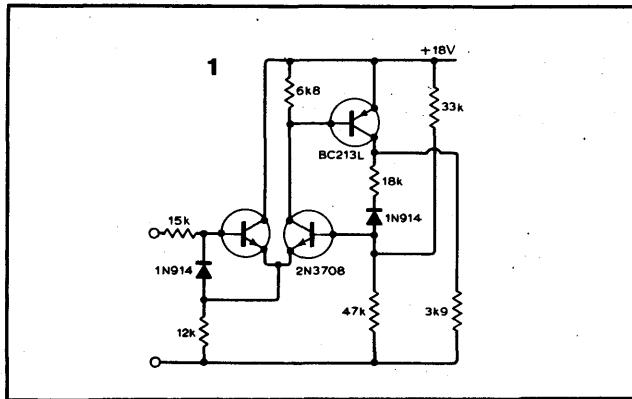
Cross references

Series 2, cards 2 & 5.

Schmitts and comparators

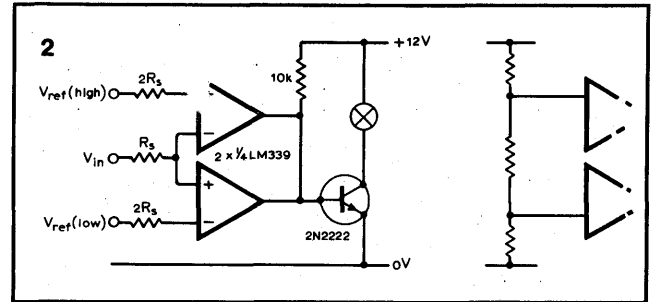
1. The circuit shown is a modification of the standard Schmitt circuit in which the input transistor is replaced by long-tailed pair allowing for better temperature compensation. Signal outputs are available from either collector, and the use of common-emitter output stage is helpful in getting a large output swing. Positive feedback is heavy, because the system of which the circuit is part can suffer from spurious triggering due to contact-bounce.

Orr, T. and Thomas, D. W. Electronic sound/synthesizer: part 3, *Wireless World*, vol. 79, 1973, pp.485-90.



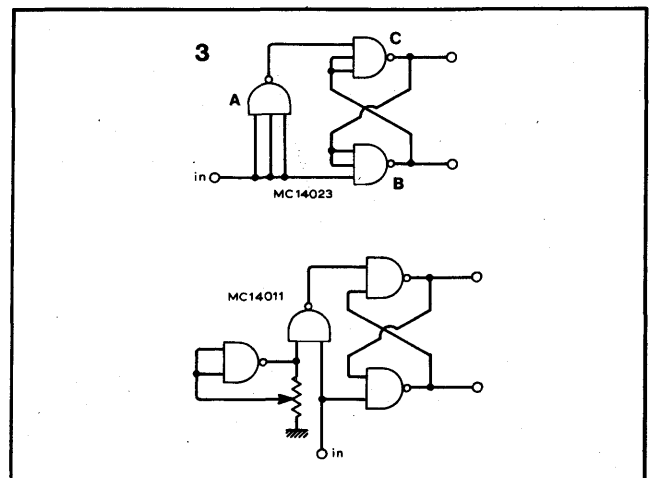
2. Recently-developed comparators such as the LM339, MC3402 can be operated from a single polarity supply voltage, with the zero supply line included within their common mode range. Minimum supply voltage can be as low as 2V, the output is an open collector device and the whole circuit can be t.t.l./d.t.l./c.m.o.s. compatible. In addition there are four such comparators in a single package allowing complete subsystems to be constructed with the minimum of interconnections. The example shown has the output devices of both comparators off when V_{in} lies between the reference voltages i.e. the transistor receives base current, illuminating the lamp. For inputs $> V_{REF} (high)$ or $< V_{REF} (low)$ the lamp is off. The reference voltages can be provided by a simple potential divider chain across the supply.

Smathers, R. T., Frederiksen, T. M. and Howard, W. M. LMI39/LM239/LM339 A quad of independently functioning comparators, National Semiconductor application note AN74, 1973, p.5.



3. The simplest Schmitt trigger circuit using c.m.o.s. has positive feedback across a single non-inverting stage or a cascaded pair of inverters. In these circuits the thresholds are defined by paralleling inputs of multi input gates e.g. for two inputs high the threshold of the third is $\approx 5V$ for a 15V supply. For three inputs in parallel the threshold is 8.5V. For output C high, the threshold for A is 8.5V and for B is 5V. The absence of resistive feedback keeps the input impedance extremely high, while gates C and B form a set - reset flip-flop with rapid switching once the thresholds are reached. The second circuit adds a variable bias which allows the hysteresis to be varied from zero to 50%.

Halligan, J. Schmitt trigger with c.m.o.s. gates, *Semiconductors* (Motorola) 1974/1, pp.29-30.



Wireless World Circard

Series 3: Waveform generators

The article on the next two pages shows the waveforms considered in this set of cards; sinewave oscillators and modulated trains of pulses were excluded. It provides an excellent "potted" account of the various methods of waveform generation.

Five cards show triangular-wave generators. As four of them are based on the integrator plus hysteresis switch approach, they also double as rectangular-wave generators. (The fifth uses an emitter-coupled circuit switching a constant current alternately through both directions of a capacitor.)

Card 12, whilst not properly describing a waveform generator, gives some waveshaping circuits for converting triangular waves into approximately sinusoidal ones. One technique described uses an integrator with d.c. negative feedback but the approximation is crude, typically giving 4% harmonic distortion. Better methods (<1% distortion) use non-linear elements in a feedback loop and two circuits are given with diodes and an f.e.t. A fourth waveshaper on this card gives a sawtooth wave from a square and triangle generator.

Examples of the two basic techniques for staircase generation are included—an interesting improved "pump" circuit and the ladder kind of digital-to-analogue converter.

- Basic op-amp square/triangle generator 1
- Emitter-coupled triangular wave generator 2
- Diode-pump staircase generator 3
- Unijunction sawtooth generator 4
- Voltage-controlled square/triangle generator 5
- Complementary transistor sawtooth generator 6
- Digital-to-analogue converter waveform generator 7
- Triggered ramp/trapezium generator 8
- Stable waveform generator using single i.c. 9
- Simple multi-waveform generator 10
- Op-amp/c.m.o.s. square/triangle generator 11
- Simple wave-shaping circuits 12

Waveform Generators

As electronics is largely concerned with the generation and processing of electrical signals, the subject of this article could include half of all known circuits. It will be easiest first to exclude certain well-defined classes of generator to be dealt with in later series. For example sine waves are perhaps the most commonly generated waveforms, and the great variety of such circuits warrants separate treatment. Similarly, pulse waveforms have ever-widening applications in communication and digital systems and correspondingly numerous circuits have been published. There remain a number of well-defined waveforms of general usefulness, sufficiently important to qualify for a series to themselves. They include triangle, ramp (linear and exponential), staircase and trapezium waveforms, together with others that are conveniently generated from them. For example, an excellent means of generating a triangular wave requires the repetitive reversal of bias to an integrator by a Schmitt trigger, the output of which is thus a square wave. Where it is possible to modify an existing waveform simply, to provide an approximation to some other desired function, the method is indicated below. It is not possible within the confines of a short article or single set of Circards to do more than outline such methods.

The majority of these generators depend on the charging of a capacitor, though dual circuits using inductors may be devised. To provide a repetitive waveform, the direction of charge flow has to be periodically reversed, and this leads to a major subdivision. The reversal may be such that the charge-rate is comparable for the two directions, in which case the capacitor waveform belongs to the triangular class. If the charge rates differ markedly the waveform approximates to a saw-tooth. In either case the charge rate may be constant or may vary during a half-cycle.

If charge is passed into the capacitor in discrete quantities at regular intervals, then a third type of waveform, the staircase waveform, Fig. 5, is produced. Digital waveform generators using digital-to-analogue (or d/a) converters can also produce stepped waveforms, the staircase being a particular version. If the steps are small enough and sufficient in number then waveforms can be synthesized to any required accuracy and the cost of such generators continues to fall with the increasing range of digital circuits of low cost. To remove the steps from the output is not possible but the use of suitable filters, such as those of series 1 Circards, can reduce the ripple at the step frequency without seriously distorting the synthesized waveform. A particular advantage of the d/a converter method is that no reactive elements are

employed as the output frequency is defined exactly by the clock frequency and the division ratio of the counter used.

Ramp and sawtooth generators

In ramp generators a capacitor receives a defined current, constant if the ramp is to be linear. If the current flows for a period determined by some external agency the ramp is said to be triggered, and normally at the end of the ramp period the capacitor is discharged and the source of current removed (or the current by-passed). Such an action is required for the triggered timebase of oscilloscopes. A second mode of operation is where the capacitor discharge is immediately followed by the re-establishment of current flow and the restarting of the cycle is a free-running mode. The parameters of the waveform of interest are linearity of the ramp, accuracy of definition of the end-points and duration of the interval between the end of one forward stroke and the next (flyback). Most applications require a linear ramp, Fig. 2, though the natural tendency is for current and hence rate of charging to fall as the capacitor p.d. increases, Fig. 3. Some circuits have bootstrapping of the voltage drive, so that the p.d. across the current-defining resistor remains constant. Compensation for leakage-current effects to maintain linearity or over-compensation to obtain waveforms as in Fig. 4 are other possibilities if the bootstrap circuit has a gain greater than unity.

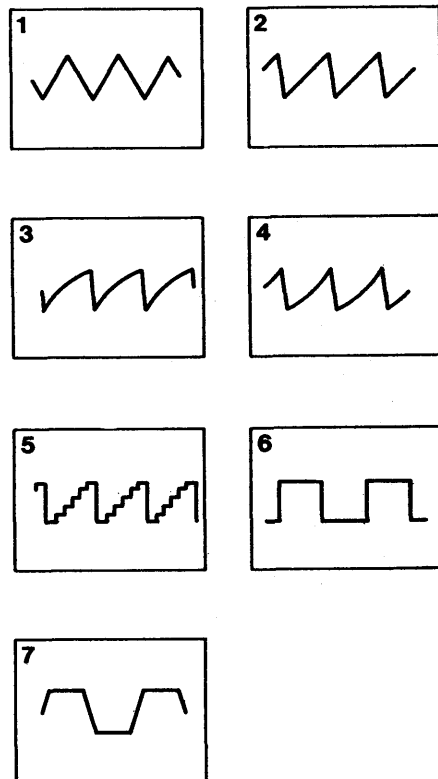
Flyback duration is determined by the rapidity with which the charge may be withdrawn from the capacitor, i.e. by the current capability of the discharge device/circuit. Perhaps the simplest device having a well-defined firing point, reasonably low leakage and satisfactory discharging capability is the unijunction transistor. With the so-called programmable unijunction device, which is internally more akin to a thyristor, a wide variety of circuits has been devised with linearity and definition of end points of around 1%. Constant-current sources are used to achieve linearity. Recent developments have included integrated circuits consisting of comparators, bi-stables and switching transistors, which though designed as timing circuits can produce similar waveforms. All such circuits suffer from the disadvantage that the load current flows in the charging circuit, disturbing the linearity.

If the capacitor is placed in the feedback path of an amplifier, as in the integrator due to Blumlein, though now more often called the Miller integrator, then a constant current may flow in the capacitor without load current disturbing that flow. This method is widely used in the circuits of the following section.

Triangular wave generator

The slopes of a triangular wave generally have equal positive and negative values, giving a symmetrical waveform. If desired, inequality may be introduced which if great enough yields an asymmetry that results in a near-sawtooth waveform. Combining an integrator of this type, often using an i.c. operational amplifier, with a Schmitt trigger (see series 2 Circards) both triangular and square waves are produced: a positive output from the Schmitt trigger drives the integrator to produce a negative-going ramp that in turn reverses the trigger output at some defined potential. Asymmetry is introduced by varying the effective input resistance of the integrator on the output swing of the trigger circuit for the two sections of the cycle. Control of the triangular wave amplitude is by means of the trigger-circuit input switching levels.

There are many applications which require voltage control of the frequency, e.g. remote programming, frequency modulation, amplitude-frequency response testing. To achieve this, the output of the



These waveshapes illustrate those dealt with in series 3 of Circards, being triangular (1), sawtooth (2), 'exponential' (3), over-compensated (4), staircase (5), rectangular (6) and trapezoidal (7) waveforms.

square-wave circuit is used to reverse the charging current, rather than feeding the integrator directly, leaving the magnitude of the current controlled by some external voltage or current source. Several methods are possible. An inverting amplifier provides a second voltage of equal magnitude but opposite sign to the control voltage, and the square wave activates electronic switches such as f.e.t.s to select these opposite-polarity voltages alternately. Amplifiers may be designed in which the gain can be switched from positive to negative leaving the magnitude of the output again proportional to the controlled source. Finally, the integrator itself may be designed so that the switching is achieved directly, the current flow being reversed within the integrator.

In all these approaches the accuracy with which the frequency tracks the applied voltage depends on the switch. It should have that legendary performance of zero on-resistance and infinite off-resistance. Field-effect transistors, both junction and insulated gate, come close to achieving the latter parameter, but the low-cost units have on-resistances high enough to introduce errors, though first-order compensation by a deliberate unbalancing of the symmetry control could be used. In other cases bipolar transistors or even diodes are

applicable, while for the highest accuracy more complex switches involving pairs of complementary f.e.t.s provide a solution.

Square/trapezium generators

A square-wave output is an integral part of many triangular-wave generators. Rise-time is defined by the particular op-amp or comparator used, while the amplitude may approach supply voltage levels (though saturation effects are significant at lower supply voltages). By applying the square wave to a second integrator with sufficient over-drive, a trapezium wave, Fig. 7, with defined amplitude and variable-slope rising and falling edges results. The slopes may again be independently controlled if the drive conditions differ for the positive and negative portions of the input.

Staircase generators

The diode pump is the classic circuit for obtaining a stepped output waveform that approximates to a linear ramp when the steps are of equal height and provided there is no change in amplitude during the interval between the steps. This implies that the capacitor should not discharge appreciably between input pulses and puts a lower limit on the repetition rate for any given circuit. To maintain equal step size at all outputs,

amplifiers may be introduced to provide functions similar to bootstrapping as in the ramp circuits. This brings with it the upper frequency limit of the amplifier.

Digital-to-analogue converter methods extend the amplitude response down to d.c., with high accuracy also available, and an indefinite variety of wave-shapes that can be synthesized by selection of suitable resistor values.

Wave-shaping circuits

In waveform-shaping circuits the options are very wide and the topic requires separate treatment, but some simple methods of shaping a triangular wave into an approximate sine wave can be suggested. In a saw amplifier the gain can be reduced as the magnitude of the input increases in two distinct ways, each of which gives a rounded peak to the output when fed by a triangular wave. The first method places a f.e.t. in the input with its increasing slope resistance at high amplitudes, while the second uses p-n diodes across the feedback network.

A second type of shaping involves the use of a switch to reverse a given waveform at some point in the cycle. This technique can be used to convert triangular waves into saw-tooth waves for example.

Wireless World

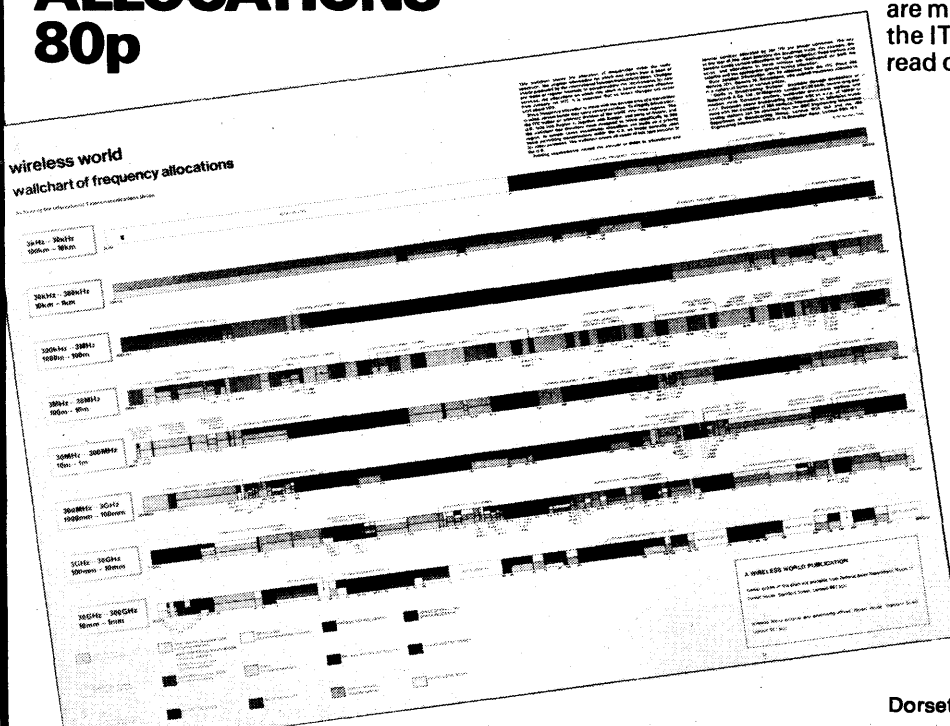
FULL COLOUR WALLCHART

OF FREQUENCY

ALLOCATIONS

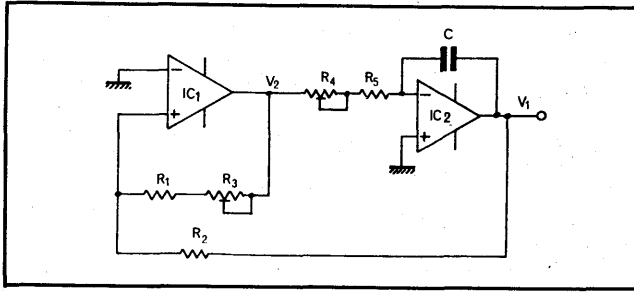
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The wallchart shows the allocation of frequencies within the radio spectrum ranging from 3 kHz to 300 GHz and is scaled on eight logarithmic bands contriving 15 main categories of transmissions which are identified by colours. All the important spot frequencies and 'special interest' frequencies are marked. The information is taken from the ITU and has been condensed into easily read chart form. Measures 2' 11" x 1' 11".



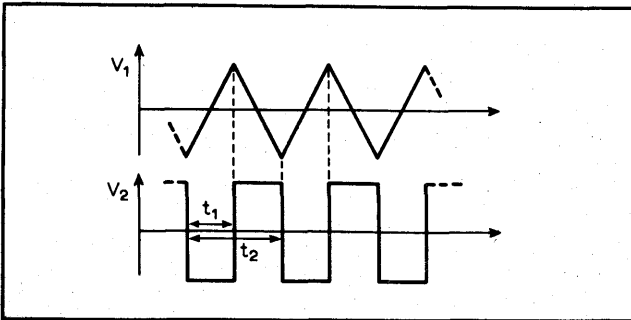
IPC Electrical-Electronic Press Ltd.
Dorset House, Stamford Street, London SE1 9LU

Basic op-amp square/triangle generator



Data

IC ₁ , IC ₂ : 741	Frequency: zero to 3kHz
Supplies: ± 6 to ± 15 V	R ₁ : 10k Ω ; R ₂ : 8.2k Ω
V ₂ : ± 0.9 V	R ₃ : 1M Ω ; R ₄ : 220k Ω
V ₁ : 1 to 20V pk-pk	R ₅ : 1.2k Ω ; C: 0.1 μ F



Circuit description

When a positive voltage is applied to the input of the inverting integrator, consisting of R_4 , R_5 , C and IC_2 , the current flow causes C to charge, with its input end positive w.r.t. its output end. Negative feedback through C and the high gain of the amplifier jointly ensure that the inverting terminal retains a potential very close to that of the non-inverting terminal. The output must therefore go negative and provided the amplified input current is much less than the constant current in R_4 and R_5 the output voltage rises linearly with time. At some value of V_1 the negative current fed back through R_2 will overcome the positive current in R_1 and R_3 , and the resulting negative current in the non-inverting input of IC_1 initiates a negative going transition in V_2 . This allows the negative current in the non-inverting input to further enhance the output swing by this positive feedback action. The integrator output then reserves its slope and eventually becomes positive and finally switches V_2 back to its original positive value. Hence V_2 is a square wave and V_1 a triangular wave. Resistor R_4 gives independent frequency control and R_3 varies the frequency and the magnitude of V_1 .

Component changes

- The low frequency of operation of this circuit is due mainly to the limited slew-rate of a 741 op-amp as the active element IC_1 . A 301 op-amp will permit frequencies of up to 10kHz to be achieved, the square wave degenerating visibly before the triangle.
- R_1 and R_5 limit the current drawn from IC_1 when R_4 and R_3 are in their minimum positions and could possibly be omitted.
- R_2 may be varied widely but must not be so low that IC_2 is heavily loaded and not so high that IC_1 fails to switch before IC_2 reaches saturation.
- C can also be changed, bearing in mind that the slope of the triangle is inversely proportional to $C(R_4 + R_5)$.

Circuit modifications

- The triangular wave can be given a d.c. offset of either polarity by applying a bias signal V_6 as shown left in which $R_6 = 10k\Omega$. The bias can be increased to the point at which the integrator saturates without changing the state of IC_1 .
- A sawtooth waveform can be achieved by adding a d.c. signal V_3 to the integrator output, as shown middle. The magnitude of V_3/R_7 must be less than $[V_2/(R_5 + R_4)]$ otherwise the integrator output will not change direction as V_2 changes sign. Time t_1 is greater than $t_2/2$ if V_3 is positive and t_1 is less than $t_2/2$ if V_3 is negative. The ratio V_3/R_7 must be comparable to $V_2/(R_5 + R_4)$ if a large mark-space ratio is required. Independent frequency control through R_4 is lost when this is done but may be regained by varying C .
- A sawtooth waveform can also be produced by the circuit, right, which does not require an external signal. Any general purpose diode will do. With the diode as shown $t_1 < t_2$, but $t_1 > t_2$ if the diode is reversed. As shown, the integration rate on the negative-going side of the triangle is controlled with $R_4 + R_5$ and on the positive-going side by R_8 in parallel with $R_4 + R_5$.

- The output of IC_1 may be clamped to a well-defined level by inserting a series resistor in the output lead and taking a pair of back-to-back zener diodes to ground. This produces a better defined integration rate and makes t_1 more nearly equal to $t_2/2$. Drive point for the circuits is taken as the junction of the resistor and zeners. Clamping on many i.c.s is possible at low signal levels by means of terminals on the i.c.s (cross ref. 2).

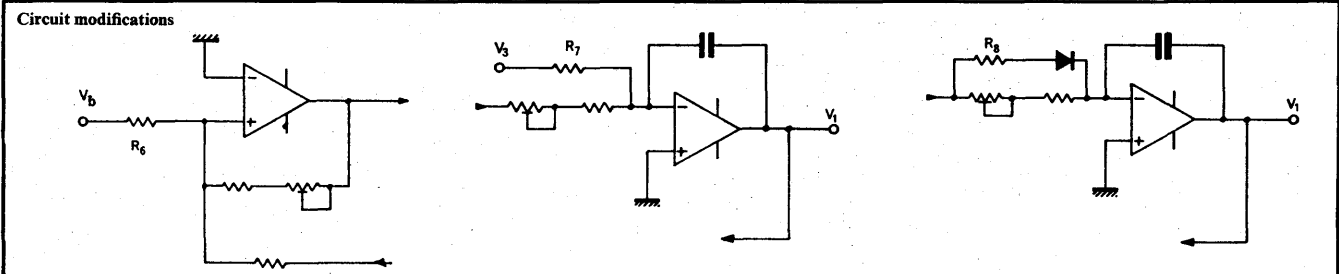
Further reading

Clayton, G. B. Triangular square wave generator *Wireless World*, vol. 75, 1969, pp.586/7.
 Tobey, G. E., Graeme, J. G. & Huelsman, L. P., *Operational Amplifiers*, McGraw Hill 1971, pp.373-81.
 Linear Applications Handbook, National Semiconductor application note AN31-6, 1972.

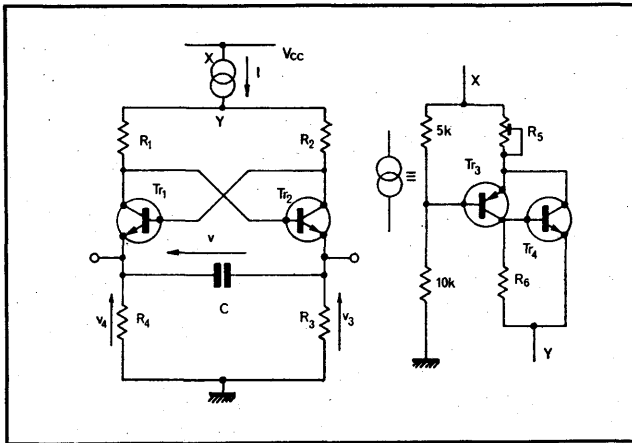
Cross references

- 1 Series 3 cards 2 & 11.
- 2 Series 2 cards 1 & 3.

Circuit modifications



Emitter-coupled triangular wave generator



Typical performance

Tr₁, Tr₂: BC125
 Tr₃: BC126
 V_{CC}: +15V
 C: 0.1μF
 R₅: 2.2kΩ
 R₁, R₂: 330Ω
 R₃, R₄: 4.7kΩ
 R₆: 10kΩ
 f: 5.3kHz
 f ∝ 1/C

v₃ = v₄: Antiphase triangular waveforms 0.4V pk-pk on a d.c. level of 4V. As R₅ is reduced, d.c. level rises towards 8V; frequency increases to 7kHz, as long as triangular waveform is maintained. A ramp voltage at 2f is available at Y.

Circuit description

The circuit is an emitter-coupled astable circuit normally fed from a voltage source. This results in sharp transitions in the voltages across R₃ and R₄ at the circuit switching points. These can be eliminated by driving from a constant-current source, so that only the direction of charging current in capacitor C is reversed, the magnitude varying little throughout the cycle. Consider Tr₁ fully conducting, Tr₂ off. The charging circuit is then as shown in the above diagram. Provided the conditions v < v₃ and v ≪ v₄ are maintained, the capacitor charges linearly, but in any case for R₃ = R₄, any rise in v₄ must be accompanied by an identical fall in v₃, to maintain a constant total current. Hence there are two outputs v₃ and v₄ which are of identical shape but anti-phase, and are

also good approximations to triangular waveforms. The transition will occur in the above example when v ≈ -0.5V at which condition Tr₂ begins to conduct, positive feedback rapidly completing the transition. Increasing the source current increases the charging rate, and hence frequency, with little change in amplitude. The output has an amplitude of ~1V pk-pk at a mean level of ~5V, depending on the controlled supply current. Supply current is defined by the constant-current source connected between X and Y, where R₅ determines the value of I.

Component changes

Useful range of C: 1μF to 3.3nF

Frequency range: 0.5 to 130kHz

A 20% reduction in V_{CC} varies frequency by about 5%.

Circuit modifications

- With R₅=0, C=0.1μF, a condition arises where the drive is not from a constant-current source, but the circuit is connected as an oscillator with d.c. supply ≈ 15V. The shape of the output waveform is then as shown left which has a d.c. level of 11V and a swing of 6V pk-pk.

- Voltage spikes occur at the positive and negative peaks of the normal triangular waveforms, due to the change in p.d. across the oscillator causing sharp current spikes from stray capacitance existing across the oscillator; i.e. circuit is temporarily operated from a constant voltage rather than a constant-current source. These can be eliminated by connecting capacitor C_x in the range 3 to 10C, as shown middle. Typically, for C = 0.1μF, R₅ = 2.2kΩ, C_x = 1μF. The waveform across R₄ is doubled, the frequency change being < 1%.

- Resistors R₃ and R₄ replaced by a 10-kΩ potentiometer R₇ as shown right. As R₇ is varied, the output triangular wave peak amplitude remains unchanged, though the slopes alter asymmetrically. Typically for C = 1μF, R₅ = 2.2kΩ, f = 0.5kHz when R₇ is set at mid-point. For a setting of 2:1, frequency reduces by 10%.

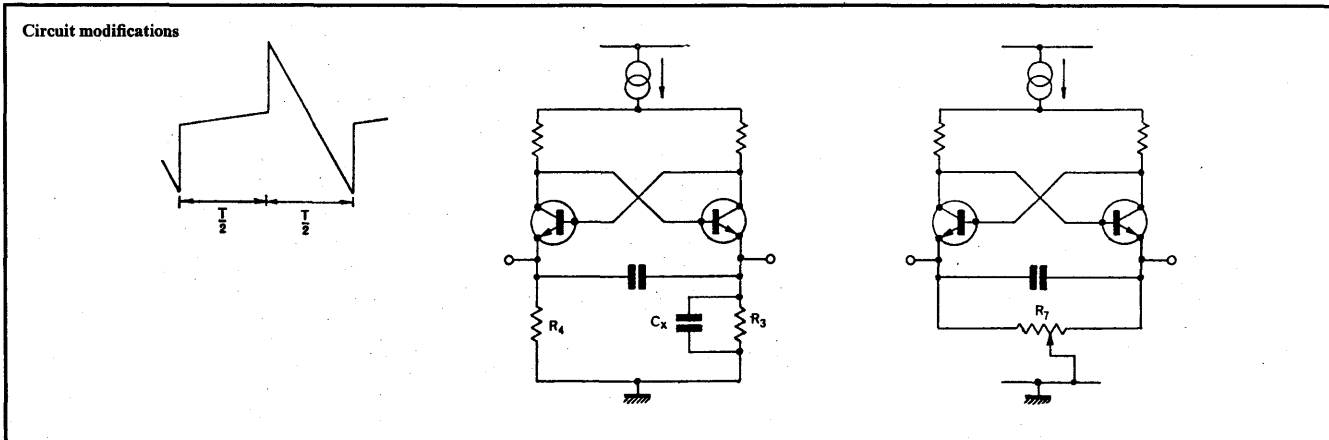
Further reading

Hemingway, T. K. *Electronic Designer's Handbook*, Business Publications 1966, pp.191-4.

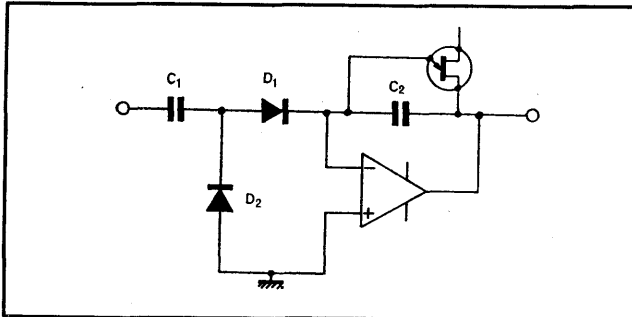
Transistorized all-waveform generator, in *Electronic Circuit Design Handbook*, Tab 1971, p.252.

Cross references

Series 3 card 1.



Diode-pump staircase generator



Typical performance

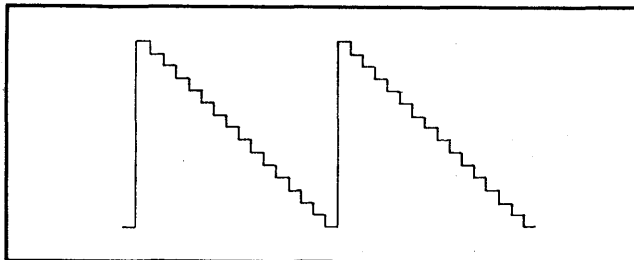
IC: 741
 Tr: 2N2160
 C₁: 3.3nF ± 10%
 C₂: 3.3nF ± 10%
 D₁, D₂: 1N914
 V_{in}: +3.6V pulses
 repetition rate: 1kHz
 pulse width: 200μs
 Output step ~ 0.28V
 Ramp height ~ 8V
 No. of steps ~ 28

Supplies: ±15V
 V': +4V

Step size $\approx \frac{(V_{in} - 1) C_1}{C_2} \cdot V'$

No. of steps = $\frac{\text{height}}{\text{step size}}$

Ramp height dependent on unijunction but ~ 2V'



Circuit description

The basic diode pump has diode D₂ feeding capacitor C₂ (grounded), and without the amplifier. On the first positive input pulse D₁ conducts and provided the pulse duration is long enough the pulse amplitude is shared between C₁ and C₂ - the same charge producing the larger portion of the p.d. across the smaller capacitance. On each succeeding pulse the p.d. established across C₂ opposes any fresh flow of charge, and the step in the output voltage diminishes progressively to zero when the p.d. across C₂ equals the input pulse amplitude. In the circuit shown, the amplifier virtual earth prevents the p.d. across C₂ from influencing the charge flow on successive cycles and the p.d. builds up in equal steps. In each case

the charge acquired by C₁ during the pulse is lost to ground through, D₁ when the input returns to zero i.e. C₁ commences each cycle in an uncharged state. Departures from the ideal are: p.d. across each diode when conducting is ~ 0.6V for silicon, reducing the effective input pulse amplitude by ~ 1.2V; amplifier input draws a small but finite current that adds a continually varying output due to integration via C₂; to make the circuit free-running a device such as a unijunction transistor must be added to provide periodic discharge of C₂, and such devices contribute additional leakage currents.

Component changes

C₁: 100pF to 1μF
 C₂: 100 pF to 1μF
 D₁, D₂: general-purpose Si diodes
 IC: any general-purpose compensated op-amp e.g. 307.
 V_{in}: 1 to 20V pk

Pulse rise time should not be too small a fraction of pulse width or excessive transient currents appear at amplifier input.
 Pulse width: < 1μs to > 1s
 Mark/space ratio: 1:100 to 100:1
 Repetition rate: 1Hz to 100kHz.

Circuit modifications

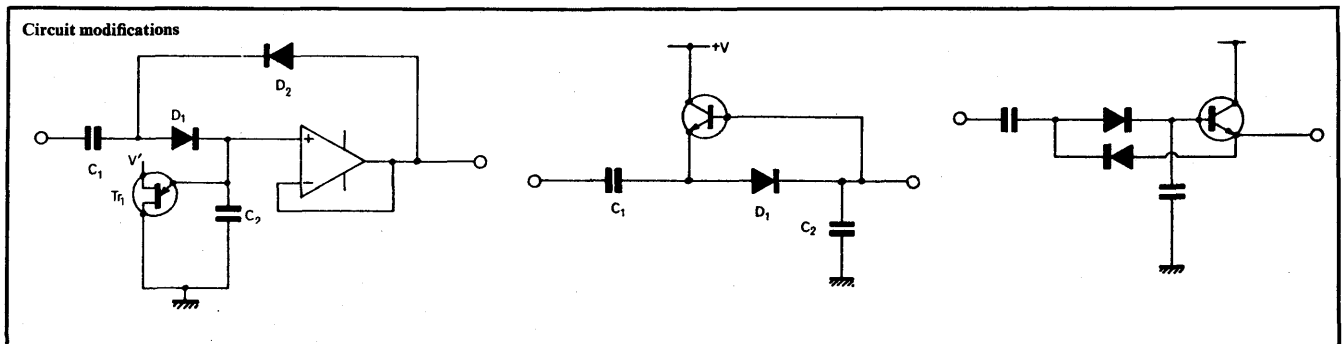
- Use of bootstrap technique returns r.h. end of C₁ to output through D₂ at end of each positive pulse. This ensures that on next positive pulse D₁ begins to conduct at start of pulse even if p.d. across C₂ and hence at output is greater than pulse height in. Ramp steps of constant size and ramp height limited only by amplifier. Again unijunction may be used to end ramp. (left)
- An alternative range of transistor-pump circuits may be devised. On the positive edge, D₁ conducts and C₁ and C₂ charge with p.d. shared between then in inverse ratio to capacitance. On negative edge, Tr₁ conducts clamping C₁ to just below output while discharging C₂ only by base current of Tr₁. (middle)
- An alternative form of bootstrap circuit comparable to first modification (right) with emitter follower replacing voltage follower.

Further reading

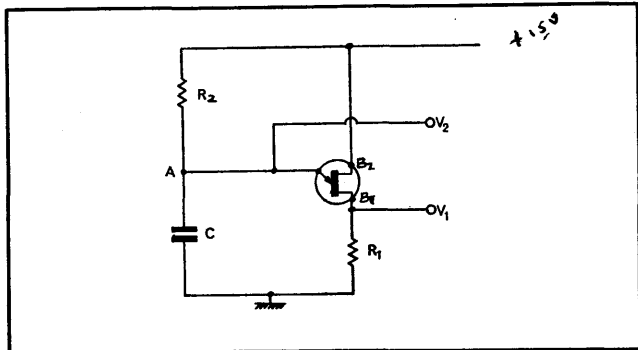
Hemingway, T. K., *Electronic Designers Handbook*, Business Publications 1966, pp.215-24.
 Staircase wave generator, in *GE Transistor Manual*, 7th edition, p.345.
 Clayton, G. B., Resistive feedback networks, *Wireless World*, Aug. 1972, pp.391-3.

Cross references

Series 3, cards 4 & 5.



Unijunction sawtooth generator



Data

Typical output waveforms obtained with
 R₁: 100Ω ± 5%
 R₂: 10kΩ ± 5%
 C: 1μF ± 5%

V: +15V
 Tr: 2N2160
 Supply range can be 4 to 20V at least.

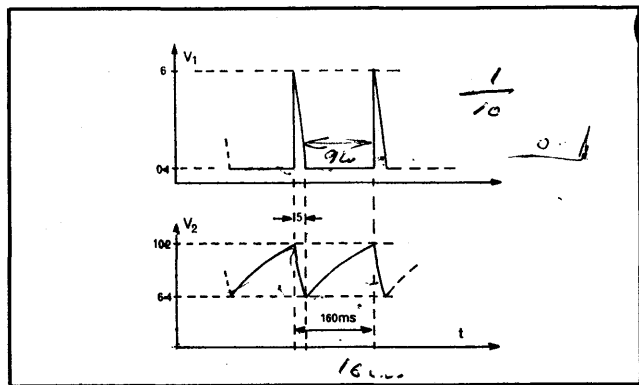
places an upper limit on the value of R₂ for guaranteed operation. The firing potential is temperature dependent because of the p-n junction p.d. at the emitter junction. This leads to temperature-induced frequency instability which can be compensated for by the insertion of a small series resistor in series with B₂. The rise in the B₁, B₂ path resistivity with temperature reduces the current and hence the p.d. across this resistor, leaving a larger part of the supply voltage at the junction.

Component changes

Reduction of R₁ to zero causes V₁ to become zero but has little effect on V₂. Any standard unijunction transistor may be used. Motorola 2N2646 will produce a smaller lower limit on V₂ and consequently reduced frequency for the same C & R.

Circuit modifications

● Discharge time through R₁ may be greatly reduced by the modification, shown left, in which V₂ is used to short the capacitor to ground. This makes the pulses of V₁ much narrower and alters the frequency slightly.



● The unijunction transistor may be replaced by the two transistor version, shown middle, with R₁ 100Ω, Tr₃ BC126, Tr₄ BC125 and the potentiometer 2.2kΩ. The lower value of V₂ in this case comes much closer to zero. The potentiometer is set to the maximum value of V₂ required plus V_{be}.

● Circuit shown right may be attached to any of the circuits to remove the error arising when the supply is switched on, at which point V₂ is at 0V rather than the minimum value it later achieves on the first discharge cycle. Resistors R₅ and R₆ are chosen so that the transistor conducts, charging C rapidly to this minimum value (ref 2).

● Charging resistor R₂ can be replaced by a defined current source e.g. a constant-current source will produce a linear ramp instead of an 'exponential', (Circards, series 3, card 2).

Circuit description

Circuit is used as a sawtooth (V₂) or a trigger pulse generator (V₁). Capacitor C charges through R₂ until the unijunction transistor V_p is reached and then discharges via R₁ until the transistor changes state at approximately 0.5V_p(sat); C then starts charging through R₂ again. Waveform frequency ∝ 1/R₂C. With R₂ fixed at 10kΩ and C varied, the waveform details (apart from the period) remain identical as C is reduced down to 220 nF. At 10nF, V₁ is reduced to half its previous value and the pulse width increases to approximately 1/10th of the period. At 1nF the pulse height is further reduced and V₂ becomes rounded.

Emitter leakage current modifies the charging waveform and

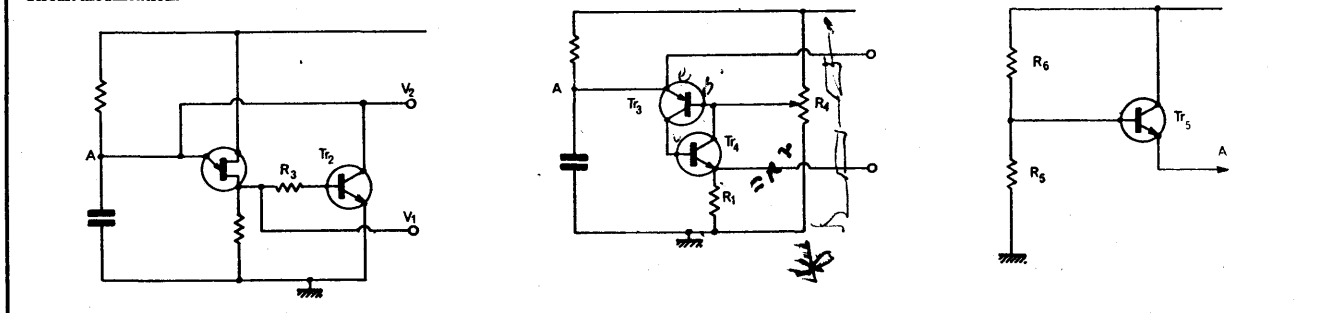
Further reading

- 1 Ultra-linear ramp generator uses UJT to drive Darlington, in 400 Ideas for Design, Hayden, p.119.
- 2 Electronic Circuit Design Handbook, Tab 1971, p.173.
- 3 Triac, UJT & FET give linear ramps, in 400 Ideas for Design, Hayden, p.128.

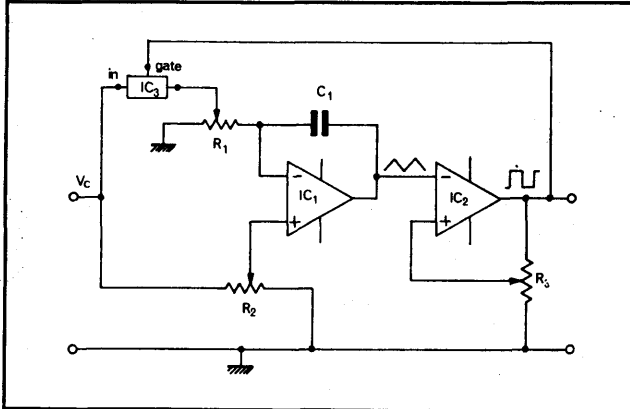
Cross references

Series 3, cards 2 & 8.

Circuit modifications

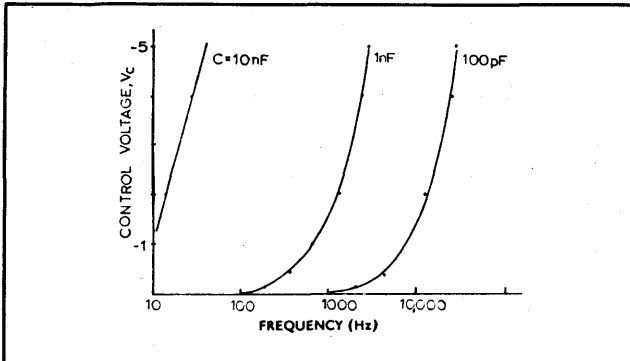


Voltage-controlled square/triangle generator



Typical performance

IC ₁ : 741	Triangular wave:
IC ₂ : 301	4V pk-pk
IC ₃ : ¼ (CD4016)	C ₁ : 1nF
Supplies: ±1V	R ₁ , R ₂ : 100kΩ (mid-position)
Control voltage, V _C : -1V	R ₃ : 10kΩ
Frequency: 660Hz	
Square wave: 8V pk-pk	



Circuit description

The basic idea of an integrator feeding a schmitt trigger may be adapted to allow voltage control of oscillator frequency. The square wave output of IC₂ controls an electronic switch IC₃ (in this case a c.m.o.s. transmission gate) which operates directly on the integrator without need for an additional reversible gain amplifier. A fixed portion of the control voltage is applied to IC₁ non-inverting input through R₂,

while the tap on R₁ is alternately open-circuited and connected to the input. With the switch closed, the inverting input receives a negative current as the full input is applied via part of R₁ to the inverting input, while the non-inverting input is held at some fraction of V_C. For an open switch the inverting input is returned to ground via R₁ while the inverting input is still maintained at a constant negative voltage. A convenient setting, if the switch is ideal, is for R₁ to be centre-tapped with the non-inverting input tapped onto R₂ at ½V_C. Either can be replaced by corresponding fixed resistors with the other varied to obtain best symmetry i.e. compensating for finite on resistance. The Schmitt circuit is conventional while the particular switch may be replaced by any switch that can connect the slider of R₁ to V_C.

Component changes

- Frequency is linearly related to control voltage V_C up to -4V.
- Useful range of C₁: 100 pF to 0.1μF.
- Positive feedback via R₃ must be <75% to maintain triangular shape, because of saturation of the IC₁ for the low supply voltage used.
- Adjustment of R₂ controls the mark/space of the square wave and slopes of the triangular wave, without altering the amplitude. Typically, C = 1nF V_C = -4V, mark/space can be 1:15 at f = 1250Hz to 17:1 at f = 460Hz.

Circuit modifications

In the circuit shown left, the basic form of the integrator and Schmitt circuit remains the same, but the electronic switch now operates in a shunt mode. A simple analysis to indicate appropriate potentiometer settings to ensure symmetrical triangles is shown above. Note that the control voltage is now positive with respect to ground. The linear relationship between V_C and f is indicated right for C = 1nF, n = 1, k = ½ and a supply of ±5V.

Triangular output is 4V pk-pk. Operation at 1Hz is easily achieved, but some readjustment of R₂ necessary to retain symmetry. Effect of supply voltage on frequency for the above components is also indicated.

Further reading

- LM311 Voltage Comparator. National Semiconductor data 1970.
- 2 Tidley, R. J., Voltage-controlled triangle/square generator, *Wireless World*, May 1972, p.239.

Cross references

Series 3, card 11.

Circuit modifications

For switch open $I = \frac{V_C - kV_C}{(n + 1)R}$

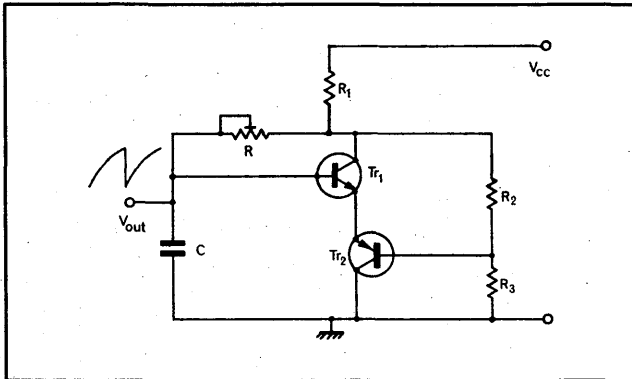
For switch closed $I = \frac{-kV_C}{R}$

For equal slopes: $(n + 1)k = 1 - k$

$$n = \frac{1}{k} - 2$$

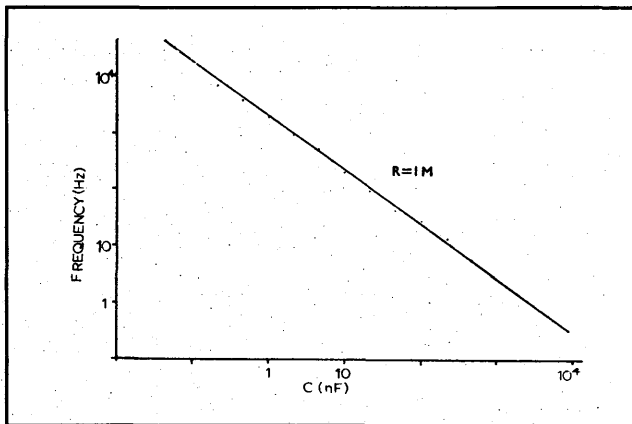
∴ For n = 1 k = ½.

Complementary transistor sawtooth generator



Typical performance

V_{CC} : +15V
 T_{R1} : BC125, T_{R2} : BC126
 R_1 : 2.2k Ω , R_2 : 22k Ω
 R_3 : 15k Ω , R: 1M Ω (pot)
 C: 10nF
 V_{out} excursion is 2.8 to 7V at 1kHz.
 Supply current: 0.5mA



Circuit description

This circuit is related to the corresponding trigger circuit described in Circards series 2. Consider the capacitor in an initially uncharged state. The base potential of T_{R1} is zero and no current flows in either T_{R1} or T_{R2} . The p.d. across R is a large fraction of V_{CC} provided the current in it and in the potential divider are small enough to avoid a large drop across R_1 . As the capacitor charges the p.d. across R falls and with it the rate of charge. When the potential at the base of

T_{R1} exceeds that at the base of T_{R2} by $\sim 1V$ the transistors begin to conduct. This reduces the potential at T_{R2} collector and at the base of T_{R2} through the potential divider. The increased p.d. between the bases that results completes the positive feedback action, ensuring a rapid switching, with the p.d. across the capacitor falling to a low value (determined by the saturation characteristics of T_{R1}). Similarly the potential at T_{R1} collector falls. After the switching transient, the recharging cycle begins. Returning R to the collector rather than V_{CC} provides negative feedback that reduces risk of circuit latching into permanently stable d.c. state.

Component changes

Minimum $V_{CC} = 4V$, oscillation ceases at 3.4V. With $C = 1nF$, $R_{min} \approx 47k\Omega$, $R_{max} \approx 2.6 M\Omega$. Useful range of C: 47pF to 32 μF (tantalum bead). Maximum useful frequency $\approx 70kHz$.

Changing the ratio R_3/R_2 alters the voltage to which C charges.

Circuit modifications

● A resistor may be included in T_{R2} collector (R_4 in Fig. on left) to provide a train of narrow pulses typically of amplitude 0.6V when $R_4 = 100\Omega$. Anti-phase pulses, of amplitude $\approx 14V$ are available at T_{R1} collector.

● Resistor R may be returned to the V_{CC} rail instead of T_{R1} collector, as shown middle. To increase R above 2.6M Ω , current gain of T_{R1} could be increased by replacing it with a Darlington unit.

● Speed-up capacitor C_1 may be added, as shown right, to increase the maximum repetition rate.

● The complementary pair may be replaced by a BFR41-BFR81 pair and all resistors can be scaled down by a factor of about ten to give higher current operation, for example, larger output pulses at R_4 . A V_{CC} up to about 90V may then be used.

Further reading

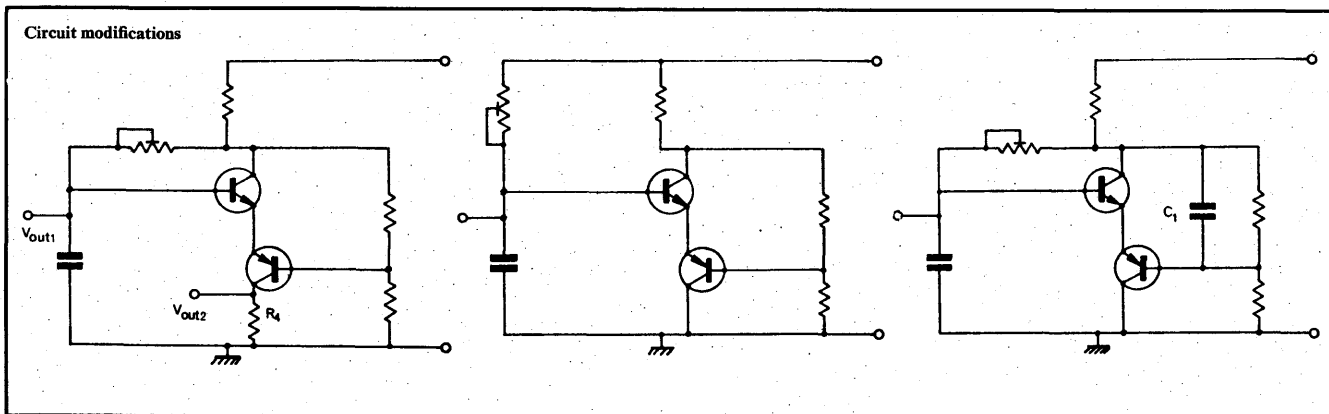
Electronic Circuit Design Handbook, Tab 1971, pp.172 & 184.

Cairns, J. B. F., Linear ramp generator, *Wireless World*, vol. 77 1971, p.604.

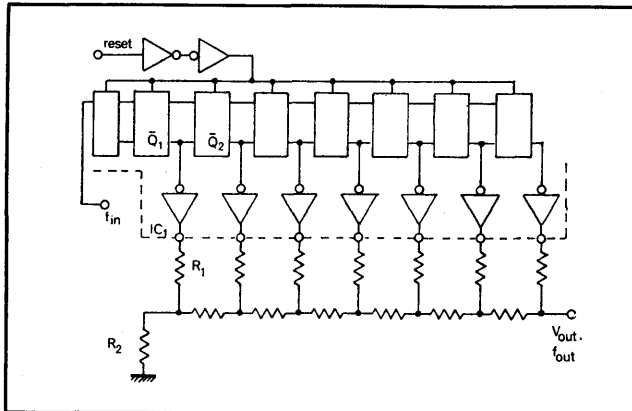
Cross references

Series 2 card 12.

Series 3 card 9.



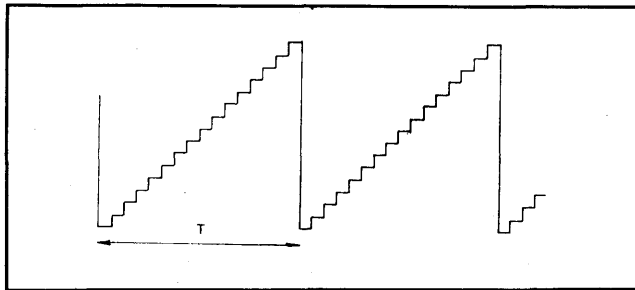
D/A convertor waveform generator



Typical performance

IC₁: CD4024A
 Supply: + 5V
 R₁: 94kΩ
 R₂: 47kΩ
 f_{in}: 12.8kHz
 f_{out}: 100Hz i.e. for
 waveform shown, T =
 10ms.

For a 7-bit counter,
 waveform comprises
 128 steps.
 Minimum input level: 1V
 Minimum input pulse:
 3.5μs



Circuit description

If the output of a binary counter is used via buffer stages to drive a resistor network, a stepped output voltage is obtained which repeats for each cycle of the counter. If the counter is clocked at a definite frequency then the output frequency is fixed by the division ratio introduced by the counter. If the clock rate is variable so is the output voltage with no change in wave-shape, while modifying the network changes the shape without affecting the frequency. The circuit shown is one example where a seven-stage binary counter feeds a resistive ladder network. The buffer elements are contained within the IC package and provide a drive voltage which is accurately defined for light loading. Using identical resistors along the chain, the change from logical '0' to logical '1' at Q₁ causes a change at the input to the ladder which is progressively attenuated, halving for each succeeding stage in the counter provided $R_1 = 2R_2$. Thus the least significant bit from the counter contributes only half the output contributed by the next bit. The result is an output voltage that is an analogue representation of the total number of bits stored in

the counter, and for constant repetition rate and n stages, staircase waveform results with 2^n equal steps.

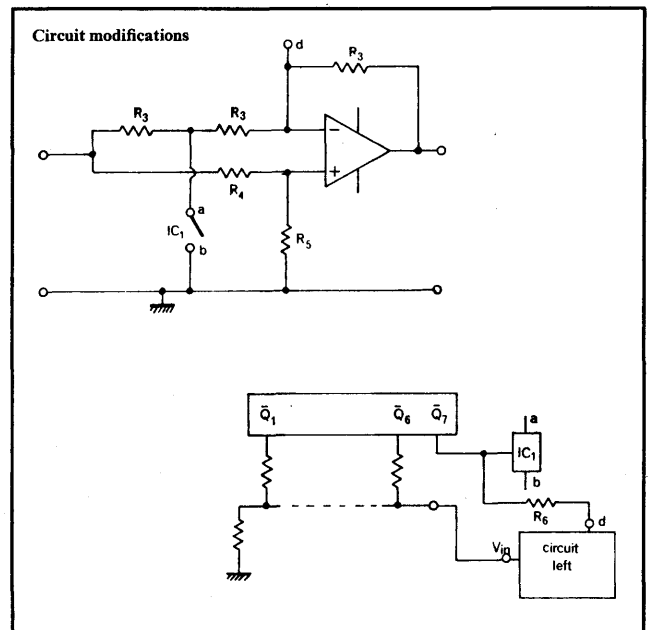
Component changes

Maximum useful output frequency: 1kHz, demanding an input p.r.f. of 128kHz. Minimum pulse level at this rate is 2V, though this varies $\pm 50\%$ with package substitution. An output repetition rate of 0.01Hz is easily achieved. Minimum pulse level is linearly related to supply voltage variations in the range 5 to 10V.

Circuit modifications

An up and down staircase waveform may be generated by inverting each alternate cycle. A suitable inverting amplifier is shown top. Resistor R₃ is 33kΩ; IC₁: $\frac{1}{4}$ (CD4016) c.m.o.s. transmission gate. Resistors R₄ and R₅ are 100kΩ, R₄: 84kΩ and R₅: 16kΩ.

Diagram on bottom indicates the overall connection, where



only six outputs from the counter are used to generate the staircase. The most significant bit-driving pulse is now used to switch both the c.m.o.s. gate and trigger the op-amp inverter. Resistor R₆ is 750kΩ for the above values of R₄, R₅.

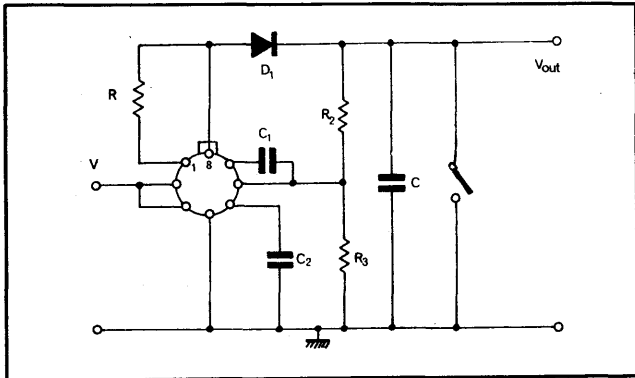
Further reading

COS/MOS Digital Integrated Circuits, RCA, 1972, p.113.
 Naylor, J. R., Digital and analog signal applications of operational amplifiers, *IEEE Spectrum*, May 1971, pp.82-4.
 Staircase-wave generator uses integrated circuits, in 400 Ideas for Design, Hayden, 1971, p.111.
 COS/MOS Integrated Circuits RCA, 1971, pp.72-81.

Cross references

Series 3, cards 3, 11 & 12.

Triggered ramp/trapezium generator



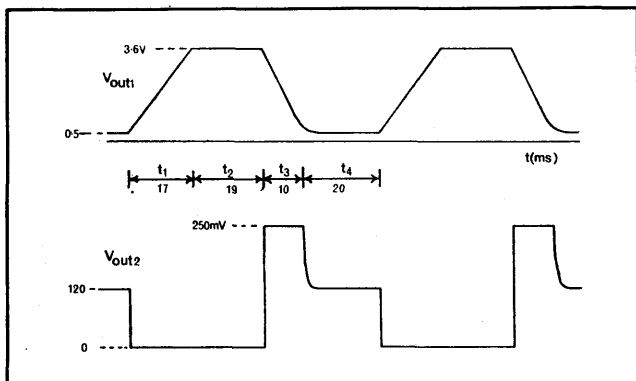
Typical performance

Supply (V): +10V
IC: LM305

R₁: 220Ω
R₂ R₃: 100kΩ
C₁: 47pF; C₂: 100nF
C₃: 22μF; D₁: SD2

With transistor version

(over, left), where Tr₁ is BFR41, R₄ 1kΩ, R₅ 47Ω, V_p = +1V, period is 66ms, width is 30ms, and source res. 50Ω; waveforms are typically as shown



Circuit description

A ramp with an accurately defined maximum value may be desirable for some applications. This can be provided simultaneously with good control of ramp slope, by using an i.c. voltage regulator having internal current limiting. At switch on, capacitor C₃ is uncharged and the switch is open. The regulator remains in its constant-current mode, charging C₃ until the p.d. across it produces a potential at the junction of R₂ and R₃ that matches the internal reference of the regulator. At this point the regulator reverts to its constant-

voltage mode and the output voltage remains constant at a value that may be controlled by the ratio R₂/R₃. During the ramp, the current drawn by the potential divider increases as the p.d. across it rises and this, combined with variation in the current-limiting action at different load p.d.s, gives rise to some non-linearity. For this reason R₂ and R₃ are increased though this marginally reduces output voltage stability. Any convenient means may be used to discharge the capacitor to initiate a following cycle and Tr₁ driven from a pulse source is one example.

Component changes

Maximum useful frequency ≈ 100kHz.

With R₃ equal to 100kΩ, variation of R₂ over the useful range 22 to 150kΩ varies V_{out}, between 8.5 and 3V. Output voltage waveform becomes a ramp either when R₁ is increased to about 470Ω or C₃ increased to about 32μF.

With C₃ equal to 22μF, max. useful R₁ is about 10kΩ (ramp amplitude no longer defined by regulator feedback resistors R₂ and R₃).

With R₁ equal to 1Ω, max. useful C₃ value is about 3000μF.

With R₁ set at 220Ω, V_{out} becomes a square wave with C₃ of 1nF.

Output voltage waveform can be made triangular by adjustment of time constants, e.g. triangle is 1.2V pk-pk, clamped at 3.6V with R₁: 330Ω, C₃: 1000μF & R₅: 100Ω.

Circuit modifications

To give a higher output current rating and to provide a fold-back (negative resistance region) to the regulator, the circuit can be modified to the form shown in the middle diagram.

The waveforms shown right are typical of those obtained with the following V: +10V, R₁: 5Ω, R₂: 3.9kΩ, R₄: 1kΩ, R₅: 15Ω, R₆ + R₇: 1kΩ, C₁: 47pF, C₂: 100nF, C₃: 1000μF, Tr₁: BFR41, Tr₂: BFR81, V_p: +3.6V. Period: 45ms, pulse width: 26ms, pulse source resistance: 50Ω.

With R₆ + R₇ equal to 1kΩ, R₆ should not be greater than about 200Ω to obviate excessive instability of the regulator due to the negative resistance characteristic. Lower level of V_{out} is less well-defined than its upper level due to dependence on V_{CE} (sat) of Tr₁ and V_p amplitude.

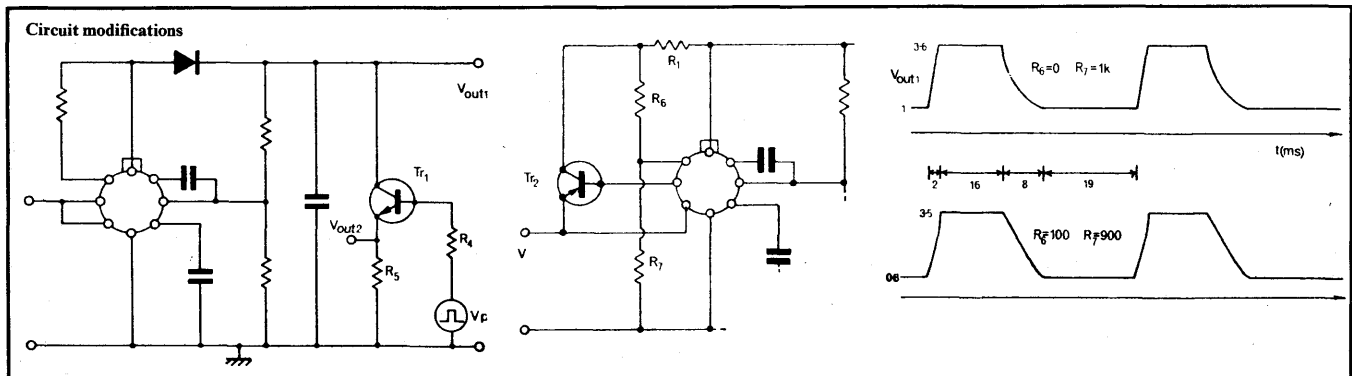
Further reading

Elmgren, K., *Journal of Physics E: Scientific Instruments* (Letters), vol. 5 1972, p.296.

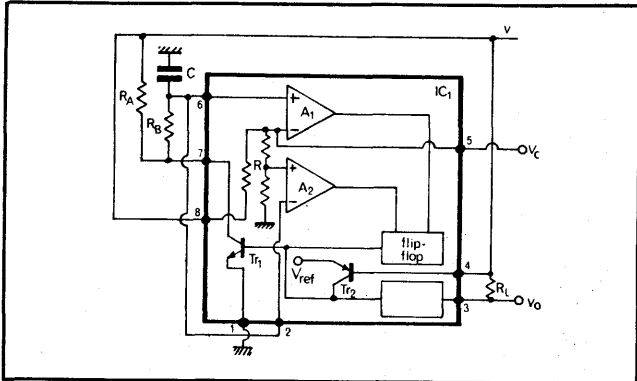
Linear Integrated Circuits Data Book, National Semiconductor 1971, p.31.

Cross reference

Series 3, card 4.



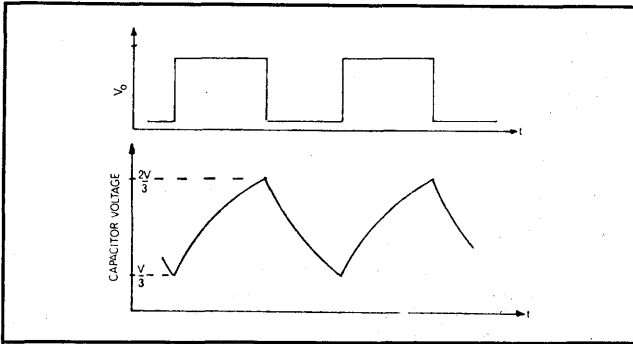
Stable waveform generator using single i.c.



Typical performance

IC: NE555V (Signetics)
 V: +5V
 $R_A: 1\text{ k}\Omega \pm 5\%$
 $R_B: 100\text{ k}\Omega \pm 5\%$
 $C: 10\text{ nF} \pm 5\%$
 $f: 710\text{ Hz}$

Charge time $\sim 0.69 \times (R_A + R_B)C$
 Discharge time $\sim 0.69 \times R_B C$
 Period $\sim 0.69(R_A + 2R_B)C$
 Duty cycle: $R_B / (R_A + 2R_B)$



Circuit description

The i.c. was designed as a versatile timer capable of operation in the astable mode. Frequency and amplitude of the waveform across the capacitor are very stable, and the waveshape can be modified by changing the charge/discharge circuit. Consider the flip-flop in the state that leaves Tr_1 non-conducting. The capacitor charges through $R_A + R_B$ until the high-level comparator reverses the flip-flop. Transistor Tr_1 conducts, discharging C through R_B until the low-level comparator returns the flip-flop to its initial state allowing the cycle to re-start.

For $R_B \ll R_A$ the flyback time is very short, and sawtooth waveforms are possible. Conversely for $R_B \gg R_A$, the time-

constants for the two sections of the cycle become comparable. Comparator input currents are low and high values of R_A and R_B may be used without deterioration of the waveform or loss of timing accuracy. Capacitor waveform is defined by the comparator levels to lie between $V/3$ and $2V/3$.

Unless the load resistance is $\gg R_A$ and R_B , buffering of the output from the capacitor is required. A square pulse output is available which can supply load currents of $> 100\text{ mA}$ with respect to either supply line and without disturbing frequency. A reset function is available that over-rides the charging action and a control voltage that changes the comparators' reference potentials i.e. allows modulation.

Component changes

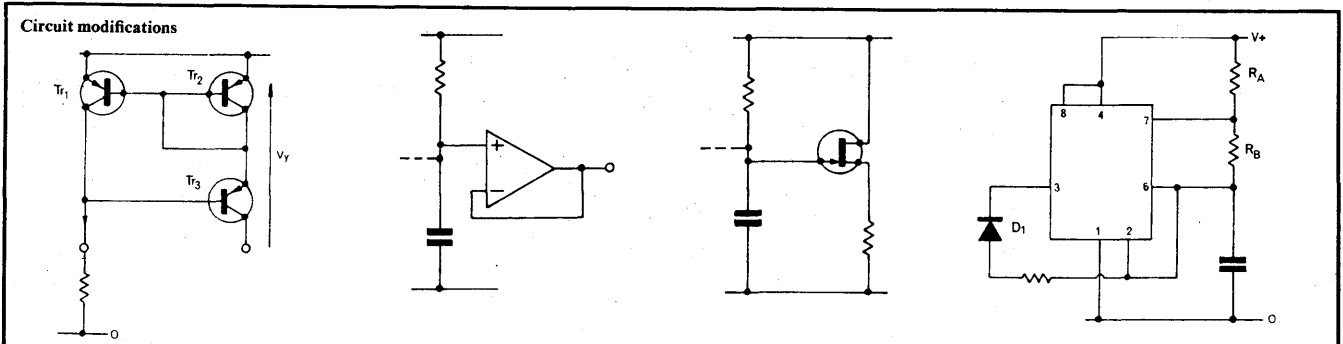
- V: +4.5 to +16V
- $R_A: 1\text{ k}\Omega$ to $1\text{ M}\Omega$
- $R_B: 1\text{ k}\Omega$ to $1\text{ M}\Omega$
- C: 100 pF to $100\mu\text{F}$
- Control voltage (pin 5) varies on and off levels in same ratio, allows modulation of frequency, but also changes amplitude of capacitor waveform.
- Addition of silicon diode in parallel with R_B , conducting on forward stroke makes charge time dependent mainly on R_A . Discharge still depends on R_B i.e. Duty cycle adjustable $\geq \frac{1}{2}$. Diode drop affects accuracy, particularly for low V.
- Output may be synchronized with external waveform fed to control input (pin 5) or triggered by input to trigger point (pin 2).

Circuit modifications

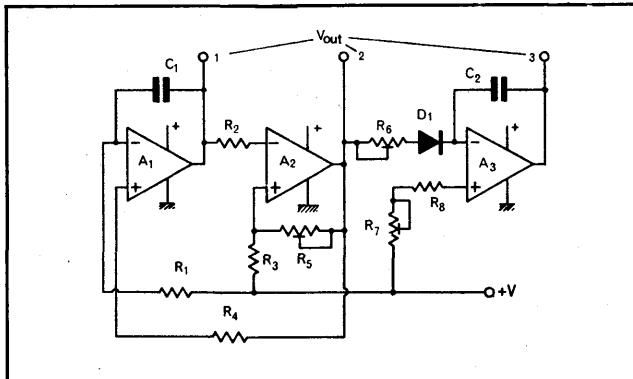
- For linear ramp generation, constant-current charging is required. Matched transistors (as in RCA CA3084) form current mirror in which collector current of Tr_3 (left) is set by current in Tr_1 , with only small influence of collector-emitter p.d. Any alternative current generator with p.d. $< \frac{1}{3}V$ may be used such as that on card 2.
- Capacitor cannot be loaded resistively without disturbance to waveform. Operational amplifier used as voltage follower, or f.e.t. as source follower, are suitable buffers for this (middle diagrams), and corresponding portions of cards 3, 4 & 6 may benefit from the same technique.
- For minimum flyback time the discharge current must be increased. If the flyback time is negligible compared to ramp time, then linear voltage control of the latter gives linear control of frequency. Diagram right shows the main output returned through D_1 to the capacitor i.e. using output current capability to reduce flyback time. Fall-time of $< 1\mu\text{s}$ for $C = 0.1\mu\text{F}$ is possible at $V = +10\text{V}$.

Cross references

Series 3 cards 2, 4 & 6.



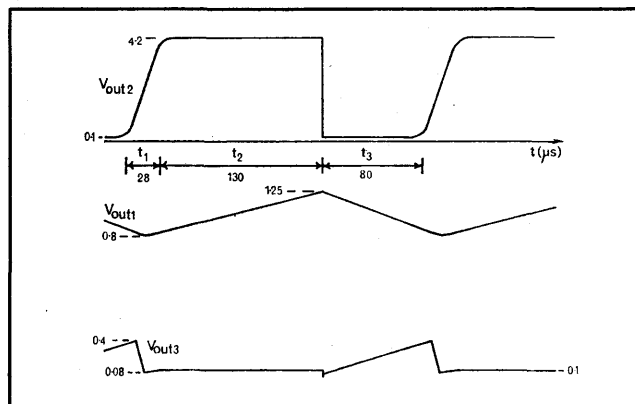
Simple multi-waveform generator



Typical performance

Supply: +5V
 A1-A3: LM3900
 R1: 1M Ω ; R2: 100k Ω
 R3: 1.2m Ω ; R4: 470k Ω
 R5, R6, R7: 1M Ω pot.
 R8: 22k Ω ; C1; C2: 1nF

D1: PS101
 With R5, R7 = 1M Ω &
 R6 = 0 output wave-
 forms are typically as
 shown right.



Circuit description

Operational amplifiers whose output depends on the difference between two input currents can be used as novel waveform generators. Thus A1 integrates the difference in current in R1 and R4. The former is constant and the latter switches between some positive value and substantially zero (the potentials at the inputs of all these amplifiers are about +0.6V w.r.t. the common line, using a single-ended positive supply). If the on-current in R4 is double that sustained in R1, the difference then changes polarity with equal magnitude

for the two polarities. The output of A1 is a linear ramp that at some potential provides a current in the inverting input of A2 that initiates a switching action, reverses the output of A2 and causes the integration to proceed with opposite slope. Resistor R5 controls hysteresis on A2, amplitude of the triangular wave and also frequency. Control of frequency with a single resistor is more difficult than for circuits using conventional op-amp circuits as the ratio R1/R2 has to be maintained for symmetrical triangular waves. The output is fed to a second integrator A3 but at an amplitude sufficient to ensure eventual saturation. Slopes of the edges can be varied by R6 and R7.

Component changes

Maximum useful frequency \approx 20kHz.

Useful C1 range \approx 10pF to 22 μ F.

R5 min \approx 100k Ω (V_{out1} switches to +4.2V).

With R5 = 105k Ω , V_{out1} = 3.7V pk-pk, V_{out3} = 3.3V pk-pk, t_1 = 28 μ s, t_2 = 1.26ms, t_3 = 800 μ s.

R6 max \approx 770k Ω (V_{out3} becomes: +4.3V d.c.).

With R5 = 105k Ω , R6 \approx 490k Ω : V_{out3} is triangular;

R6 > 490k Ω , V_{out3} is a trapezium clamped at 4.3V;

R6 = 0, R7 = 433k Ω to 1M Ω , V_{out3} is a ramp;

R6 = 0, R7 = 11 to 433k Ω , V_{out3} is a trapezium;

R6 = 0, R7 = 5k Ω , V_{out3} is a square wave (anti-phase with V_{out2}) and may be made a trapezium or a ramp waveform by selecting C2 between 10pF and 100nF.

Circuit modifications

● Remaining amplifier in the LM3900 package may be used as an inverter (see left) fed from any one of the existing outputs to provide a pair of antiphase triangular, square, ramp or trapezoidal outputs.

● This amplifier may also be used as a summer for two or more, of the existing outputs (middle) to produce more complex waveshapes.

● If dual polarity supplies are used then R_x above should be reduced to 1M Ω and connected to 0V instead of the +V rail.

● For certain values of R6 and R7, V_{out3} will not be clamped to either a low or a high level. To remove this indefinite state, d.c. feedback can be added to A3 as shown right.

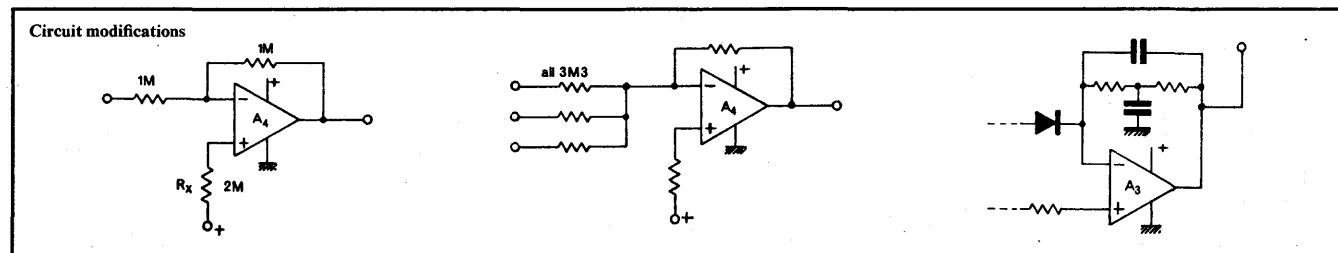
Further reading

LM3900 Quad Amplifier, data sheet and application note, National Semiconductor, 1972.

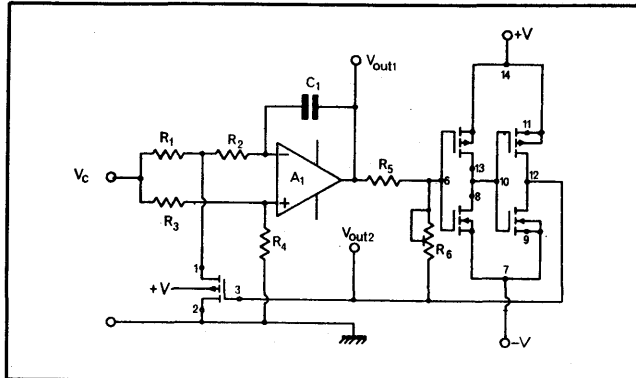
Gledhill, B., Analogue module applications, *Electronic Engineering*, March 1970, pp.64/5.

Cross reference

Series 3, card 11.



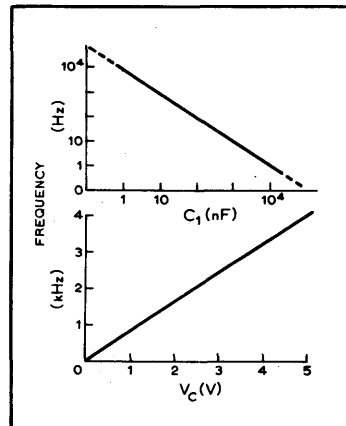
Op-amp/c.m.o.s. square/triangle generator



Typical performance

Supplies: +5V
 A₁: 741
 c.m.o.s. inverters: $\frac{1}{2} \times$
 CD4007AE
 c.m.o.s. switch: $\frac{1}{2} \times$
 CD4007AE
 (R₁ + R₂); (R₃ + R₄):
 47k Ω
 R₅: 100k Ω , R₆: 1M Ω pot.
 C₁: 1nF

of 1.3V pk-pk and V_{out2}
 is 1:1 square wave,
 10V pk-pk.



Variation of frequency
 as function of C₁ (V_C =
 + 1V) and of V_C (C₁ =
 1nF) shown right, with
 R₁ = 2R₂, R₃ = 2R₄ and
 R₆ = 1M Ω .

At 10kHz, V_{out1} is sym-
 metrical triangular wave

Circuit description

One limitation to the basic triangle-square generator is that the output amplitude of the Schmitt depends on the saturation limits of the amplifier/comparator used—the hysteresis and hence triangular wave amplitude and frequency are device-temperature variable. The circuit shown uses a c.m.o.s. Schmitt whose output swings almost exactly to the supply limits provided it is lightly loaded. Such a circuit can be provided by a single c.m.o.s. package while leaving at least one m.o.s. device free to act as a switch driven by the Schmitt output. The switch may be used to invert the current flow within the integrator, or to invert the gain of a preceding

amplifier where the circuit is to be used as a voltage-controlled oscillator.

The circuit makes very economical use of the lowest-cost c.m.o.s. package to provide triangular and square waves whose amplitudes are constant for constant supply voltage. A further advantage is that the Schmitt current is negligible except at the switching point. The main disadvantage is that the c.m.o.s. threshold voltage, while close to V/2, has some tolerance and the triangular wave will have a non-zero mean potential.

Component changes

Maximum useful frequency: 50kHz (C₁ = 1nF; V_C = +1V)
 R₆ min \approx 60 k Ω (loss of triangular wave).

Increasing R₁/R₂ increases +ve slope of V_{out1}, without changing its -ve slope, increases V_{out}, and increases mark-space ratio of V_{out2} (22:1 when R₂ zero).

Decreasing R₁/R₂ reduces +ve slopes of V_{out1}, without affecting its -ve slope, reduced frequency (f_{min} \approx 600Hz), and reduces mark-space ratio of V_{out2} (1:32 min at f \approx 700Hz).

Increasing R₃/R₄ reduces +ve and increases -ve slopes of V_{out1}, reduces frequency (f_{min} \approx 300Hz), and reduces mark-space ratio of V_{out2} (1:60 min at f \approx 400Hz).

Reducing R₃/R₄ increases +ve and decreases -ve slopes of V_{out1}, reduces frequency (f_{min} \approx 20Hz), and increases mark-space ratio of V_{out2} (500:1 max. at f \approx 100 Hz).

Circuit modifications

- Variable voltage control input (V_C) may be derived from a potentiometer (R₇ \approx 5k Ω , left) connected between +V and 0V rails, which provides first-order compensation for supply voltage changes.

- Useful frequency range of the circuit may be extended to about 250kHz by using a fast integrator. Middle diagram shows an example where A₁ is an LM301A, C₂ 10pF, C₃ 150pF and R₈ 5k Ω .

- In place of a c.m.o.s. device, the switch may be realized by a discrete transistor e.g. BC125. Inclusion of R₉, typically 10k Ω , as shown right allows the triangular output to swing symmetrically with respect to 0V.

- A third output is available at pin 10 of the CD4007AE which provides a square wave in push-pull with V_{out2}.

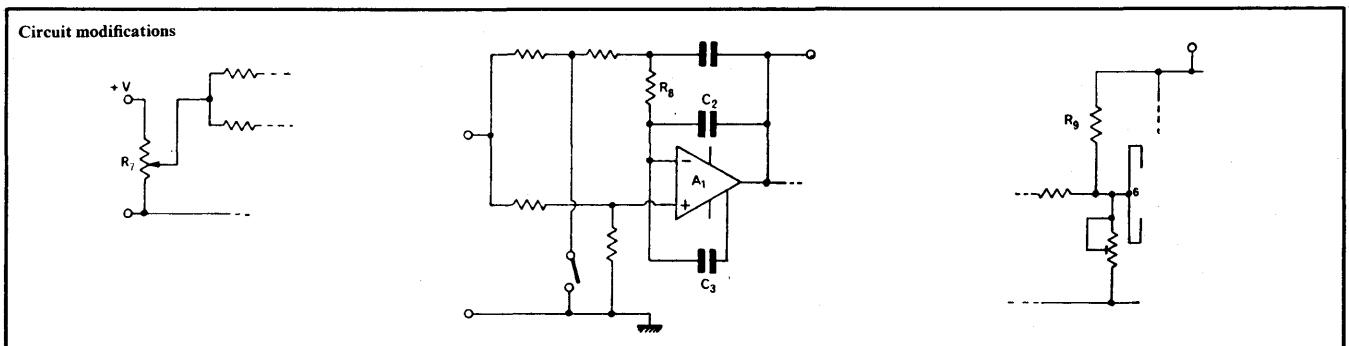
Further reading

Simple square-triangle waveform generator, *Electronic Engineering*, Oct 1972, p.29.

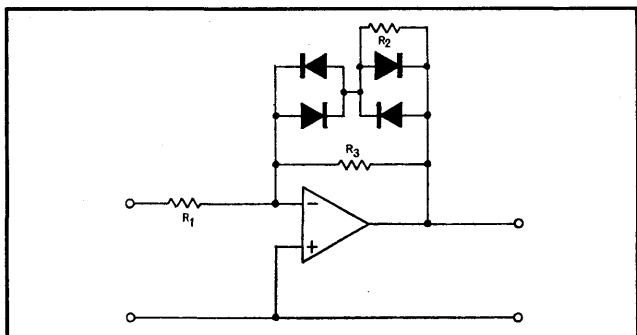
Smith, J. I., *Modern Operational Circuit Design*, 1971, p.224/5.

Cross references

Series 3 cards 1, 2, 5, 7 & 10.

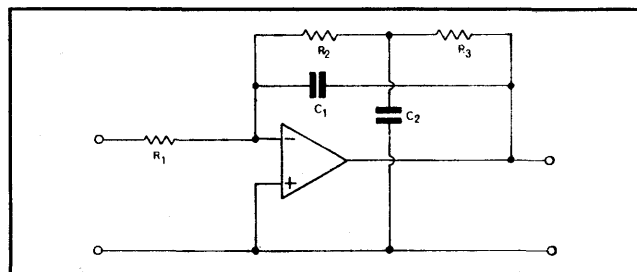


Simple wave-shaping circuits



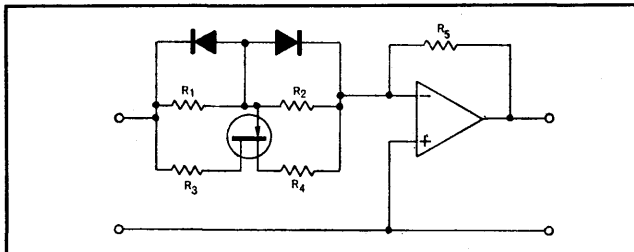
f : 1kHz
 R_1 : $22k\Omega \pm 5\%$
 R_2 : $100k\Omega \pm 5\%$
 R_3 : $27k\Omega \pm 5\%$
 Diodes 1N914
 Adjust input triangular wave amplitude for minimum output distortion ($\sim 2V$ pk-pk) Distortion $< -43dB$ using distortion factor meter. Visual adjustment possible to $-35dB$ to $40dB$.

If a repetitive waveform is fed to an amplifier with a non-linear transfer function, the output waveform differs from that of the input. In the circuit shown the diodes across the feedback resistors are non-conducting for small signals and the output waveform is an inverted version of the input. As the amplitude increases, the diodes are progressively brought into conduction and the output increases more slowly than the input. With the values shown an input triangular wave produces a sinusoidal output with total harmonic distortion $< 1\%$ on two conditions: that the input contains no significant d.c. component, and that the input resistor is adjusted for the particular value of input voltage. Component values were determined empirically with the diode non-linearities smoothing the transitions between the defined regions of the transistor function.



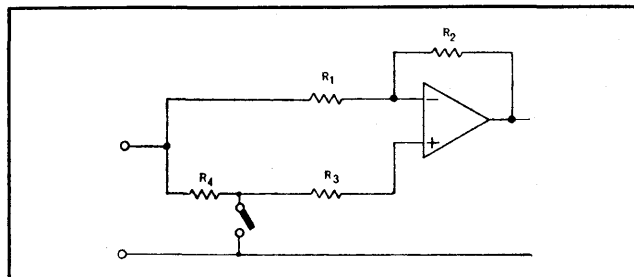
f : 1kHz
 R_1 : $15k\Omega \pm 5\%$
 R_2 : $100k\Omega \pm 5\%$
 R_3 : $100k\Omega \pm 5\%$
 C_2 : $1\mu F \pm 10\%$
 C_1 : $10nF \pm 10\%$
 Output $\propto (1/fR_1C_1)$
 Output amplitude \approx input amplitude for above typical values and sinusoidal/triangular wave inputs.

A third method of generating an approximate sine wave which is not strictly wave-shaping in the above sense is to apply a triangular wave to an integrator with overall decoupled d.c. negative feedback to define the mean output voltage. The integral of a linear ramp is parabolic in form, and the combination of two successive parabolas corresponding to the positive and negative slopes of the triangular waves gives a crude approximation to a sine wave with harmonic distortion of about 4%. The one advantage of this circuit over the previous is that the wave shape is independent of input amplitude, but the output amplitude is inverse to frequency.



f : 1kHz
 R_1 : $1M\Omega \pm 5\%$
 R_2 : $1M\Omega \pm 5\%$
 R_3 : 470Ω
 R_4 : 470Ω } matched to $\pm 2\%$
 Diodes 1N914
 Adjust input triangular wave amplitude as for 1. Distortion $< -40dB$. Select R_5 for desired output, typically 1 to $10k\Omega$.

Placing a non-linear element in the input path also modifies the output, and to convert a triangular wave into an approximate sinusoidal wave an f.e.t. may be used. The source and drain are interchangeable and the diodes ensure that for either polarity of input the f.e.t. is effectively operated with low V_{gs} . At low input voltages, the f.e.t. has a low and relatively constant slope resistance, rising progressively as the input brings it towards pinch-off. If the input voltage has $V \approx V_p$ for the f.e.t., the output peak is just flattened and distortion of less than $< 1\%$ is again possible. The gate bias resistors should be large and the source and drain resistors equal. If the f.e.t. or diode networks are reversed, so is their action—the magnitude of the transfer function increases as the input amplitude rises. Any other devices with controlled non-linearities may replace the above.



Using an ideal switch and $R_1 = R_3$, gain is exactly inverted. Switch may be driven by the squarewave of a square/triangle generator, and the circuit then inverts alternate ramps to give saw-tooth.
 R_1 to R_4 : $100k\Omega$.

An alternative to controlled non-linearity is to introduce an instantaneous change in gain at some precise point in a waveform. This can be conveniently done if a square wave is available simultaneously with the waveform to be modified, as in the triangle/square generators described earlier. If the triangular waveform is passed through an amplifier whose gain is inverted at each peak of the triangular wave, the result is a sawtooth wave at twice the frequency. The switching of the amplifier gain may be carried out using f.e.t. switches as described in previous cards. A further modification involves the superposition of a portion of the square wave on the sawtooth, producing a sawtooth at the original frequency but with a transient at the ramp mid-point.

Cross reference
 Series 3, card 7.

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 multi-meter, by measuring mean or peak values and calibrating them in r.m.s. values. For sinusoidal waves, the relation $V_{rms} = 1.11 V_{mean}$ 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).

- Basic diode rectifiers 1
- Peak/mean/r.m.s. calibrated rectifier 2
- Absolute-value circuits 3
- High-frequency voltmeter for a.c. 4
- Class-B economy rectifier 5
- Potentiometric peak-sensing circuit 6
- Low-frequency measurement of a.c. waveforms 7
- High-current peak/mean rectifier 8
- Simple precision rectifiers 9
- Positive/negative peak detector 10
- Square-law meter circuit 11
- AC adaptor for digital voltmeter 12

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 im-

possibly 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

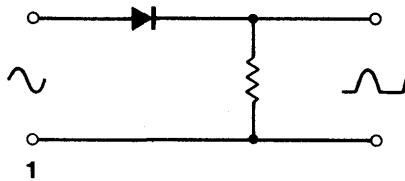
output which is ideally equal to the input when the latter is positive, and an output which is zero when the input is negative. The ideal diode would pass zero current for all conditions when the anode is negative with respect to the cathode, and have zero p.d. when the polarity is reversed. In practical 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.

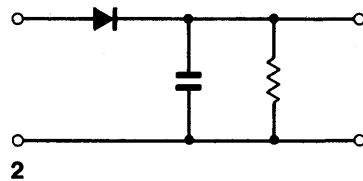
A direct voltage may be obtained directly as in Fig. 2. The capacitor charges on each positive peak of the input, losing some of that charge between peaks into the resistance of any load. To minimize such losses and make the output a more accurate measure of the repetitive peak input voltage, the time-constant 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.)

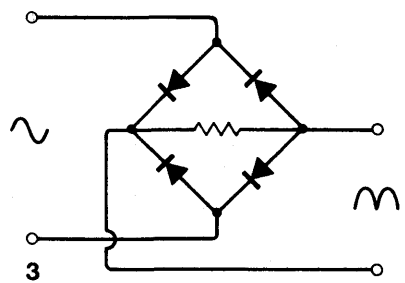
Two methods are available. Bridge rectification as in Fig. 3 requires four diodes to



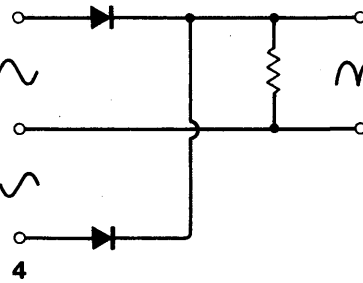
1



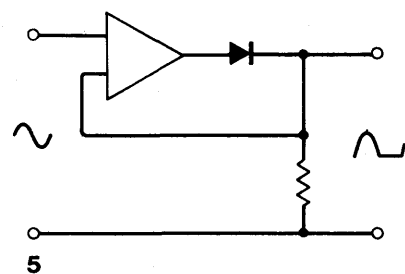
2



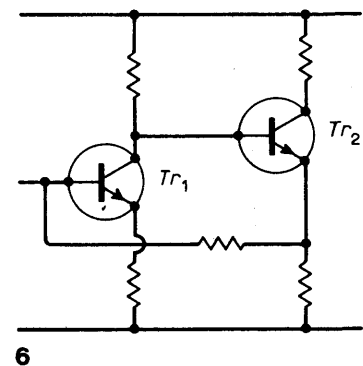
3



4



5



6

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 ($\ll 1$ Hz) 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

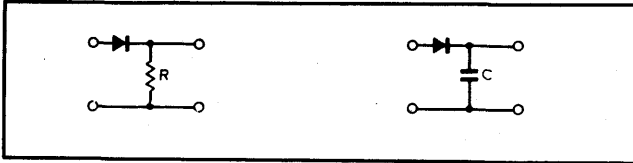
- thermocouples generating an e.m.f. dependent on the power dissipated in a load
- non-linear amplifiers approximating to square-law characteristics where the output can be averaged to give a mean-square reading. A second squaring circuit in the feedback path of a following amplifier gives a square-root 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 ob-

tained as ready-made units from the makers of instrumentation amplifiers. Methods using the square law characteristics of f.e.t.s 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.d.s 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 $V_3 = V_4 = \text{constant}$, 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.

Basic diode rectifiers



Typical performance

Circuit left:

R: 1k Ω ; diode: PS101

Input signal level:

3V r.m.s.

Source impedance: 50 Ω

Useful frequency range:

up to 18MHz; onset of

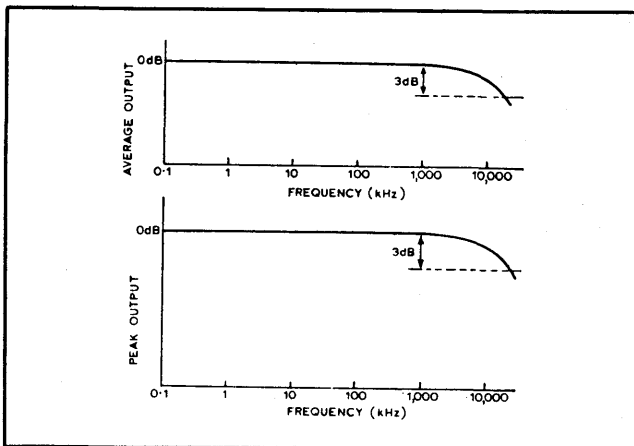
distortion occurs around
2MHz

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- μ A meter movement would allow a 1- μ F 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

Angelo, E. J. Electronics: UJTs FETs and Microcircuits.

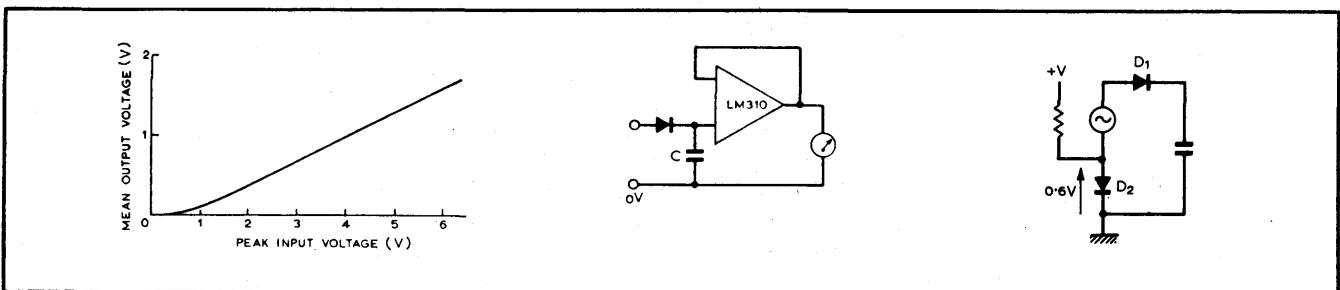
McGraw Hill, 1969. pp.25–34.

Stewart, H. E. Engineering Electronics, Allyn and Bacon, 1969. pp.130–40 and 738–44.

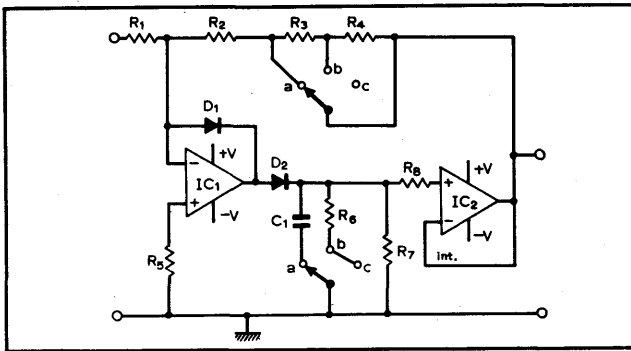
Hemingway, T. K. Circuit Consultant's Casebook, Business Books, 1970. pp.179–95.

Cross reference

Series 4 card 2, 5, 10



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



Components

IC₁: LM301A*IC₂: LM302Supplies: $\pm 15V$ R₁, R₂, R₃: 100k Ω R₄: 22k Ω ; R₅: 47k Ω R₆: 6.8k Ω ; R₇: 100M Ω R₈: 2.2k Ω ; C₁: 68nFD₁, D₂: 1N914

* needs 30pF compensating capacitor

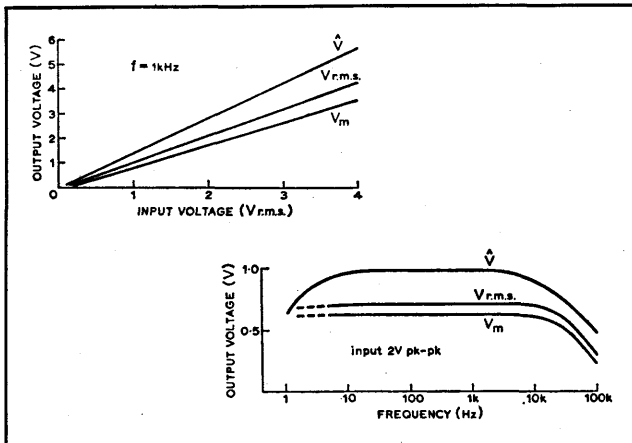
Typical performance

at 1kHz (all ranges)

Input res.: 100k Ω Output res.: < 5 Ω

Load current: 0-10mA

Stability: < 0.5%

 $(V_s \pm 7$ to $\pm 15V, V_{in}$
> 250mV r.m.s.)Accuracy: $\pm 0.5\%$,
 \pm errors in R₁ to R₄

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 D₂ conducts, the feedback path is closed and the high gain of IC₁ 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 D₂ 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 IC₁ reverse biases diode D₂ and the capacitor stores the peak voltage, which value is transferred to the output by IC₂. 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 D₂ is not conducting, D₁ is used to clamp the output of IC₁ by feedback action; this minimizes the recovery time of the circuit prior to the next period when C₁ is to be charged. Mean reading is achieved by removing C₁ 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 R₁ from 1k Ω to 1M Ω gives proportional change in input resistance and input voltage required for given output.

- Increase C₁, R₇ to allow peak detection for lower frequency inputs e.g. for C₁ > 1 μ F peak rectification possible for signals of frequency < 1Hz.

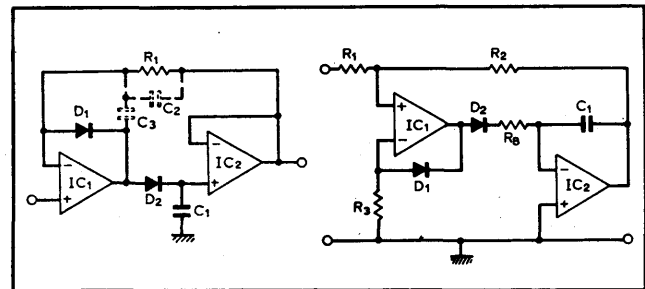
- For true mean-value half-wave rectified let R₃ = 0. To retain r.m.s. equivalent increase R₄ to 120k Ω .

- Replace IC₁ by any general-purpose op-amp (741 or 748 with 30-pF compensation capacitor). Replace IC₂ 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 IC₁. 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 C₂, C₃ in conjunction with R₁ may be necessary with some combinations of amplifiers to avoid risk of high-frequency oscillation. For R₁ = 100k Ω C₂, C₃ may be around 100pF. (left)

- Using IC₂ as an integrator, the additional inversion provided within the feedback loop must be countered by taking the feedback to the non-inverting input of IC₁. Diode D₁ still provides clamping to avoid saturation of IC₁ and R₈ limits charge rate to C₁. (right)



- In general for all such circuits, corresponding sample-and-hold circuits may be constructed by replacing D₂ 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

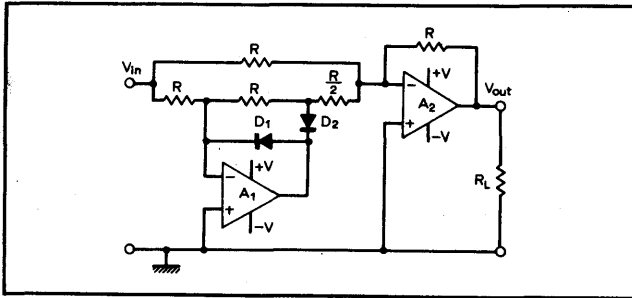
National Semiconductor application note AN20, 1969 p.9.

National Semiconductor application note AN31, 1970, p.12.

Cross references

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

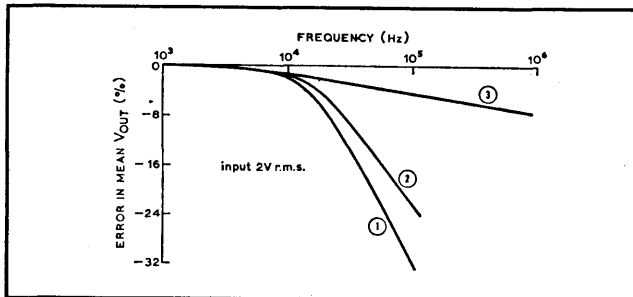
Absolute-value circuits



Typical performance

A₁, A₂: 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 A₂. For negative input voltages, the output of the operational amplifier A₁ swings positive causing D₁ to conduct and D₂ to be reverse biased. Thus, for this polarity of input voltage, there is no contribution of current to A₂ through its R/2 inverting input path. The only input current to A₂ 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 A₁ causes its output to swing negative which reverse biases D₁ and brings D₂ into conduction. Amplifier A₁ thus acts as a unity-gain inverter causing the voltage at the junction of D₂ and R/2 to be -V_{in}. Amplifier A₂ 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 A₂, 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 A₁ are critical if accurate reversal of the gain of the system is to be achieved because for positive input signals the current fed to A₂ represents the difference between those in the two parallel paths. Slew-rate limiting of A₁ 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 A₂ 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 A₁ output goes negative so that D₁ conducts and D₂ is reverse biased. Therefore the junction of R₂ and R₄ is at -V_{in} and as A₂ acts as a unity-gain inverter (non-inverting input of A₂ is virtually grounded), V_{out} = V_{in}. For negative inputs A₁ output goes positive, D₂ conducts and D₁ is reverse biased. The input current in R₁ now divides between R₃ and R₂ plus R₄ in the ratio 2:1, so that A₁ output (and non-inverting input of A₂) is at -2V_{in}/3. Amplifier A₂ now acts as a follower with a gain of 1 + R₅/(R₂ + R₄) = 3/2 for R₂ = R₄ = R₅, hence V_{out} = -V_{in}. Thus V_{out} is a full-wave rectified version of V_{in}. The inverting input of A₁ 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.

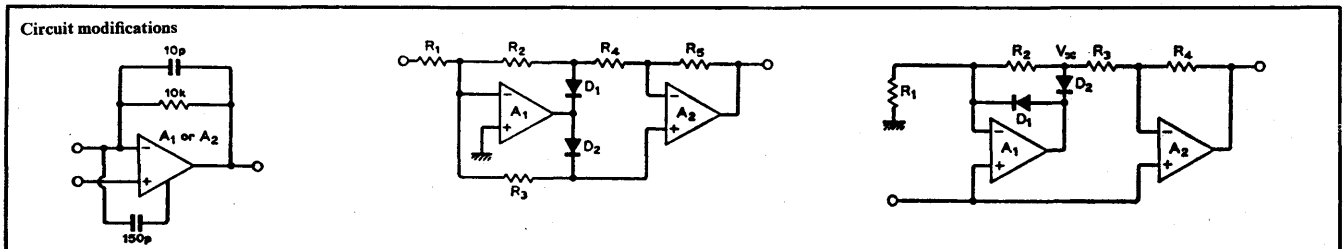
- Circuit shown right provides an output that is the absolute value of V_{in} when R₁ = R₂ = R₃ = R₄/2 and has a high input impedance since the signal source sees the high common-mode input impedances of A₁ and A₂. For positive inputs A₁ acts as a unity gain follower as D₁ conducts and D₂ is reverse biased. Thus V_x = V_{in} and as R₄ = 2R₃, V_{out} = -2V_x + 3V_{in} = V_{in}. For negative inputs D₁ is reverse biased and D₂ conducts so that A₁ acts as a follower with a gain of 2 making V_x = 2V_{in} and again as R₄ = 2R₃; V_{out} = -2V_x + 3V_{in} = -4V_{in} + 3V_{in} = -V_{in}. Hence V_{out} is the absolute value of V_{in}.

Further reading

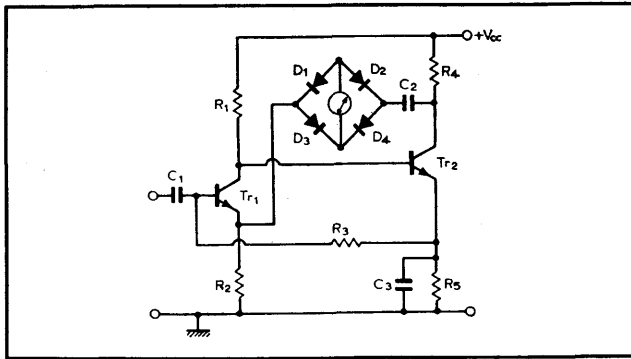
- Application Manual for Operational Amplifiers, 2nd edition Philbrick/Nexus Research, 1968, p.59.
- Egan, F. (Ed.), 400 Ideas for Design, vol. 2, Hayden, 1971, pp.152/3.
- Graeme, J., Op-amps form self-buffered rectifier, *Electronics* vol. 43, no. 21, 12 Oct. 1970, p.98.
- Smith, J., Modern Operational 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

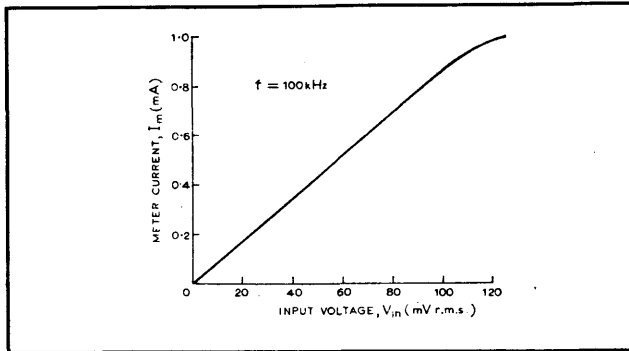


High-frequency voltmeter for a.c.



Typical performance

Supply: +12V, 12mA
 Tr_1, Tr_2 : 1/5 \times CA3046
 Diodes: PS101
 R_1 : 2.7k Ω ; R_2 : 100 Ω
 R_3 : 39k Ω ; R_4 : 470 Ω
 R_5 : 270 Ω
 C_1 : 10 μ F (tantalum)
 C_2, C_3 : 22 μ F (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 100 Ω 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 100 Ω 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.

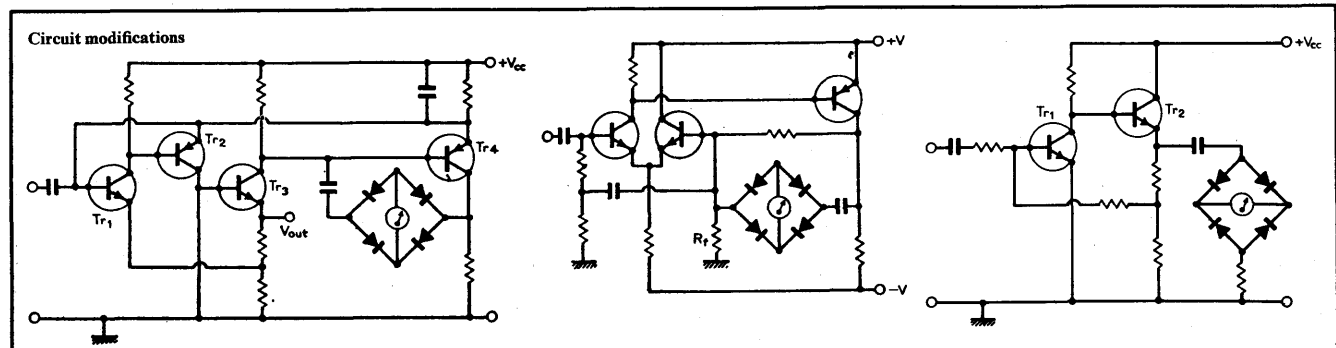
● Shown centre is a three-stage amplifier with overall series-applied shunt-derived negative feedback that raises the input impedance, fixes the voltage gain and provides a well-defined current into the fourth transistor. This has a bridge-rectifier giving shunt-applied feedback for low input impedance 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.

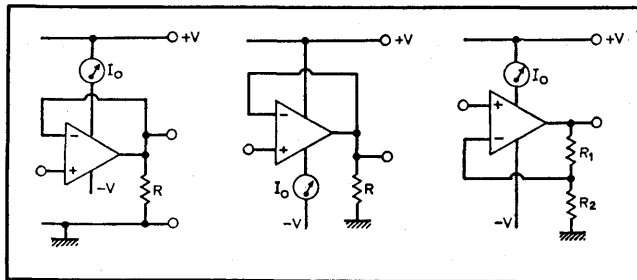
Further reading

Clayton, G., "Operational amplifiers", *Wireless World*, vol. 75, 1969, pp.482/3.

Application of Linear Microcircuits, vol. 1, SGS-UK Ltd, 1968, p.90.

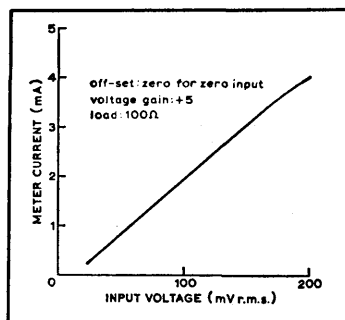


Class-B economy rectifier



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 restricted 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 $\sim 25mA$ and the current in the output stage may be $\sim 1mA$. 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 $\rightarrow 0$; 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

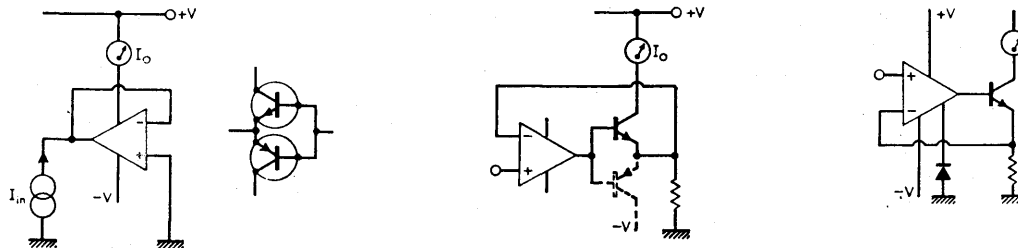
National Semiconductor application note AN31-11, 1972.

Blomley, P., New approach to class-B amplifier design, *Wireless World*, vol. 1971, pp.57-61 and 127-131.

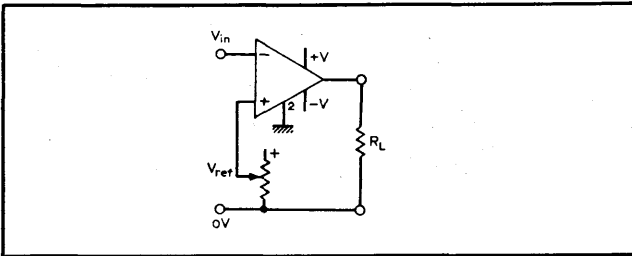
Cross references

Circard series 4, cards 1, 2 & 8.

Circuit modifications



Potentiometric peak-sensing circuit



Typical performance

IC: SN52710

Supplies: +12V, -6V

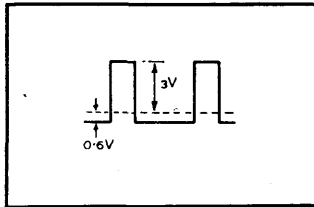
P.R.F.: up to 2 MHz

Minimum pulse width: 200ns

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 100 Ω .
- 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 220 Ω 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.

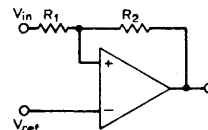
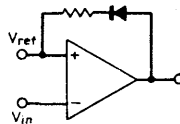
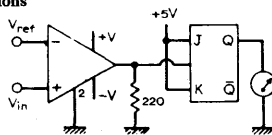
Further reading

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

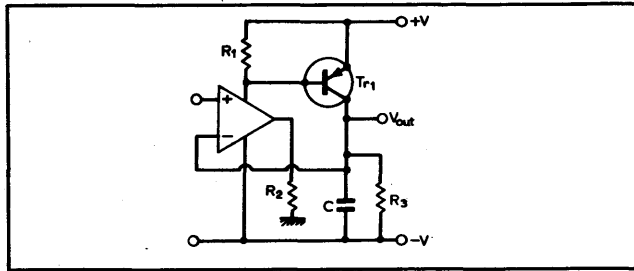
Cross reference

Series 2 card 6.
Series 4, card 10

Circuit modifications

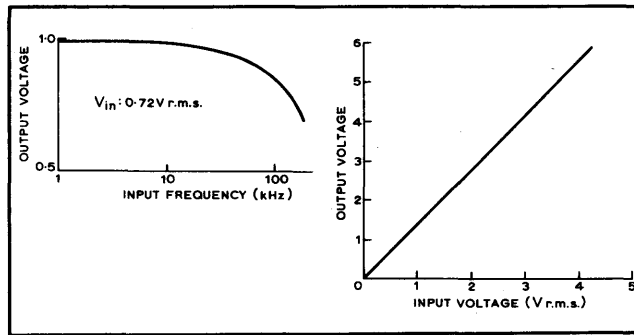


High-current peak/mean rectifier



Typical performance

IC: 741
 Tr₁: BFR81
 Supplies: ±10V
 R₁: 180Ω
 R₂: 470Ω
 R₃: 1MΩ
 C: 0.47μF
 V_i: 1.44V r.m.s. at 1kHz
 V₀: 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 R₂ being drawn through R₁. The p.d. developed across R₁ drives Tr₁ 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 R₁ falling to some minimum level insufficient to maintain conduction in Tr₁. For the remainder of the cycle C discharges under the combined action of R₃ 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 R₁ falls below level at which Tr₁ conducts when input is zero or negative.

C: 1nF to 1000μF
 R₃: 100k to 10MΩ
 R₁: 100 to 220Ω

Raising R₁ further is likely to leave Tr₁ conducting permanently; too low a value requires excessive signal drive.

R₂: up to 1kΩ. With s/c the non-linearities are exaggerated; high R₂ 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 (R₂ + R₁)/R₁ (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 IC₁ drives Tr₁ into conduction and any current through resistors R is absorbed. For negative-going input IC₂ similarly drives Tr₂.

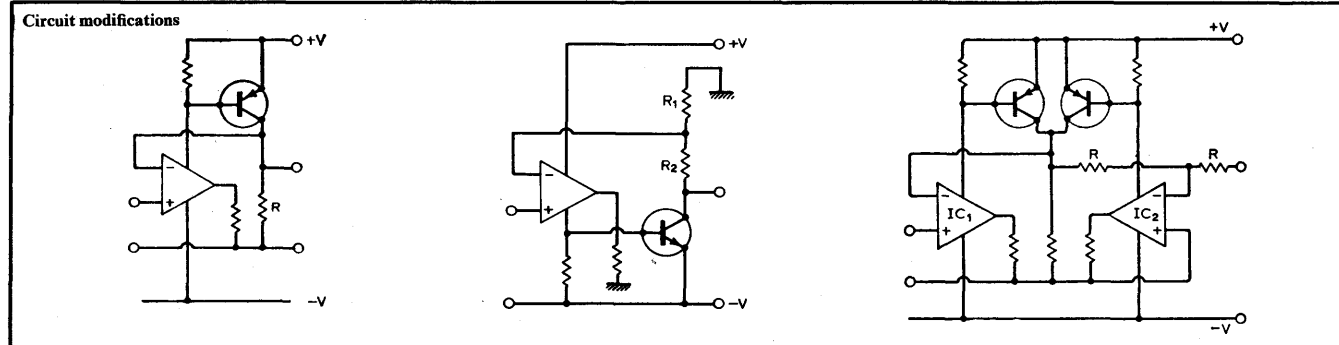
Further reading

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

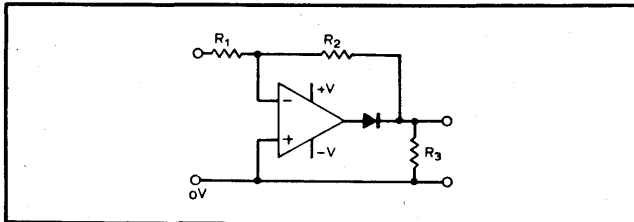
Cross references

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

Circuit modifications



Simple precision rectifiers



Typical performance

IC: 741
 Diode: PS101
 Supplies: $\pm 15V$
 R_3 : $10k\Omega$; R_2 : $3.3k\Omega$
 R_1 : $6.8k\Omega$
 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 :

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

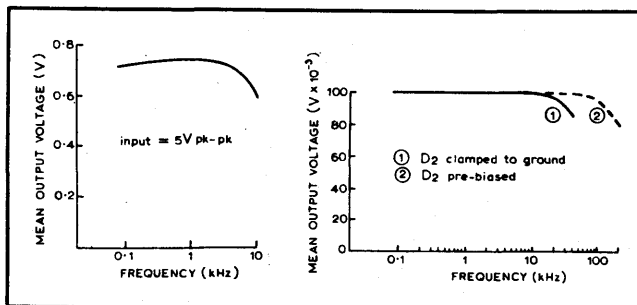
$$= \frac{V_{in} \cdot R_2}{R_1}$$

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

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 $1V/\mu s$.

Component changes

- Supply voltages can be reduced to $\pm 3V$ 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 : $10k\Omega$, R_2 : $9.8k\Omega$. Useful range of R_3 : 3 to $10k\Omega$, R_4 : $6.8k\Omega$. 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} .

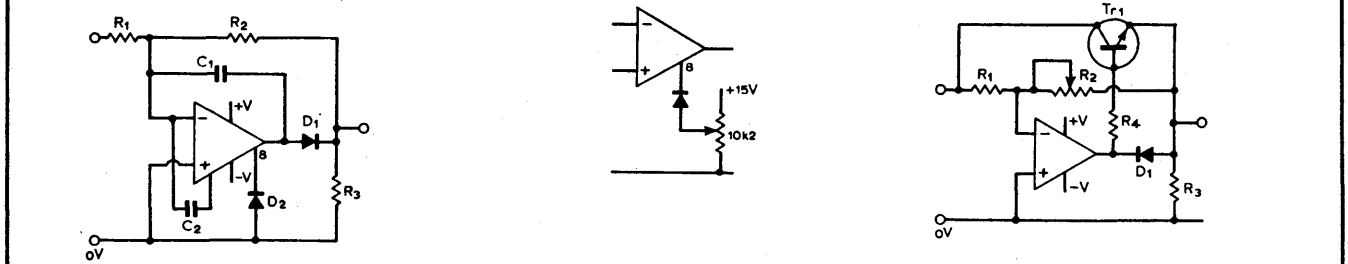
Further reading

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.

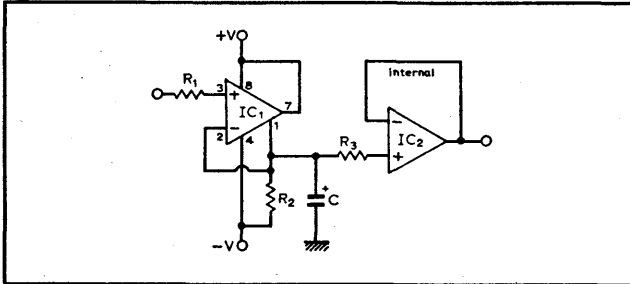
Cross reference

Series 4, cards 2 & 3.

Circuit modifications



Positive/negative peak detector



Typical performance

IC₁: LM311
 IC₂: LM310
 Supplies: $\pm 15V$
 R₁: 2.2k Ω
 R₂: 1M Ω
 C: 1 μF (35-V tantalum)
 R₃: 10k Ω
 Ripple <1% down to 200Hz

Peak detection for inputs <100mV to >20V pk-pk.
 For supplies of $\pm 10V$, 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 R₂ 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

IC₁: 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).

IC₂: 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.

R₁: 1.5k to 4.7k Ω .

R₂: 220k to 3.3M Ω ; 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 R₂. Use of tantalum for higher values ensures that risk of r.f. oscillation on peaks is minimized.

Supply voltage: ± 5 to $\pm 18V$.

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.t.s 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 $\pm 5V$ 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 $\pm 5V$ 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

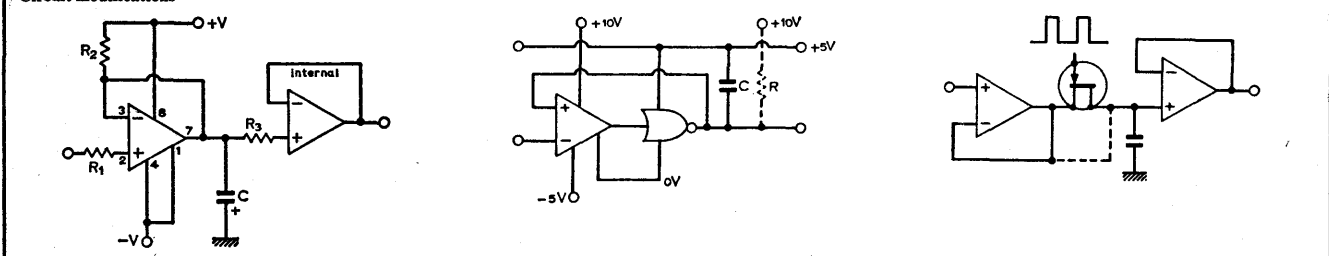
National Semiconductor, LM311 voltage comparator data sheet, 1970.

Positive peak-detector for fast pulses, Applications of Linear Microcircuits, SGS, vol. 1, p. 98.

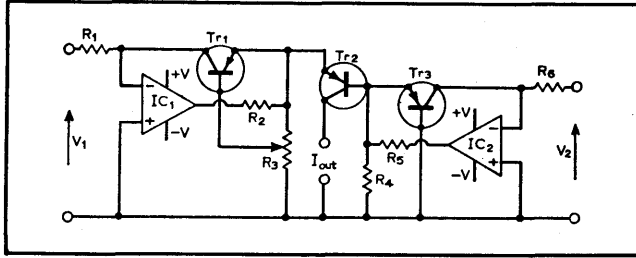
Cross references

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

Circuit modifications

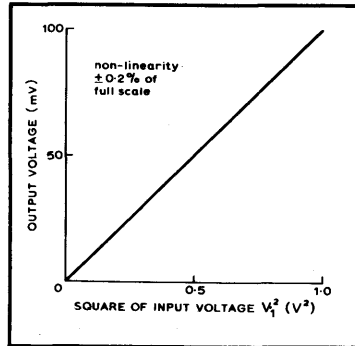


Square-law meter circuit



Typical performance

- $V_s: \pm 10V$
- IC₁: 741
- IC₂: 741
- Tr₁, Tr₂, Tr₃: matched n-p-n transistors, CA3046 (RCA) using three out of the five transistors in package
- R₁: 10k Ω ; R₂: 220 Ω
- R₃, R₄: 100 Ω
- R₅: 220 Ω ; R₆: 10k Ω



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 IC₁ and Tr₁, IC₂ and Tr₃ 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 Tr₂ is of the form $\log_e A - \log_e B$ where A and B depend on V_1 , V_2 respectively. Collector current of Tr₂ is then proportional to the antilog_e of the p.d. between its base and emitter (see Theory). By tapping the base of Tr₁ onto R₃ the effective output voltage from this first circuit can be made any desired multiple of Tr₁ base-emitter p.d. provided that the base current is much less than the current in R₃. The output voltage fed to the emitter of Tr₂ then becomes proportional to $2\log_e V_1$ or $\log_e V_1^2$ when R₃ 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

- IC₁, IC₂: 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.
- R₂, R₅: may be omitted if other means used to avoid oscillation (they reduce loop gain in conjunction with R₃, R₄).
- R₄: May be omitted subject to above precautions or may be tapped as with IC₁.
- R₁, R₆: 1k to 100k Ω , setting sensitivity of circuit. At both high ($\gg 1mA$) and low ($\ll 1\mu A$) current transistors depart from log law; op-amp input current limits low-level operation for particular circuit given.

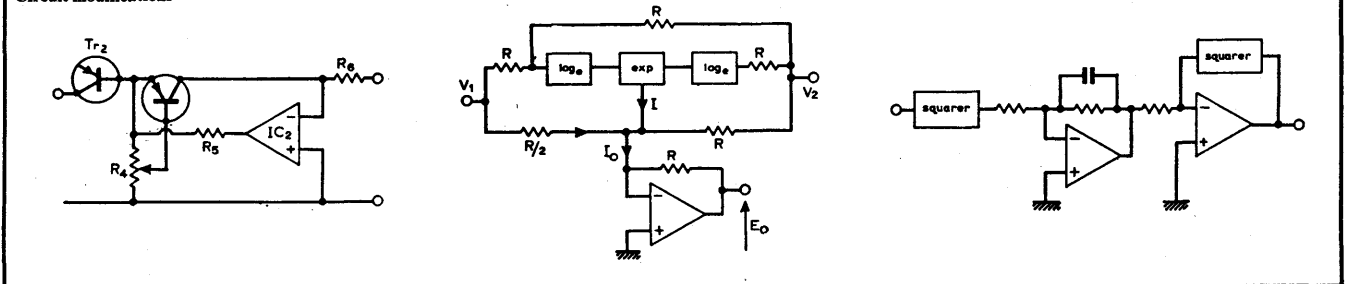
Circuit modifications

- If IC₂ drives the transistor, but with the base of the transistor taken to a tap on R₄, 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 Tr₂ 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 Tr₂ 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.

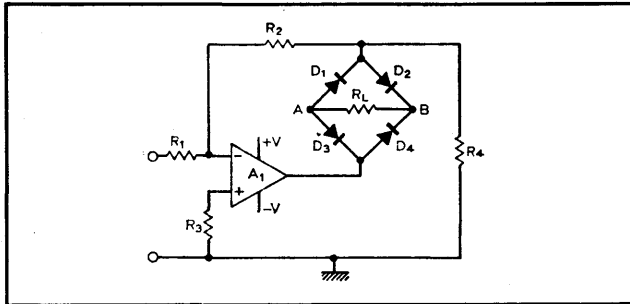
Further reading

- Ehram, 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



A.C. adaptor for digital voltmeter



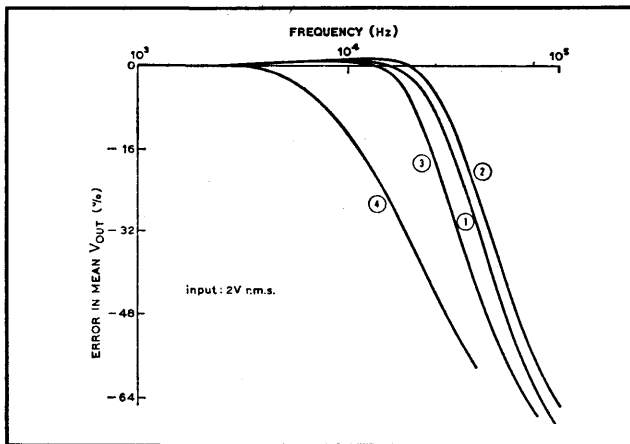
Typical performance

Supplies: $\pm 15\text{V}$, $\pm 2\text{mA}$ A₁: 741

Diodes: CA3019 (part)

R₁: $20\text{k}\Omega$, $\pm 5\%$

R₂, R₃, R₄, R_L:
 $10\text{k}\Omega \pm 5\%$
 Source res: 60Ω
 Error in mean V_{out} :
 $\approx +3.7\%$



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 sinewave 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 A₁ an operational amplifier that allows the use of feed-forward compensation, as shown in card 3.

Component changes

Replacing the CA3019 diode bridge with $4 \times$ PS101 silicon diodes typically produces the response shown in curve 2: low-frequency error in V_{out} (mean) $\approx +3.7\%$.

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) $\approx +3.7\%$.

Useful range supplies: ± 3 to $\pm 18\text{V}$.

Useful range of V_{in} : 350mV to 4.2V r.m.s. R_4 min for V_{in} min and no significant peak clipping $\approx 15\Omega$.

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 Tr₁: BC126, Tr₂: BC125, R₅, R₆: $1\text{M}\Omega$, R₇, R₈: $10\text{k}\Omega$, C₁: $4.7\mu\text{F}$, C₂: $22\mu\text{F}$, A₁: 741, D₁ to D₄: 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. Tr₁ 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 C₂ value. V_{out} can be made equal to the r.m.s. value of V_{in} by scaling $R_2 = (2\sqrt{2R_4})/\pi$.

● 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

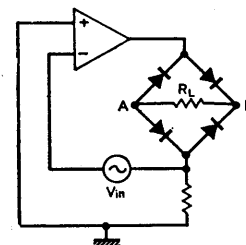
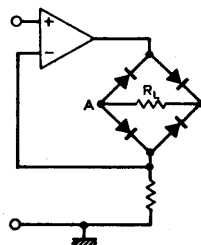
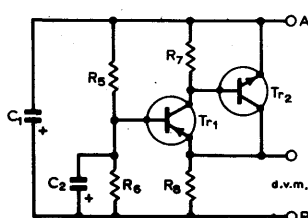
Gellie, R. W. & Klein, A. G., Accurate a.c.-d.c. converter for low frequencies, *Electronic Engineering*, 1967, p.484.

Dromgoole, M. V., Op.amp. a.c. millivoltmeter, *Wireless World*, 1970, p.75.

Cross references

Series 4, cards 3 & 7.

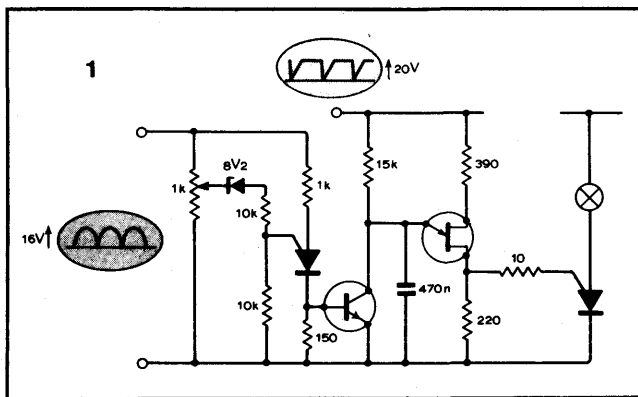
Circuit modifications



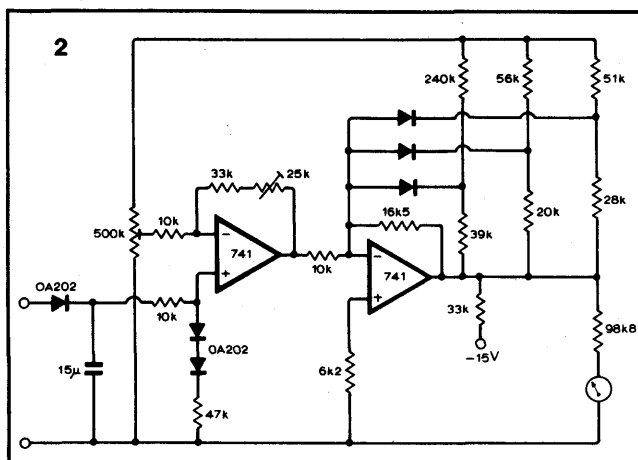
A.C. measurements

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 falls 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).



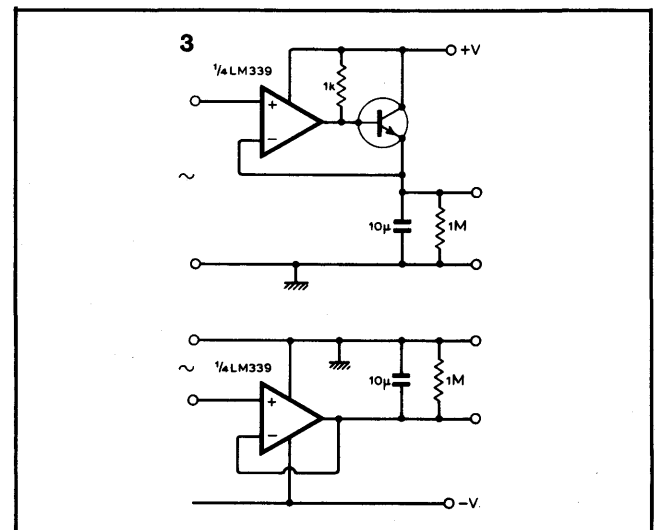
2. For measurement on telephone circuits and other equipment conforming to similar specifications, a meter scaled linearly



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.

Wireless World Circard

Series 5: Audio circuits

The vast majority of these low-level audio circuits use the 741 operational amplifiers, which need dual-polarity power supplies. The tape head preamplifier is one exception, where the low-noise National or RCA op-amps are recommended. One card, entitled economy i.c. audio circuits, gives a selection of eight circuits using the LM3900 in a way that allows use of a single-ended power supply. Five cards deal exclusively with preamplifiers of various kinds, one a multi-input type catering for all the usual signal sources. Card 1, detailing an equalizer for magnetic pickups, shows the technique of reducing cable capacitance by using double-screened cable.

Most of the remaining cards deal with higher-level circuits, for mixing, filtering and tone controls. Cards 3 and 7 show simple 12dB/octave scratch and rumble filters, with useful modifications that can give a variable attenuation slope and cut-off frequency with a single potentiometer, the rate varying between 6 and about 18dB/octave.

The two tone control cards also give useful circuits, card 6 particularly with its simple multi-section circuit (the "graphic" or "room" equalizer) using only RC (Wien) networks, rather than the more common RLC circuits. The circuits of card 2 give two variants of the now-famous Baxandall tone control circuit, one using an op-amp and one using a field effect/bipolar transistor combination to give better linearity.

- Magnetic cartridges/RIAA equalization 1
- Tone control circuits/Baxandall 2
- Rumble filters 3
- Tape-head preamplifier 4
- Audio mixer/operational amplifier 5
- Multi-section tone control system 6
- Scratch filters 7
- Microphone preamplifiers 8
- Impedance matching and transforming 9
- Economy i.c. audio circuits (using LM 3900) 10
- Ceramic cartridge preamplifier 11
- Multi-input preamplifier 12

Audio Circuits

Cinemascope or the Magic Lantern? The breadth of choice available to the user of equipment reproducing audio signals is just as great. We are here seeking the happy medium and sidestepping such difficulties as to whether the medium should be disc or tape — or for Menuhin fans, the message.

The starting point is the assumption that the signals though complex can be represented as a mixture of sinusoidal waves of different amplitudes with frequencies lying between certain limits, say 20Hz to 20kHz. Generally the aim of good audio equipment is to produce at the ear of the listener a pattern of sound most closely resembling that which he would have heard as a direct listener to the original sound source. The system has to take account of the characteristics of the transducers at both ends of the chain as well as any intervening media used for storing or transmitting this signal.

If the input transducer had a linear amplitude response and gave the same output voltage for a given sound intensity regardless of frequency, then the following amplifiers could themselves have a linear response. The design of such amplifiers, with the aid of modern technology in the form of operational amplifiers is by now routine. There are three distinct departures from this idealized existence.

- The output voltage for constant signal strength may be frequency dependent in some controlled manner e.g.: tape-head e.m.f. proportional to frequency for constant amplitude recorded signal.

- The signal may have been recorded and/or processed by some preceding stage with some characteristic defined according to some standard (R.I.A.A., B.S., C.C.I.R. etc).

- Imperfections in some other part of the system may have resulted in anomalies in the desired response e.g.: resonance effects in transducers.

Any one of these would call for correcting action in the amplifying chain, though in some cases as in the design of loudspeakers, resonance effects in the speaker itself can be dealt with by careful design of the enclosure. As each transducer is a very complicated mechanism involving the interaction of several electrical and mechanical properties it is common to operate them with amplifiers whose impedance characteristics are closely controlled, thus eliminating one possible source of variation in performance. This article considers only the input transducers, such as microphones, tape-heads, and assumes that any succeeding power amplifier/loudspeaker combination can have its

imperfections accounted for by tone controls.

The matching problem at the input reduces the design of amplifiers whose input impedance is either equal to, much less than, or much greater than that of the source. Equal source and input impedances are used in line amplifiers where, for example, input, output and attenuator resistances might be 600Ω . This allows for easy calculation of power levels at all points in a system, and for the interconnection of multiple elements in a system. On the other hand, even within such a system the power output amplifier might be designed to have an output resistance $\ll 600$ ohms so that several such loads might be paralleled without diminishing the power fed to each.

A second important feature of the matched impedance condition is that it maximizes power transfer from a source of given e.m.f., and internal resistance. In most modern circuits using heavy negative feedback, the natural impedances tend to be either very high or very low and there is then no advantage from a power transfer standpoint of artificially modifying their terminal impedances to some arbitrary value. To do so simply throws away power in the passive network added for this purpose.

In the case of very small signals where noise is a severe problem, matching of impedances plays an important part. A moving-coil microphone having a low internal impedance (e.g. 200Ω) generates a low e.m.f. of, say, $100\mu\text{V}$ r.m.s. Fed directly to a semiconductor amplifier, the input noise voltage would be relatively large, while it would be possible to have a high input impedance using series negative feedback. A step-up transformer of large turns-ratio would greatly increase the signal e.m.f. at the amplifier input and would dominate the noise voltage. However the effective source impedance seen by the amplifier would also be raised by the transformer action and with it the contribution to noise due to the amplifier's input noise current. The optimum condition is when the contributions due to noise voltage and current generation are comparable. Other parameters such as amplitude response are also affected but the condition chosen is often close to the matched condition.

For microphones, the mechanical properties are normally designed so that they are self-equalized, i.e. that they give an output e.m.f. that depends only on the sound intensity and not on frequency. The most common microphones are magnetic in some form, variants including moving-coil, moving-iron and ribbon

microphones. Reduction of the moving mass to extend response tends to reduce both sensitivity and impedance with the problems described above. Crystal microphones are used for low-cost applications such as simple cassette recorders and require a high input impedance pre-amplifier to avoid attenuation at low frequencies where the capacitive reactance of the microphone increases. The pre-amplifier design is similar to that for crystal/ceramic pickups, i.e. a flat response and generally an impedance in excess of $1\text{M}\Omega$, possibly up to $10\text{M}\Omega$ or more for low capacitance units with extended low-frequency amplitude response. Alternatively, the feedback may contain a capacitance whose change in reactance compensates for that of the transducer.

These ceramic elements could in principle be designed to give a frequency-independent output when used for record reproduction, but another factor enters the argument. During recording, signals are first passed through frequency-dependent amplifiers. These have strictly controlled characteristics, usually referred to as R.I.A.A. and further defined in BS1928. If all signals were recorded with a so-called constant-velocity characteristic it would be found in practice that the amplitude at low frequencies would result in breakthrough between neighbouring sections of the groove. This is because constant velocity fixes the velocity at the zero-crossing point of the signal.

At low frequencies the longer period would allow proportionately larger excursions. Hence low-frequency signals are recorded with amplitude proportional to signal e.m.f. (whereas a velocity-proportional recording would otherwise have merits since a magnetic playback element would re-convert that velocity back into e.m.f. proportionately). This constant amplitude characteristic merges into a constant-velocity region at around 1kHz, but at still higher frequencies the recording again changes to constant amplitude. The reason is different. The majority of the noise in any system is concentrated in the higher octaves as in most cases noise is proportional to bandwidth. By emphasizing high frequency signals during the recording process and reversing the procedure on playback, the overall amplitude response remains linear, but any noise due to the record surface and playback pre-amplifier is diminished as it is relative to a much larger signal. Noise accompanying the original signal emerges from the system at an unchanged ratio.

This recording characteristic of BS1928 accommodates the larger low-frequency amplitudes common in music, and does not lead to distortion at high frequencies as the signal amplitudes are relatively small. The playback transfer function is

$$T_0 = k \frac{(1+j\omega T_2)}{(1+j\omega T_1)(1+j\omega T_3)}$$

Where $T_1 = 75$, $T_2 = 318$ and $T_3 = 3180\mu s$. To achieve this with a magnetic cartridge, the preamplifier input resistance should be higher than that of the cartridge at all frequencies of interest, or should have a fixed value that can be allowed for in tailoring the cartridge response in terms of its electro-mechanical properties. A typical value is $50k\Omega$. The voltage gain must fall between 50Hz and 500Hz at 6dB/octave, passing through a point of inflection at 1kHz, and falling again at 6dB/octave beyond 2.2kHz. These three time-constants may be defined by three separate CR circuits; in some cases two of the time constants are achieved using a single capacitor in a suitable network of resistors.

With ceramic cartridges, equalization is not a result of circuit design but of the transducer itself. The various parameters such as compliance as well as resonances are carefully combined to provide a good approximation to the desired equalization subject to correct loading as outlined above. Where an amplifier has in-built equalization (i.e. for magnetic cartridge) then a separate network may be inserted between the ceramic cartridge and that input to remove the effect of that equalization — a cumbersome process that might be called re-de-equalization.

Tape-recorded signals followed a C.C.I.R. characteristic recently re-defined and extended in BS1568 part 1. There is a low-frequency time constant identical to that in the R.I.A.A. curves, i.e. a time constant of $3180\mu s$, with one further time-constant depending on tape speed, but giving a response that is constant above a particular frequency. These characteristics are quite independent of any imperfections in particular combinations of heads and tapes though intended to optimize their operating conditions. Feedback networks are then similar to those for magnetic cartridge pre-amps equalized as above though requiring only two CR time constants in the ideal case. In practice the imperfections of the system may force for example the addition of some treble boost on playback, operating in the 5 to 20kHz region. This is not covered by any standard, but may readily be incorporated by a further decrease in feedback factor in these regions.

Once these preamplifiers have converted the transducer outputs into voltages bearing a nominally linear relationship to the original sound intensity, it might be thought possible simply to amplify the signals further and apply them directly to an output transducer. Such trusting simplicity exposes one to ridicule

Table 1: proposed classification for loudspeakers.

Frequency range (Hz)	Title
10- 30	grunter
30-100	boomer
100-300	roarer
300- 1k	crooner
1k- 3k	howler
3k- 10k	screamer
10k- 30k	screacher

for ignorance of that recent discovery. Finagle's axiom on reproducing circuitry and equipment — FARCE for short — viz "All signals are equalized but some are more equalized than others". Most audio systems use one or more circuits to modify the amplitude response of the signals passing through them to correct for this effect.

Where unwanted material occurs at the extremes of the spectrum then low-pass or high-pass filters are used for sharp attenuation of these unwanted signals with minimum attenuation of the desired signals. These filters when used in audio equipment are generally called scratch and rumble filters respectively but the basic principles underlying them are the same (see Circards series 1). Second- or third-order filters are used, and as the ultimate judgement of these audio systems is rightly the subjective one of a listening test, the choice of filter characteristic is often empirical.

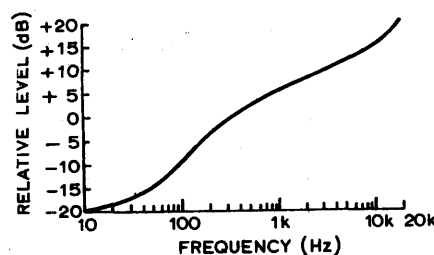
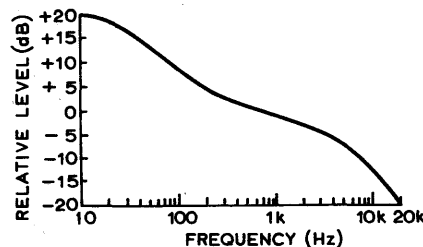
During such a listening test, the parameters of the room housing the loudspeaker plays a large part, while sound sources including commercial recordings are not above suspicion in respect of the linearity of amplitude response. Even if all such sources reached the impeccable standards which the engineers concerned strive so successfully to meet, there would remain the personal preferences of the user. It takes a brave man to refrain from just-a-touch on the tone controls when demonstrating the

superiority of his latest equipment to a fellow enthusiast (competitor?). Of all the tone controls proposed, the most generally accepted is due to P. J. Baxandall, basing itself on a feedback rather than a passive network. This allows for true boost or cut to either low or high frequencies relative to an unchanging centre frequency, generally 1kHz. Two potentiometers are used, adjusting the feedback in the two frequency regions separately around a virtual earth amplifier such that the gain in these regions varies typically from 0.1 up to 10 i.e. 20dB. The higher the quality of the sources and other links in the chain the smaller the range covered by these tone controls need be.

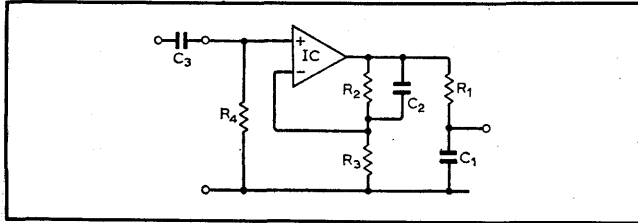
More complex tone controls may be used to sub-divide the frequency spectrum still further; though purists will reject this approach as it smacks of gimmickry, there can be a case for it for various forms of electronic musical instruments and in sound effects. One possibility is the use of parallel channels each consisting of a low-Q band-pass filter using sufficient channels that the mixed signal has very little ripple in its overall amplifier response characteristic when all controls are level. It is convenient for producing relatively small amounts of boost and cut at selected regions in the spectrum and may be augmented by active filters with higher Q if stronger effects are needed.

The mixer circuits used in such a system, as when mixing inputs from tape, disc, radio, are now frequently based on the see-saw amplifier feeding to the virtual earth through appropriately scaled resistors. If it is desired to obtain significant voltage gain from the mixer as well as having multiple inputs, the bandwidth restrictions are more severe, being in effect determined by the total gain used, i.e. the sum of the gains with respect to various inputs. Phase shift in the operational amplifier at all frequencies above 10Hz is such as to make the virtual earth point have a largely inductive impedance, i.e. one that rises proportional to frequency.

As a final comment on the possibility of multi-band operation of audio systems it can be argued that limiting the number of loudspeaker drive units to two (a woofer and a tweeter) with the occasional addition of a mid-range squawker is too restrictive. Accordingly we suggest a new classification scheme of dividing the spectrum from 10Hz to 30kHz into seven bands each to be handled by a separate loudspeaker, see table. Combined with quadrasonic operation surely an export boom must be the result?

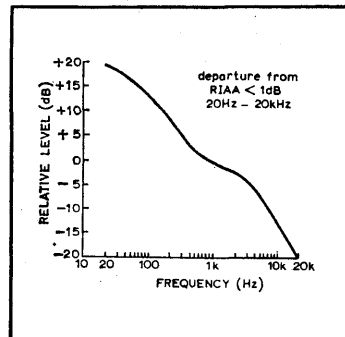


Magnetic cartridges/R.I.A.A. equalization



Typical performance

IC: 741
 R_2 : 680k Ω
 R_3 : 76k Ω (82k Ω //1M Ω)
 R_1 : 15k Ω
 R_4 : 68k Ω
 C_3 : 0.33 μ F.
 C_2 : 4.7nF
 C_1 : 4.7nF
 Supplies: \pm 10V
 Distortion: 0.05% at
 350mV r.m.s. input
 100Hz, 1kHz. 0.1% at
 10kHz



Circuit description

The required transfer function to meet the playback characteristic of BS 1928: 1965 and RIAA is of the form

$$T = K \frac{(1 + j\omega T_2)}{(1 + j\omega T_1)(1 + j\omega T_3)}$$

with $T_1 = 75$, $T_2 = 318$ and $T_3 = 3180\mu$ s. The circuit shown achieves this by using an external passive network R_1 , C_1 which gives the attenuation corresponding to $1/(1 + j\omega T_1)$ and a feedback network R_2 , R_3 , C_2 which controls the gain of the amplifier to give the remaining frequency-dependent terms. In this way three time constants are provided using only two capacitors. At very low frequencies the gain is constant at $(R_2 + R_3)/R_3$ provided the input time constant is low enough. As frequency rises X_{C_2} becomes comparable with R_2 and the gain falls, until X_{C_2} becomes comparable to R_3 . At higher frequencies amplifier gain becomes unity, with the passive network following contributing an increasing attenuation at still higher frequencies. In some cases the limited bandwidth of the amplifier might be relied on to approximate to this final high-frequency performance, eliminating R_1 and C_1 . Input impedance is determined by R_4 except at very high frequencies where amplifier input

capacitance becomes significant. Output impedance is determined by R_1 , C_1 . To obtain high-accuracy capacitors at low cost, the time constants may be obtained using resistors for R_2 , R_3 that result in a d.c. offset due to input current that conveniently cancels that due to R_4 . Lower values of resistor minimize hum pick-up problems in difficult environments.

Component changes

IC: Any general purpose op-amp whose gain-bandwidth product 20kHz.

R_2 : 6.8 to 680k Ω

R_3 : 680 Ω to 100k Ω

R_1 : 470 Ω to 47k Ω

C_3 : 0.1 to 10 μ F (may be dispensed with if head allowed to carry small input current).

C_2 : 1nF to 1 μ F

C_1 : 1nF to 1 μ F

Constraints: $R_1 C_1 = 75\mu$ s, $R_2 C_2 = 3180\mu$ s and

$R_2 R_3 C_2 / (R_2 + R_3) = 318\mu$ s.

Within these constraints the input resistance should be around 50k Ω for matching standard magnetic cartridges; the resistor values have to be compromises between drift and hum/noise problems (high R), as well as amplifier loading (low R).

Circuit modifications

● Any amplifier capable of accepting shunt-derived series-applied negative feedback may be used. Feedback is considerable and the amplifier does not need high open-loop gain to provide accurate equalization using this network. A single-ended supply may be used and the circuit is tolerant of wide variations in this supply and of transistor characteristics. The load resistance must be $\gg R_1$ if the 75- μ s time constant is not to be modified though in practice loads may be accommodated by raising R_1 and accepting some attenuation at all frequencies.

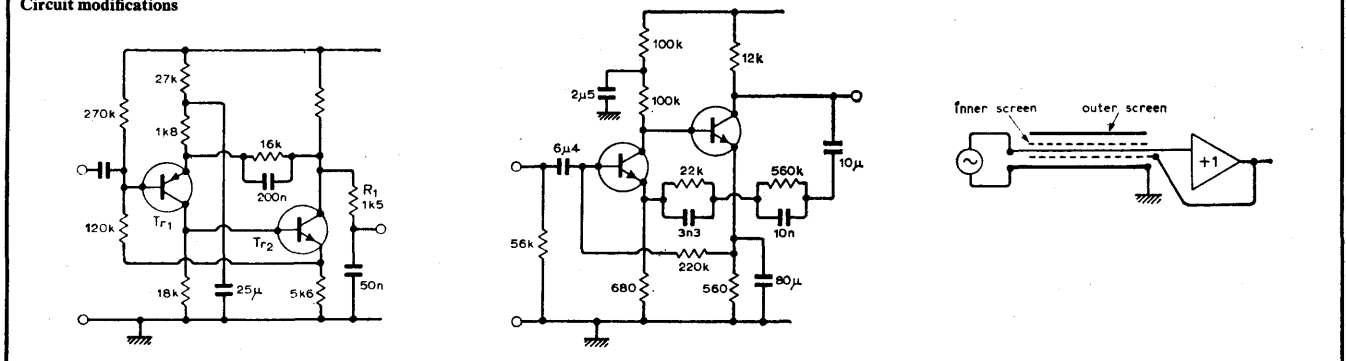
● Where the amplifier is to accommodate varying types of input shunt-applied feedback leading to a see-saw type amplifier is the usual solution; this allows control over input impedance and transfer function for a wide range of requirements. Again the gain requirements are low and one or two transistors may suffice in the amplifier.

● To minimize cable capacitance effects, twin-screened cable may be used with the outer screen grounded and the inner screen connected to the feedback path. By this means inner-screen capacitance is bootstrapped. Applicable to all series applied feedback circuits.

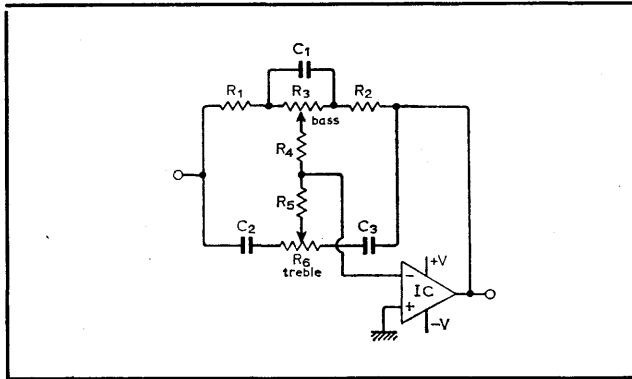
Cross references

Series 5, cards 4, 8 & 12.

Circuit modifications

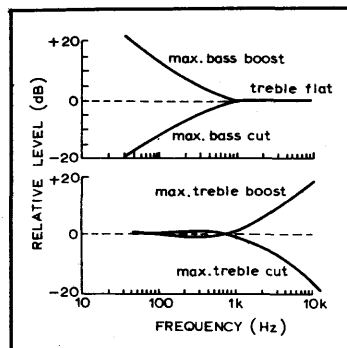


Tone control circuits/Baxandall



Typical performance

IC: 741
 Supplies: $\pm 15V$
 $R_1, R_2: 4.7k\Omega \pm 5\%$
 $R_3, R_6: 100k\Omega$
 $R_4: 39k\Omega$
 $R_5: 5.6k\Omega$
 $C_1: 47nF \pm 5\%$
 $C_2, C_3: 2.2nF \pm 5\%$
 Input signal: 2V pk-pk
 Source impedance: 60Ω



Circuit description

A tone control is a variable filter in which one or more of the elements is variable and which allows amplitude-frequency response of an amplifier to be adjusted. The above active circuit operates with a frequency-dependent feedback network and is based on the original Baxandall design. It has its greatest effect on the extreme bass and treble parts of the audio spectrum, and allows for separate bass and treble controls between which there is low interaction. The circuit features low distortion with maximum boost, and with the controls in mid-position the overall response is flat to within 1dB over the audio range. To ensure minimum restriction on the range of control available the source impedance of the driving circuit should be low. The component values used give the characteristics shown in the graphs with approximately 20dB of bass and treble boost and cut at 30Hz and 15kHz with respect to 1kHz, where the gain is unity. Excluding hum,

the total harmonic distortion and noise is better than 0.5% over the range 100Hz to 10kHz, and better than 0.01% for input signals less than 100mV.

Component changes

To make small changes in overall response of the tone control circuitry, resistors R_4 and R_5 may be increased to double their present value, or reduced to zero, e.g. for R_4 zero bass boost and cut increased by approximately 3dB for frequencies below 1kHz; for R_5 zero, treble boost and cut increased by 2dB for frequencies greater than 1kHz.

Circuit modifications

● To obtain gain with low input signals, connect feedback network to potential divider tapping across the output as shown centre i.e. gain = $(R_7 + R_8)/R_8$. Useful values for a gain of 10: $R_8: 470\Omega$, $R_7: 4.7k\Omega$. To ensure that the input and feedback parts of the network give correct balance with the controls in mid-position, the source resistance should be equal to R_7 and R_8 in parallel.

● For a discrete-component output stage, use a field-effect transistor in preference to a bipolar transistor because of better linearity and lower noise level (see last reference), and the high input impedance minimizes loading on the tone control circuit. Emitter-follower output provides a low output impedance and thus the arrangement can replace the op amp.

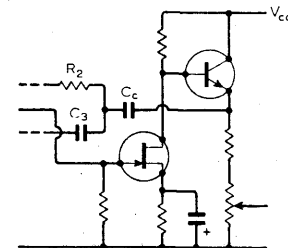
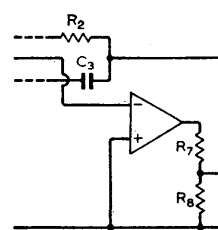
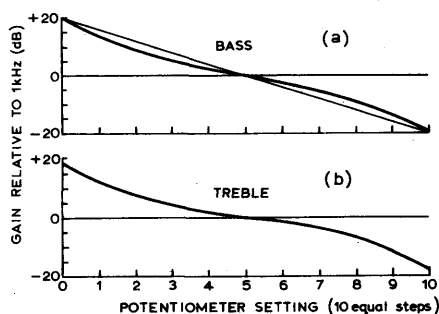
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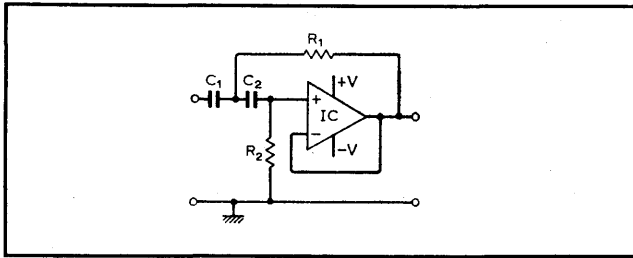
Cross references

Series 5, card 6.

Circuit modifications

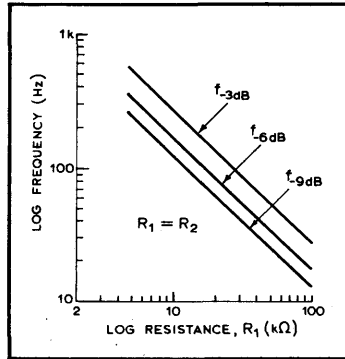


Rumble filters



Typical performance

IC: 741
 Supplies: $\pm 15\text{V}$
 C_1, C_2 : $100\text{nF} \pm 10\%$
 R_1, R_2 : see graphs
 Input: 100mV r.m.s.
 Source resistance: 60Ω



Introduction

An audio amplifier, having an amplitude-frequency response extending into the sub-audible range, can suffer from the reproduction of sub-audio-frequency signals especially when it is operated in its high output level region. Although these very low frequencies are themselves inaudible they can overload various parts of the system and/or produce inter-modulation distortion and even audible tones by mixing with wanted signal components. These rumble signals may be produced from many sources such as turntable units, tape decks and microphones, in the latter case being largely due to sudden changes in breadth or due to wind when outdoors. In disc reproduction, rumble signals are generated due to the transmission of mechanical vibrations from the turntable unit to the stylus or even along the pick-up arm to the cartridge. Sometimes these rumble signals are of large amplitude due to resonances in the turntable mounting plate or plinth or due to transmission of the sub-audible slip frequency of an induction-type drive motor. To reduce the effect of these unwanted signals many amplifiers include a high-pass, or rumble, filter which can be of active or passive form. Often the filter is a fixed-response, passive design permanently connected in the pre-amplifier and consists of high-cost inductors having high-permeability screens and capacitors. A cheaper active filter can normally be designed using transistors or integrated circuits and located between the preamplifier equalization network and tone controls. Such filters can be switched out of circuit if desired and may

have a variable response. The circuit shows a simple, second-order active filter having a cut-off frequency that is easily adjusted by means of R_1 and R_2 which could be in the form of a ganged potentiometer.

Component changes

Useful range of supplies: ± 3 to $\pm 18\text{V}$. For a required cut-off frequency C_1 and C_2 could be made 1% components and R_1 and R_2 matched to give the same time constants. Ratio C_1/C_2 can be changed to alter response. Operational amplifier 741 can be changed to a 748 or 301 with a 30-pF compensation capacitor. Resistors R_1 and R_2 may be made continuously variable with a ganged log-law potentiometer.

Circuit modifications

It is possible to cascade a number of high-pass sections, of the type shown below, to obtain filters having a higher rate of cut-off, as the output impedance of each section is very low.

● The network shown left is a modification of the basic section requiring the addition of C_3, R_3, R_4 and A_2 . Components C_3 and R_3 form an additional first-order section and R_4 is connected between the output of this section and the original second-order section. Amplifier A_2 is connected as a follower having a high input impedance so that the filter response is not significantly altered by the loading of the following stage. When $R_x = R_4$, V_{out} is in the form of a first-order high-pass filter response. When $R_x = 0$, V_{out} follows the signal at the output of A_1 which is in the form of a third-order response as the passive first-order section C_3, R_3 is then in cascade with the active second-order section; R_4 is then, ideally simply a high resistance shunt path. As R_x changes from R_4 to zero, the cut-off frequency and the rate of cut-off increase, so that it could be useful to choose the first-order and second-order section time constants to define the end-points of the response. Graph shows the variation obtained with a 741 as A_1, A_2 ; supplies of $\pm 15\text{V}$; C_1, C_2 : 100nF ; R_1, R_2 : $33\text{k}\Omega$; C_3 : $10\mu\text{F}$; R_3 : 330Ω and R_4 : $100\text{k}\Omega$.

● Circuit right shows an alternative high-pass section having a Bessel response and a cut-off frequency of $0.7/2\pi C_4 R_5$ when $C_4 = C_5 = C_6$ and $R_6 = 3R_5$.

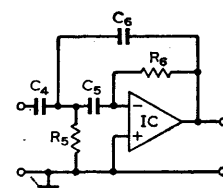
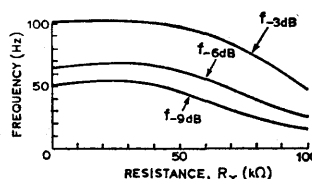
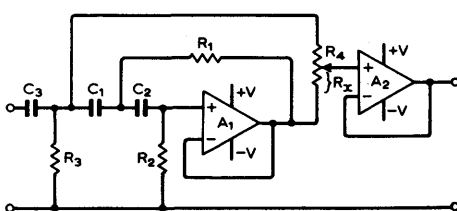
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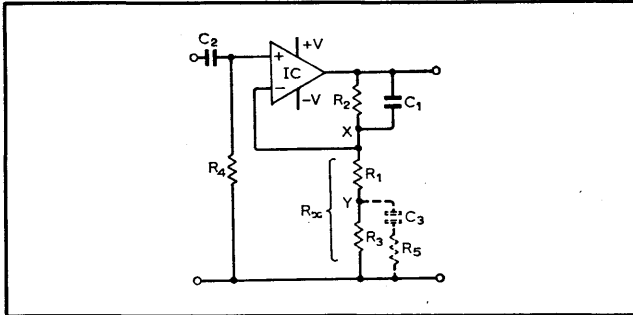
Cross references

Series 5, card 6.
 Series 1, cards 3, 5 & 6.

Circuit modifications

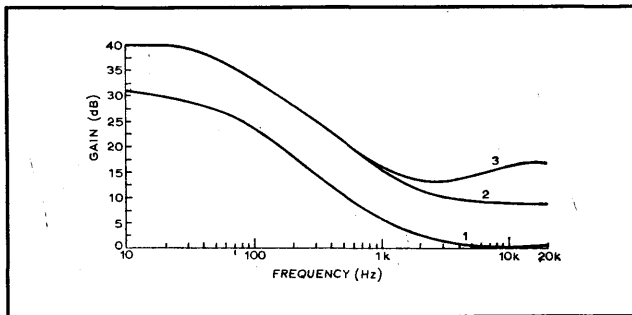


Tape-head preamplifier



Typical performance

IC: 741* values suitable
Supplies: ± 5 to ± 15 V for 9.5cm/s
 R_1 : 220 Ω ; R_2 : 10k Ω R_4 : 100k Ω ; R_5 : 22 Ω
(9.5k Ω preferred) C_2 : 47nF; C_3 : 1 μ F
 R_3 : 47 Ω *not optimum for low
 C_1 : 0.33 μ F $\pm 1\%$ noise performance



Circuit description

The circuit is closely related to that of the magnetic pickup preamplifier having a transfer function of the form $k(1 + j\omega T_1)/(1 + j\omega T_2)$.

Time constants are determined in accordance with appropriate standards. The circuit shown achieves two time-constants using a single capacitor with $T_1 = C_1(R_x$ in parallel with $R_2)$, $T_2 = C_1 R_2$. For widely-spaced time constants $R_2 \gg R_x$ and $T_1 \approx C_1 R_x$ i.e. independent control of the time-constants is possible in practice by varying R_x and R_2 separately. At low frequencies the gain rises to a maximum of $(R_2/R_x) + 1$ falling at high frequencies to unity. This fails to make use of the high open-loop gain of the amplifier, providing equalization without further amplification. The circuit also lacks the facility for introducing high-frequency lift to help overcome the practical imperfections of head and recording medium (see modifications over). Input resistance of the circuit should be high and this could be accomplished by direct connection of the tape-head to the non-inverting input. This would however allow the amplifier input current to flow in the head, partially magnetizing it and increasing the resulting tape-noise. This effect may be avoided by using C_2 (low-leakage) and by regularly demagnetizing tape-heads.

The head characteristic is such that a tape recorded with constant magnetization, irrespective of frequency, would produce an e.m.f. from the playback head proportional to frequency, as e.m.f. is proportional to the rate-of-change of magnetic field. This explains the need for the rising gain at low frequencies. A complex pattern depending on record bias

levels, high-frequency fall-off in remanence, play-back gap dimensions, record and playback gap alignments, magnetic domain size in the tape determine the number and extent of corrective actions during the record-replay process. Limiting the gain to a constant value at high frequencies is covered by the equalization standard, other actions, particularly treble boost on record and playback being considered for each specific head, tape and bias combination.

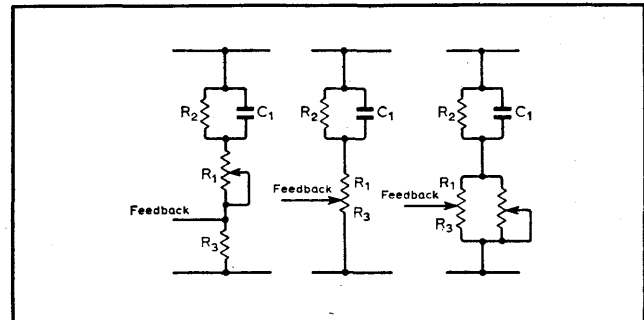
Component changes

- Improved performance obtained by using specially designed low-noise amplifiers such as the National Semiconductor LM380/381 or the RCA quad-amplifier packages CA3048 or CA3052.
- Increase gain by connecting feedback from junction of R_1 and R_3 .
- Approximately 6dB of treble boost achieved when C_3/R_5 network inserted, the upper limitation being determined by R_5 .

Circuit modifications

These are concerned with the higher frequency range where $X_{C_1} \ll R_2$.

- Network shown left is useful for variable tape speeds. If tape speed is reduced, time constant must be increased. With a lower tape speed, the output is reduced, but this will be matched by the increased gain obtained by increasing R_1 i.e. gain given by $(R_1 + R_3)/R_3$.
- Variable gain without alteration of time constant is available from circuit shown centre.
- Circuit on right shows an arrangement in which both a variable gain and time constant are independently achieved.



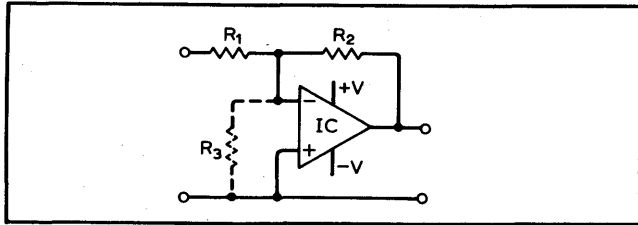
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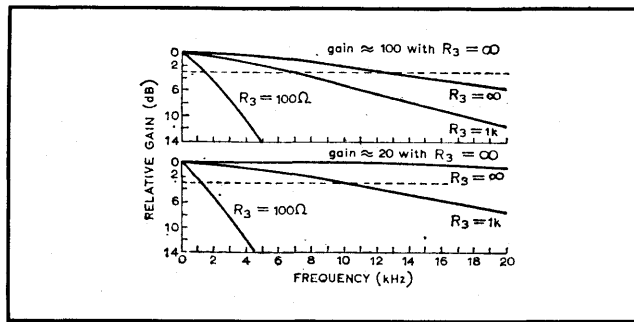
Series 5, card 1.

Audio mixer/Operational amplifier



Typical performance

IC: 741
 Supplies: $\pm 15\text{V}$
 Input: 200mV pk-pk
 Upper graph: $R_1: 1\text{k}\Omega$;
 $R_2: 100\text{k}\Omega \pm 5\%$
 $R_3: \infty; 1\text{k}\Omega, 100\Omega \pm 5\%$
 Lower graph R_1 :
 $4.75\text{k}\Omega$; $R_2: 100\text{k}\Omega \pm 5\%$
 $R_3: \infty, 1\text{k}\Omega,$
 $100\Omega \pm 5\%$



Circuit description

This is a conventional summing amplifier using a 741 op-amp, with the gain defined by the ratio of the feedback resistor R_2 , to the resistance R_1 connected to the inverting input of the amplifier, when a single source is connected. Open-loop gain-bandwidth product of the 741 is 1MHz, and if the aim is an amplifier with a gain of 100, bandwidth is then 10kHz. With multiple inputs the presence of the source resistance of these inputs connected to the nominal virtual earth point means that at high frequencies the signal is further attenuated and the bandwidth may be significantly less than 10kHz. The source resistances are effectively connected in parallel to ground and, for example, for ten 1-k Ω inputs, the effective resistance to ground is 100 Ω . Hence with a feedback resistor of 100k Ω , the feedback factor is 1/1000, and from a feedback standpoint, the system corresponds to one for which the gain is 1000, and gain-bandwidth product is reduced to 1kHz. If a 3dB cut-off of 50kHz is desired to allow a safe margin, then this could be achieved for a gain of 20. If it is fed from, say, ten inputs simultaneously, then the allowed gain for each input would be 1/10 of 20, on the basis of the previous argument that it is the

sum of the parallel admittances which determines the gain-frequency characteristic of the amplifier, and either a gain of 20 for a single amplifier, or ten separate gains of 2 can be achieved from the mixer. Thus an overall unity-gain mixer is a not-unreasonable circuit. Although it appears to waste bandwidth at first sight, it allows in practice, with ten inputs, a bandwidth approaching 100kHz.

Analysis

This shows why a large bandwidth cannot be achieved in a mixer which is simultaneously fed from a large number of inputs. The mixer may be in the form below/left, where there are N inputs and the value of the feedback resistor is kR . As far as feedback is concerned, the circuit reduces to that shown centre and the magnitude of the effective gain is $|G| = kR/(R/N) = Nk$.

For the 741, $Nk \times \text{bandwidth} = 1\text{MHz}$. If the bandwidth is 50kHz, Nk must be 20, and if N is 10, k is 2.

Component changes

Effects of parallel source resistance on magnitude and phase at 1kHz and 10kHz for nominal gain magnitudes of 100 and unity ($R_3 = \infty$) are tabled below $R_2 = 100\text{k}\Omega$, $R_1 = 1\text{k}$ or 100k Ω .

gain		frequency (kHz)	R_3 (k Ω)	phase lag (deg)
nominal	measured			
100	95	1	∞	-180
100	70	10	∞	-240
100	90	1	1	-200
100	45	10	1	-260
1	unity	1	∞	-180
1	within	10	∞	-180
1	resistor	1	1	-180
1	tolerances	10	1	-215

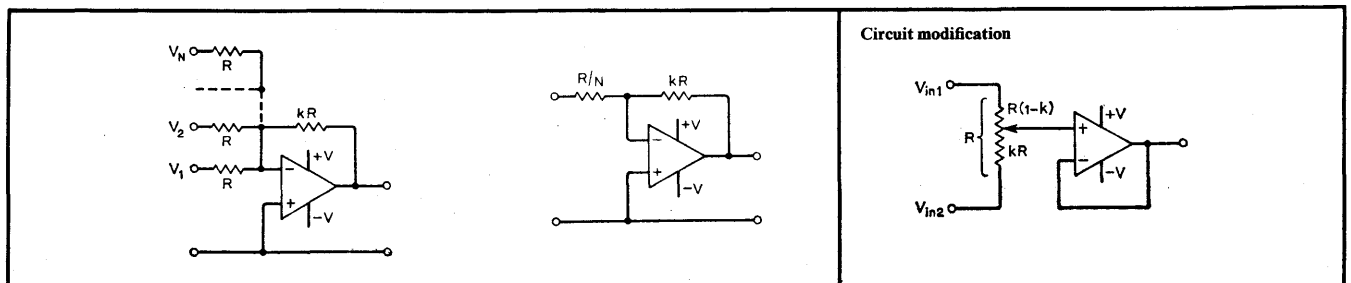
In the last two observations for unit gain, the 1k Ω shunt resistance causes a d.c. shift of -250mV.

Circuit modification

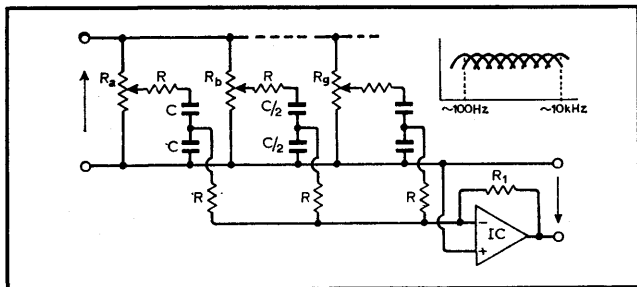
To mix two signals, use voltage follower as shown below. Output voltage is $V_{\text{out}} = kV_{\text{in1}} + (1-k)V_{\text{in2}}$. An advantage is that the degree of mixing is continuously variable, with equal mixing when $k = 0.5$. However, $R \gg$ output impedance of the signal sources, and suitable drivers would be op-amps with $R > 10\text{k}\Omega$.

Further reading

Evans, J. H. & Williams, P., Modular integrated circuit audio mixer, *Wireless World*, vol. 78, Dec. 1972, pp.564-70. Audio mixing preamplifier circuit, *Mullard Outlook*, vol. 15, no. 1. June 1965.



Multi-section tone control system



Typical performance

IC: 741

Supplies: $\pm 15V$ R_a to R_n : 10k

R: 12k

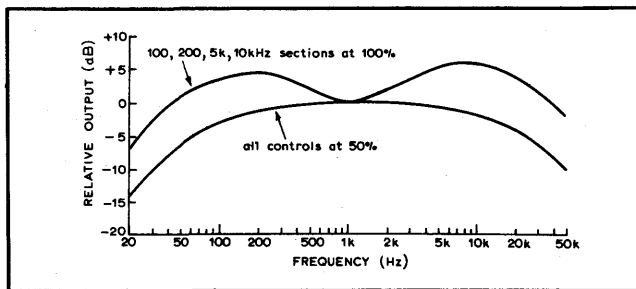
C: 100nF, 50nF, 20nF
10nF, 5nF, 2nF, 1nF R_1 : 1k Ω n : 7 stagesFor $R_a - R_g$ set to max.

output +0dB/-3dB

from 120Hz to 15kHz.

centre frequency: $1/2\pi CR$.

C.R.



Circuit description

To modify the amplitude-frequency characteristic of an amplifier over limited parts of the spectrum, filters may be used centred on the appropriate frequencies. Variations on this method are unlimited, and the circuit shown represents one possibility, in which the output is proportional to the sum of the outputs of a number of passive filters. Centre frequencies of the passive filters are scaled in 1, 2, 5, 10 steps from 100Hz to 10kHz, and the Q is low enough that for equal inputs the ripple in the response curve is < 1 dB with -3 dB points below 100Hz and above 10kHz. The summing amplifier operates on the currents flowing in the resistors forming part of what would normally be the parallel arms of the Wien networks. Raising or lowering the contribution of each network singly or in groups by means of the input potentiometers allows for moderate amounts of cut or boost in particular parts of the spectrum. If larger boosts are required then the passive networks can be replaced by active filters (see

Circard series 1). The number of sections used must depend on particular requirements; even a single section feeding the summing junction together with direct input via a resistor allows for selective boost or cut at a given frequency. The larger the number of sections the higher the Q that may be used for a given ripple in the frequency response. The C, R values chosen would result in unloaded centre frequencies greater than the nominal 100, 200Hz etc. Using relatively high values for R_a to R_g lowered the frequencies closer to nominal for potentiometer settings close to centre positions.

Component changes

Potentiometer resistances should be low to avoid changing characteristics of Wien networks (output resistance of pot. included in series arm of network). A compromise is necessary as multiple pots. present heavy load to the audio source - range 1 to 100k Ω within above constraints.

C: 100 pF to 10 μ F chosen to give centre frequencies of $1/2\pi CR$.R: 1k to 1M Ω R_1 : 1 to 100k Ω

Any number of sections may be used in principle with the restriction that the frequency limitations of the summing amplifier dictate the high-frequency response.

Circuit modifications

● Any band-pass amplifier may be added to the system, or replace an existing passive section, to increase the boost attainable at a given frequency. Such filters as those of Circard series 1 nos 1, 3, 6 & 8 may be used. As shown, the output is an inverted version of the input at the centre frequency and adjustment of the Q by varying the tapping point results in complex variations in the response, ranging from band-pass to a virtual notch. These methods must be applied with discretion but as the accompanying graph shows even a single active filter can produce dramatic variations in response (applications could include sound-effects, tone forming in electronic musical instruments).

● Combinations of low and high-pass circuits allow the raising and lowering of ranges of frequencies without the peaky response of band-pass circuits as above. Notch filters inserted in series can be used to tune out particular frequencies.

Further reading

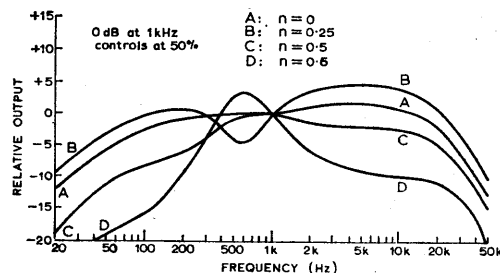
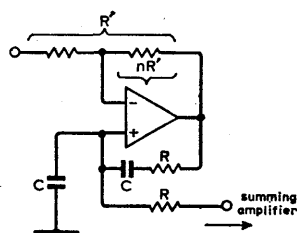
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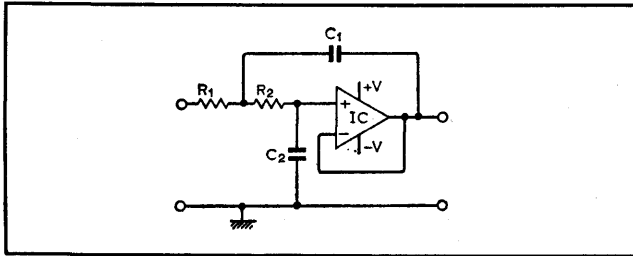
Series 5, cards 2 & 10.

Series 1, cards 1, 3, 8 & 11.

Circuit modifications

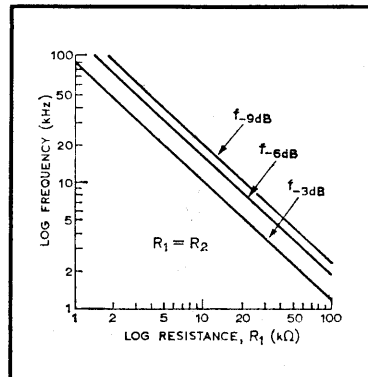


Scratch filters



Typical performance

IC: 741
 Supplies: $\pm 15V$
 $C_1, C_2: 1nF \pm 1\%$
 R_1, R_2 : see graphs
 opposite
 Input: 100mV r.m.s.
 Source resistance: 60Ω



Circuit description

Many audio amplifiers provide low-pass or top-cut filters to define the upper limit of the frequency response. These filters, normally incorporated between the source equalization pre-amplifier and the tone control circuit, are used to eliminate unwanted signals such as tape hiss, surface noise and scratches from old discs, radio interference, bias oscillator pick-up. Although some undesired h.f. signals are outside the audible range they can nevertheless overload the amplifier or introduce inter-modulation distortion components that may render the output unsatisfactory unless they are attenuated by a low-pass filter. Such filters may be passive or active networks or a combination of both types. Whatever its form, the low-pass filter will normally be switched in or out of circuit as desired. Cut-off frequency of the filter may either be infinitely-variable over a wide range or selected by a multi-position switch, typical values being 4, 8, 10, 12 and 15kHz. Selection of the cut-off frequency will normally be made subjectively depending on the amount and nature of the high frequency noise. While the rate of cut-off could also be chosen subjectively it would not normally exceed about 18dB/octave due to the increasing likelihood of severe transient distortion or ringing in the region of the cut-off frequency as a result of the amplifier becoming conditionally stable.

The circuit shown is an example of a simply-designed, second-order active low-pass filter using RC networks that can provide a wide range of cut-off frequencies with a cut-off rate that is well within the above-stated maximum. By

suitable choice of components its input impedance can be adjusted so that it does not significantly load the preceding pre-amplifier and its output impedance is low due to the operational amplifier.

Component changes

Useful range of supplies: ± 3 to $\pm 18V$.

For defined cut-off frequency C_1 and C_2 may be changed with corresponding change in R_1 and R_2 to give same time constants.

Ratio C_1/C_2 may be varied to change response.

741 operational amplifier may be replaced by a 748 or 301 using a 30-pF compensation capacitor.

R_1 and R_2 may either be switched with a two-pole unit or made infinitely variable with a ganged log-law potentiometer.

Circuit modifications

● Low-pass filters capable of providing a wide range of cut-off frequencies and a variable rate of cut-off tend to become complex networks. Circuit on left shows a modification of the basic circuit which provides a compromise between network complexity, variable cut-off frequency and rate of cut-off. Components A_1, R_1, R_2, C_1 and C_2 form the basic second-order active filter and R_3 and C_3 form an additional first-order passive section. Potentiometer R_4 is connected between the outputs of the first-order and second-order networks. When $R_x = R_4$, V_{out} provides a first-order response cascaded first-order and second-order networks. R_4 must be reasonably large to prevent significant changes in the time constants at the extreme values of R_x and A_2 serves as a high-input-impedance buffer to avoid excessive loading of the network. Between the extreme settings of R_x the filter provides a variable cut-off frequency and a variable rate of cut which increases as the cut-off frequency falls. The graph shows this effect with $A_1, A_2: 741$; supplies of $\pm 15V$; $C_1, C_2: 1nF$; $R_1, R_2: 10k\Omega$; $C_3: 10nF$; $R_3: 1k\Omega$ and $R_4: 100k\Omega$.

● Of the many alternative forms of active low-pass filters one example is shown right which provides a Bessel response – a good compromise between sharpness of cut off and transient overshoot – when $R_5 = R_6 = R_7$ and $C_4 = 3C_5$ with a cut-off frequency of $1.4/2\pi C_4 R_5$ Hz.

Further reading

Evans, J. H. and Williams, P., Modular integrated circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.

Gayford, M. L., Hi-Fi for the Enthusiast, Pitman, 1971.

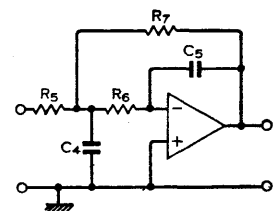
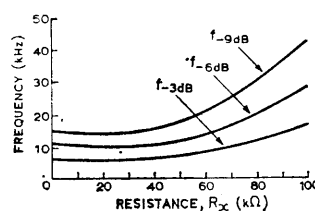
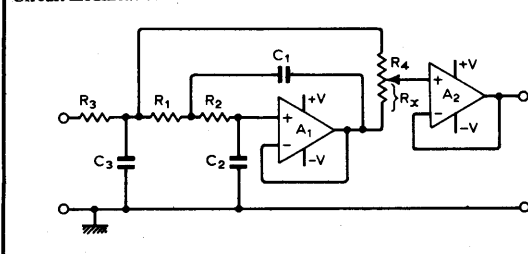
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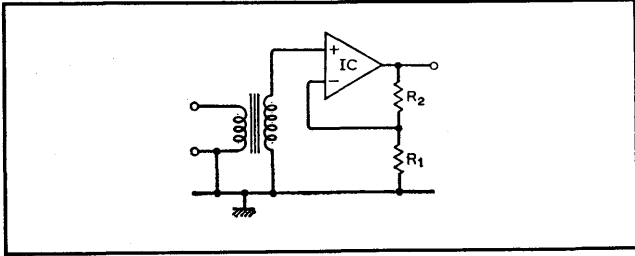
Series 5, card 6.

Series 1, cards 3, 4, 6 & 11.

Circuit modifications



Microphone preamplifiers



Typical performance

IC: 741
 Transformer: see note
 R_2 : $1k\Omega$
 R_1 : 100Ω
 Source resistance: 200Ω
 Response: 45Hz to 20kHz, +0dB/-3dB
 (Application of small bass-boost as in card 2 gave 20Hz to 20kHz ± 1.5 dB)
 Equivalent input noise: $0.5\mu\text{V}$ r.m.s. 20Hz to 20kHz
 Noise reduced by 2dB by reducing bandwidth to 10kHz.

Note:

Performance of the microphone amplifier is strongly affected by the transformer. Type used was U204 by AKG, having a 15:1 turns ratio, a secondary inductance high enough to avoid l.f. limitation, and sufficiently low self-capacitance to prevent h.f. losses. Microphones may have specific transformers designed for them and the characteristics of the two should be considered together.

Circuit description

High-quality moving-coil microphones have a low output e.m.f. and a low internal resistance ($< 1\text{mV}$ and $100\text{-}500\Omega$). Coupled directly to any semiconductor amplifier whether discrete or integrated, the resulting signal/noise ratio would be unacceptable as "hi-fi" but might be adequate for some applications (s/n ratios of about 45-50dB being the limit attainable with low-cost i.cs and low-sensitivity microphones). As the microphone has a relatively low resistance the dominant noise effect is due to the equivalent noise voltage at the amplifier input—the noise current fails to develop any significant p.d. across the microphone resistance. By transformer-coupling the signal to the amplifier input, and using a step-up ratio of 10:1 or greater, the signal e.m.f. can be greatly increased; no change in noise voltage generated at the amplifier input occurs but the noise current produces a proportionately larger contribution because the effective source impedance seen by the amplifier input is $n^2 R_s$ where n is the step-up turns ratio and R_s the microphone resistance. The optimum turns ratio which results in equal contributions from voltage and current noise generators may not give the optimum from a standpoint of amplitude response; with too

high a turns ratio the resulting high impedance may be shunted heavily by capacitance at high frequencies, while the secondary inductance limits the low-frequency response.

Component changes

The amplifier gain bandwidth product is 10^6Hz and to leave a safe margin for amplifier limitations of upper cut-off, gains of < 50 should be accepted. The transformer has a turns ratio (e.g. 5 to 50) and the amplifier output may be up to $500 \times$ the microphone e.m.f. though 100 to 200 is more likely. Keep R_1, R_2 low from low-noise standpoint e.g.

R_1 : 10 to 500Ω .

R_2 : 100 to $5k\Omega$.

R_2/R_1 : 5 to 50.

Coupling capacitor may be used at output though d.c. content is small.

Output stage of op amp is class B. Where crossover distortion is a worry, even though negative feedback is large, stage may be biased into class A by taking resistor from output to one supply line, drawing current greater than expected peak load current.

Circuit modifications

● Any non-inverting amplifier may use series-applied feedback in the same way. If conventional transistor amplifiers are used, one approach is to define the d.c. conditions separately and capacitively couple the transformer output to the amplifier. The method shown left uses the low d.c. resistance of the transformer to act as a shunt-applied d.c. negative feedback from second emitter to first base. If R_4/R_5 is relatively large (high closed-loop gain) then p.d. across R_3 is V_{be} of Tr_1 , $\approx 0.6\text{V}$ for silicon. Convenient resistors might have $R_2 = 5R_3$, $R_1 = 10R_2$, $R_2 = 10R_4$, $R_4/R_5 =$ required gain, $0.6\text{V}/R_3 =$ standing current in Tr_2 ; all for $V_s = 5$ to 15V . The values may be chosen to give a standing current in Tr_3 convenient for the required load swing, or if that is not a limiting factor, the currents may be lowered to raise the effective input impedance. The ratios are neither critical nor optimal, but are working guides.

● Using the virtual earth idea, an extremely simple self-biased stage gives no voltage gain but a lower output impedance at negligible cost. It may be similarly applied to the d.c. feedback-pair circuit or as shown to an alternative operational amplifier arrangement.

Cross References

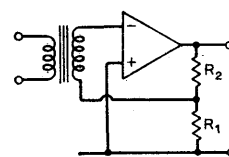
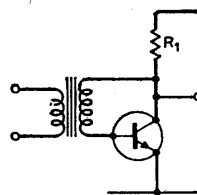
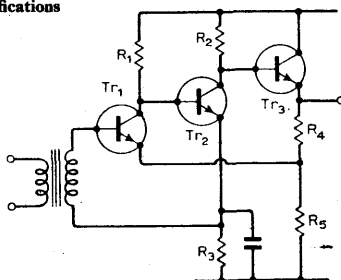
Series 5, cards 2 & 12.

Further reading

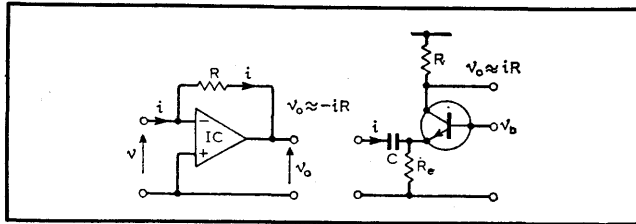
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Walker, H. P., Low-noise Audio Amplifiers, *Wireless World*, vol. 78, May 1972, pp.233-7.

Circuit modifications

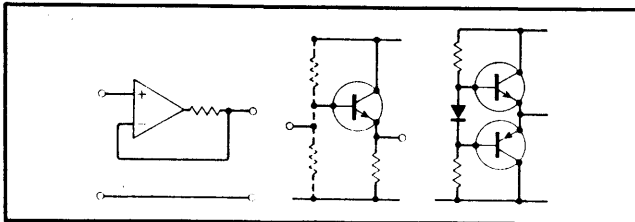


Impedance matching and transforming



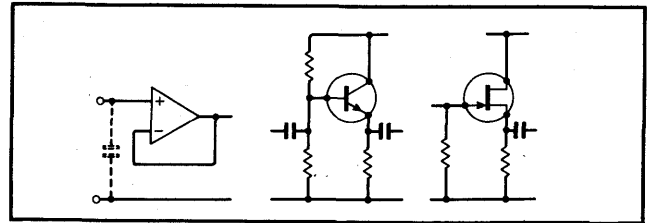
Circuit description

Shunt-applied negative feedback reduces the input resistance of an amplifier. To a first-order, for A large the input impedance is R/A , commonly called Miller effect, properly due to Blumlein. As A is frequently dependent, and for amplifiers such as 741 has 90° phase lag for $10\text{Hz} < f < 1\text{MHz}$, the feedback current lags on the applied voltage i.e. input impedance is low, predominantly inductive and is hence proportional to frequency. The assumption that the input is a virtual earth is justifiable in most cases, but may not be so at upper end of audio range ($Z_{in} \sim j1000\Omega$ at $f = 10\text{kHz}$ for $R = 100\text{k}\Omega$). If d.c. conditions not important, common-base amplifier has low input resistance ($\sim 25\Omega$ for $I_c = 1\text{mA}$) and excellent frequency response. Lower cut-off frequency determined by C and source resistance. Output may be taken at collector either as current into external load (note quiescent d.c. current) or as voltage developed across R . Values of V_b , R_e determine quiescent conditions. Input resistance non-linear and input current should be restricted to small fraction of standing current. Input resistance $\propto 1/I_c$.



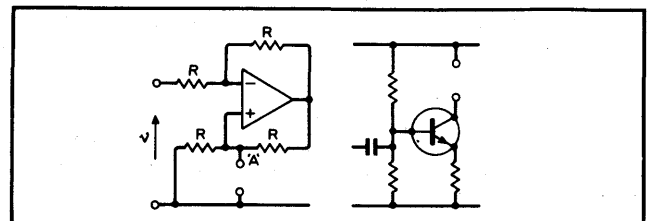
Circuit description

Amplifiers such as the 741, 301 etc. have open-loop output resistances of 50 to 150Ω , relatively frequency independent up to the useful limit of operating frequency. Falling gain of the amplifier at high frequencies does not allow the feedback to reduce the effective output impedance to almost zero as it does at low frequencies. Output impedance is largely inductive and corresponds to a few tens of microhenries e.g. at 90kHz series resonance effects can be observed with a load capacitance of 100nF . Effects are minimal at audio frequencies except where high gain is being attempted i.e. feedback much less than unity. This property is distinct from the current limiting of the amplifier into too low a load resistance, where additional internal transistors clamp the output current to $\sim 25\text{mA}$. Again emitter followers, Darlington pairs may be used. For high output currents complementary emitter followers in class B are used but the subject is then properly treated as a power amplifier.



Circuit description

The basic follower circuits (voltage, emitter and source follower) have a.c. voltage gains close to unity, supply appreciable load currents and ideally draw negligible current from the source i.e. giving high input impedance. At high frequencies the finite open-loop gain of the operational amplifier together with its phase lag contribute additional shunt capacitance effects at the input depending on the amplifier input resistance. In most cases the total input capacitance is likely to be dominated by the physical system capacitances; minimum device capacitances below 1 to 3pF are rare, total capacitance may be around 10pF . For high source resistance ($> 100\text{k}\Omega$) this may bring cut-off frequency below 100kHz but not serious in audio band. Most likely limitation remains falling gain of amplifier if set for high voltage gain initially. See card 5 on mixer circuits. Input impedance of discrete circuits may be high particularly f.e.t. ($> 1\text{M}\Omega$) but a.c. coupling required, at least at output.



Circuit description

There is no simple way of obtaining high output resistance to a grounded load using operational amplifiers that have one output grounded. One possibility shown above uses positive and negative feedback combined to raise the impedance at point A close to infinity. This may also be viewed as a negative impedance converter circuit and has the possibility of an output resistance which while large may have either positive or negative sign. Load resistance should not be too large (say $< 10R$ if possible) so that the negative resistance effect cannot dominate. Output current is then determined by V and R . At high frequencies the amplifier gain/phase shift gives an equivalent output impedance that becomes capacitive. For a 741 using $R = 10\text{k}\Omega$, a shunt capacitance of about 60pF was observed. Upper cut-off frequencies in excess of 100kHz are possible. The transistor circuit and its complementary version allow high impedance a.c. drive but superposed on a direct standing current.

Further reading

Evans, J. H. and Williams, P., Modular integrated circuit audio mixer, *Wireless World*, vol. 78, 1972, pp.564-70.

Series 5, cards 1, 8 & 11 high Z_{in}
 4 & 5 low Z_{in}
 5 fixed Z_{in}

Economy i.c. audio circuits

Circuit description

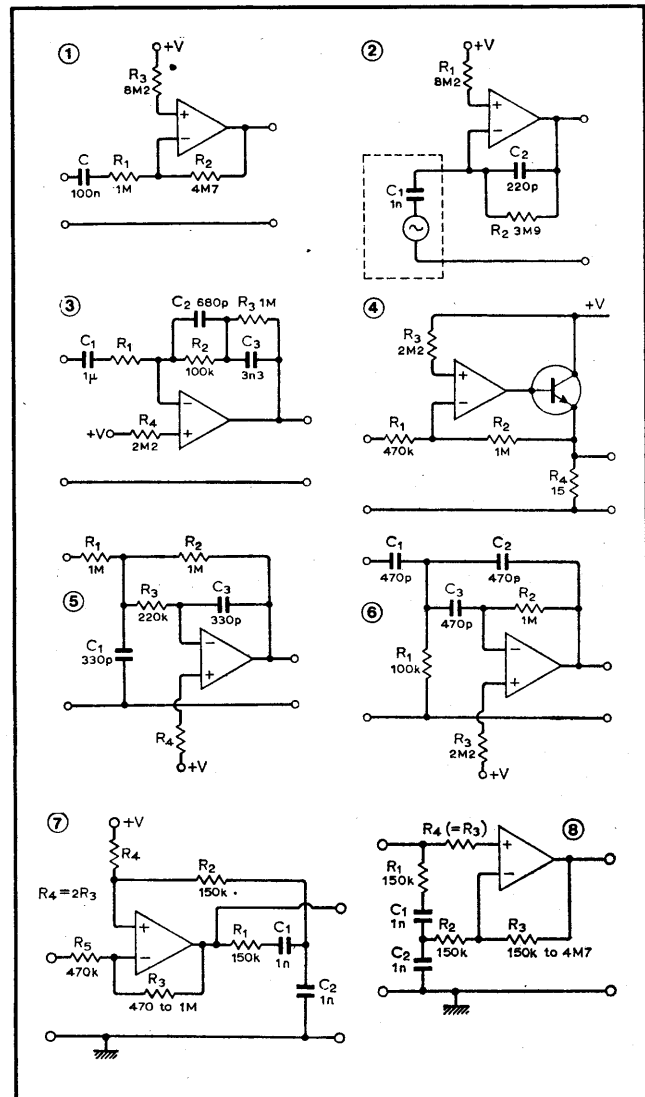
The i.c. contains four identical amplifiers each of which is similar to the d.c. feedback pair of a common-emitter stage followed by a common-collector stage. Feedback is applied externally, and the novel feature of the circuit is the second input forming part of a current mirror such that the effective current seen by the amplifier is the difference between the currents fed to the two inputs. The non-inverting input normally receives a separate bias from the positive supply or from any other convenient positive potential including the output of some preceding stage. Signal and feedback are most often applied to the inverting input with the system having properties associated with the see-saw amplifier familiar in op amp circuits. The difference is that the virtual earth is so only for a.c., having a d.c. potential equal to the input transistor V_{be} i.e. $\approx 0.6V$. This configuration allows most functions normally provided by op amps to be achieved with a single-ended supply over a wide voltage range.

Circuit 1. A basic buffer amplifier with a voltage gain of $-R_2/R_1$ and an input impedance of R_1 . If $R_3 = 2R_2$ the output voltage is approximately $V/2$. This is because the negative feedback through R_2 causes the current flowing in the inverting input to be an approximate match to operation by the d.c. at the output. Although large capacitances are then required ($> 500\mu F$) if good low-frequency response is the aim. The low voltage rating keeps capacitor cost low. Heavy feedback keeps distortion low, assisted by class A operation. Bandwidth of 10Hz to 100kHz possible with distortion $< 1\%$ in range 100Hz to 10kHz. Ripple induced hum may be minimized by centre tapping R_3 and decoupling to ground.

Circuit 2. Alternative capacitance feedback for inputs such as piezoelectric pick-ups (see card 11). Resistor R_2 necessary to provide bias, reduces low-frequency gain. Possibly centre-tap and decouple to ground using capacitance of 1nF to avoid peak in audio band.

Circuit 3. Magnetic pickup may also be accommodated using feedback network of type described on card 1. Correct bias obtained for $R_4 = 2(R_2 + R_3)$ with R_1 chosen to give correct input impedance. These amplifiers are not specified for low-noise performance but are still worth considering where economy is predominant aim. It is possible to combine the properties of 3 & 4 to give direct headphone drive with one stage from such an input, but at increased distortion and reduced bandwidth. Separate circuits for 3 & 4 give full stereo operation using a single i.c. package together with two low-cost output transistors. Major cost in such output stages must be that at the non-inverting input, and this can only be so when the output voltage is related to the supply voltage by the appropriate ratio of the resistors. An alternative bias method for an input with a d.c. resistance to ground $\ll R_1$, is to dispense with C and R_3 and choose the ratio R_2/R_1 to set the output d.c. conditions. This follows from the 0.6V at the inverting input which then defines the current in R_1 . The same current flows in R_2 as that drawn by the inverting terminal is very small. Hence $V_o(d.c.) \approx [(R_2/R_1) + 1]0.6V$. The gain option is thus restricted by the d.c. setting as the ratio R_2/R_1 is involved in both.

Circuit 4. Where higher output currents are required, a simple class A stage can be added. Output current of the amplifier is up to 10mA, allowing a load current of 100 to 200mA to be achieved without difficulty. Operating with a supply voltage of $\sim 5V$ dissipation in the output transistors is $\ll 1W$ and low-cost plastic/epoxy units are satisfactory. Current levels are such that this is a very convenient output stage for

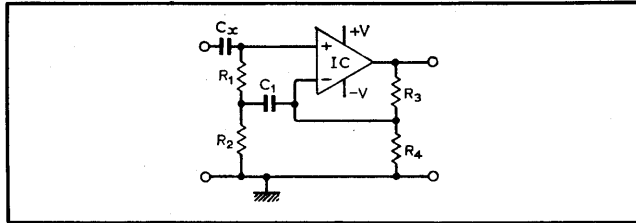


low-resistance headphones—up to quadraphonic operation using identical configurations for each of the four amplifiers in the i.c. package. Capacitance coupling to the load is then indicated since the headphones would be driven into non-linear region (power supply etc.)

Circuits 5 & 6. Rumble and scratch filters may be constructed (cards 3 and 7) but as the voltage-follower mode is not obtainable directly, different passive networks are used. Those shown are both for cut-off frequencies of around 1kHz. The damping may be adjusted without changing this frequency by varying the ratios C_2/C_1 for the scratch filter, 5, and the ratio R_2/R_1 in the rumble filter, 6. Increasing all capacitors in each circuit by a factor n reduces the cut-off frequency by the same factor.

Circuits 7 & 8. If more specialized filters are required in audio systems such as boosting response over a limited band e.g. to help overcome losses in tape heads at high frequencies, then active filters offer an alternative to LC passive sections. Circuit 7 is a low-Q bandpass filter of centre frequency $1/2\pi R_1 C_1 R_2 C_2$ while 8 represents a simple notch filter of the same frequency, adjusting R_4 to obtain the notch. R_3 sets the gain and the bias is obtained from the d.c. level of the preceding stage.

Ceramic cartridge preamplifier

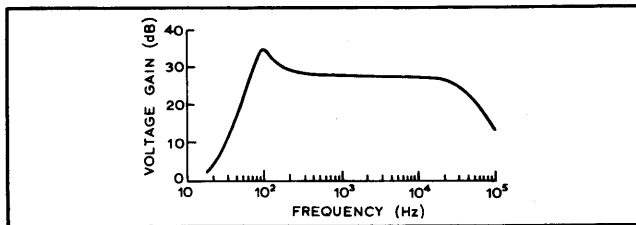


Typical performance

IC: 741
 Supplies: $\pm 15V$
 R_1, R_2 : $10k\Omega$; R_3 : $220k\Omega$; R_4 : $22k\Omega$;
 C_1 : $22\mu F$ (tantalum).
 C_x : $1.5nF$ (see circuit description)

Input: $100mV$ r.m.s.
 Source resistance: 60Ω ;
 (see curve opposite)
 Note. Components chosen to show resonant peak.

$$\text{Gain } 1 + \frac{R_3(R_2 + R_4)}{R_2R_4}$$



Circuit description

Unlike magnetic cartridges which produce an output proportional to the velocity of the stylus movement, a piezo-electric cartridge provides an output proportional to the amplitude of the stylus deflection and appears as an almost pure capacitance at the preamplifier input. The RIAA recording characteristic is essentially one of constant amplitude except for a small mid-band range of constant velocity recording. Piezo-electric cartridges are normally designed to provide mechanical equalization that caters for imperfections in the almost constant-amplitude recording characteristic. Thus, the preamplifier should provide a flat amplitude response and a high input impedance, not less than about $1M\Omega$. As the input impedance falls so also does the response at the bass end of the spectrum, leading to an approximation to an unequaled magnetic cartridge response. In the circuit shown, C_x represents the capacitance of a crystal or ceramic cartridge, normally in the range $400pF$ to $1nF$. Bootstrap capacitor C_1 allows the cartridge to be lightly loaded by the high input impedance at the non-inverting input of the operational amplifier. To achieve this C_1 must be a high-value, high-quality capacitor such as a tantalum bead type. Then, the p.d. across R_1 is essentially zero as the high open-loop gain of the operational amplifier forces its differential input p.d. to be zero. Thus, no current flows in R_1 and the cartridge sees it as an open circuit. At low frequencies, the boot-

strapping becomes imperfect and the inductive input impedance results in the resonant peak shown above. In the circuit shown below (left) $e = V/(1 + jC_1R_2)$ and $i = e/R_1$, hence $Z_{in} = V/i = R_1(1 + jC_1R_2)$ which contains an equivalent inductance $L = C_1R_1R_2$. This inductance resonates with C_x at $f_o = 1/2\pi C_1R_1R_2C_x$ ($88Hz$ for above component values). For good transient performance f_o must be well below the audio band. See below for suitable values.

Component changes

Useful range of supplies: ± 3 to $\pm 18V$.
 To make resonant peak occur below $10Hz$ with $C_1 = 22\mu F$ use $R_1 = R_2 = 180k\Omega$ for $C_x = 400pF$ and $R_1 = R_2 = 120k\Omega$ for $C_x = 1nF$.
 Ratio R_3/R_4 may be adjusted to give desired gain.
 Lower values of R_3 and R_4 make gain less dependent on C_1, R_1 and R_2 .
 Screened wire between cartridge and amplifier reduces gain and resonant peak frequency e.g. 1 metre of separately-screened pair ($136pF/m$) reduces mid-band gain by about $1.6dB$ and f_o from $88Hz$ to $78Hz$.

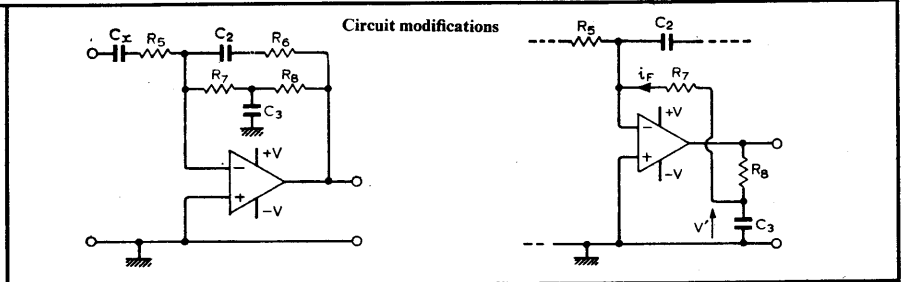
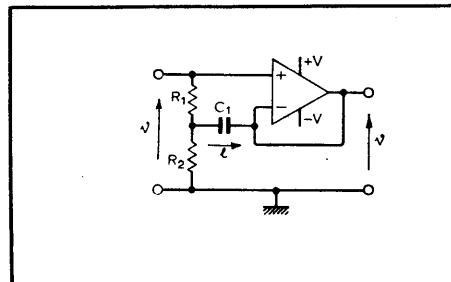
Circuit modifications

The circuit shown centre demonstrates a method of obtaining a flat amplitude response by making the time constant C_xR_5 and C_2R_6 equal, where C_x is the capacitance of the piezo-electric cartridge. With $R_6 = kR_5$ and $C_2 = C_x/k$ the gain of the amplifier is given by: $G = (kR_5 - jk/\omega C_x)/(R_5 - j/\omega C_x) = k$. Resistive loading of the cartridge is determined by R_5 which should not be less than about $1M\Omega$. The d.c. negative feedback provided by R_7 and R_8 will not significantly affect the response of the amplifier provided that R_7 and R_8 are both very much greater than R_6 and that C_3 adequately decouples them throughout the audio band. Incorrect choice of component values can again lead to a resonant peak at some low audio frequency, but as its Q is not high it could be used to give a degree of bass boost if desired.

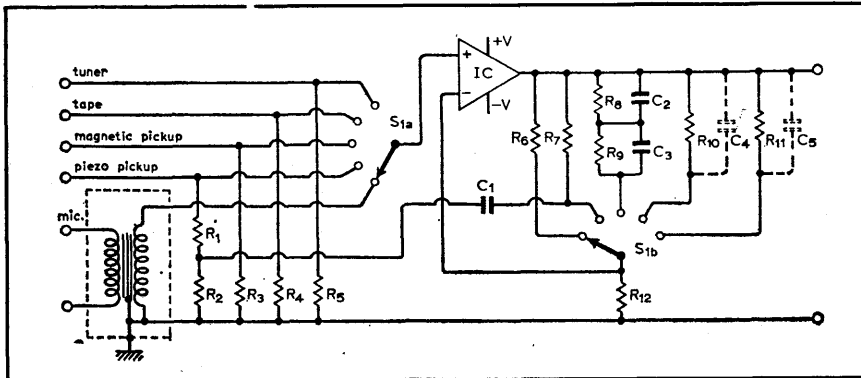
To see this effect, refer to circuit on right where $V' = V_{out}/(1 + j\omega C_3R_8)$ and $i_F = V'/R_7$. Hence the equivalent impedance (Z_{eq}) of the d.c. feedback network is given by $Z_{eq} = V_{out}/i_F = R_7(1 + j\omega C_3R_8)$. Thus, Z_{eq} contains an equivalent inductance $L' = C_3R_7R_8$. This inductance will resonate with C_2 (with which it is in parallel) at a frequency $f'_o = 1/2\pi C_2C_3R_7R_8$ Hz. To demonstrate this effect, component values were chosen as: A: 741; C_x : $1.5nF$; R_5 : $1k\Omega$; R_6 : $1.5k\Omega$; C_2 : $1nF$; R_7, R_8 : $10k\Omega$; C_3 : $22\mu F$ and the resonant peak found to occur at $108Hz$ compared with a predicted f'_o of $107Hz$.

Further reading

- Evans, J. H. and Williams, P., Modular integrated circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.
- Burrows, B. J. C., Ceramic pickups and transistor pre-amplifiers, *Wireless World*, 1970. pp.56-60
- Earl, J., How to Choose Pickups and Loudspeakers, Fountain Press, 1971.
- A preamplifier for use with crystal and ceramic pickups, Design Note 14, SGS-Fairchild, 1965.



Multi-input preamplifier



Typical performance

IC: 741
 Supplies: $\pm 15V$
 Transformer: AKG 0204
 $R_1, R_2: 150k\Omega; R_3: 47k\Omega$
 $R_4, R_5: 470k\Omega;$
 $R_6: 8.2k\Omega; R_7: 120\Omega$
 $R_8: 560k\Omega; R_9: 47k\Omega$
 $R_{10}: 120\Omega; R_{11}: 1.2k\Omega$
 $R_{12}: 820\Omega$
 $C_1: 22\mu F$ tantalum
 $C_2: 5.6nF; C_3: 1.8\mu F$
 Sensitivities:

For 350mV
 output at
 1kHz
 mic: 3.5mV
 piezo: 308mV
 mag: 4.82mV
 tape: 310mV
 tuner: 141mV

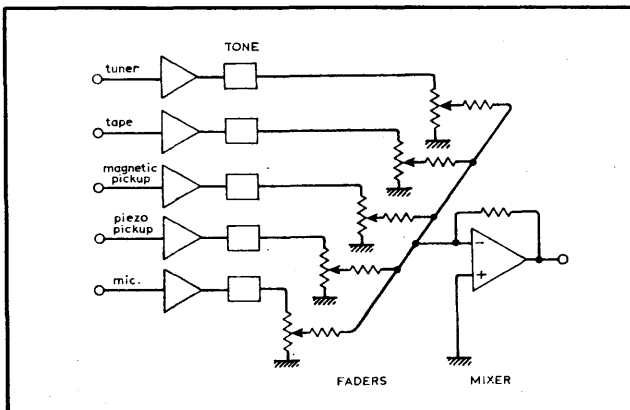
Circuit description

When designing a complete preamplifier circuit to accommodate a large range of signals from different sources, the output voltage to be fed to the main amplifier must normally be of the same value irrespective of which source is being used. The preamplifier must therefore provide different sensitivities for different sources as well as providing any equalization appropriate to a particular input. Many preamplifier circuits therefore use a single amplifier with passive input and feedback networks switched into circuit to meet the requirements of a given source. The circuit shown is of this type and uses a single integrated-circuit operational amplifier as the active block although a discrete transistor version could also be used. The operational amplifier is connected as a voltage follower with gain where the gain is defined by the feedback network components. Connected in this manner, the amplifier offers a basically high input impedance at its non-inverting input which can be reduced as far as the source is concerned by inserting suitable passive input networks where necessary. The amplifier shown accepts input signals from any one of five sources, viz magnetic microphone, piezo pickup, magnetic pickup, tape replay amplifier and a radio tuner. The amplifier was designed to give an output of 350mV at 1kHz giving, at this frequency, gains for the above inputs of approximately: 100; 1.15; 86.5; 1.15 and 2.47 respectively, with R_{12} chosen as a fixed value to simplify the switching arrangement. Components R_8, R_9, C_2 and C_3 provide an RIAA equalization when using the magnetic pickup input and R_3 was chosen to load the cartridge with 47k Ω .

Circuit description

Useful range of supplies: ± 3 to $\pm 18V$.

741 operational amplifier may be replaced by a 748 or 301



using a 30-pF compensation capacitor. Break-before-make switches could be replaced by make-before-break types to prevent sudden changes causing annoying clicks in the loudspeaker.

C_4 and C_5 may be included to roll off the flat response at high frequencies. The C_4 to R_{10} network could be changed to provide additional gain and the correct equalization when the preamplifier is fed directly from a tape head instead of from a tape replay amplifier.

Circuit description

● A resistor of the order of 10k Ω may be taken from the output of the preamplifier to supply a tape recorder.

● A disadvantage of the switched-network type of pre-amplifier is that only one signal source may be used at any given time. If each source is provided with a separate and appropriate preamplifier, any or all the inputs can be fed to a following mixer stage via faders making the system more flexible. Further versatility can be obtained by inserting a separate tone control circuit between the preamplifier and fader of each channel as shown in diagram left.

● As the various input sockets must be physically separated to some extent, the provision of a separate preamplifier for each source is useful as it can be placed in close proximity to its input socket in an attempt to reduce noise and hum pick-up.

● Diagram right shows a preamplifier using the see-saw configuration where networks X and Y are designed to provide the desired gain and equalization for a particular source, the networks again being introduced by switches.

Further reading

Evans, J. H. & Williams, P., Modular integrated circuit audio mixer, *Wireless World*, Dec. 1972, pp.564-70.

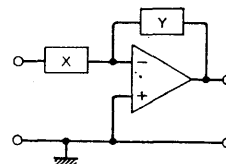
Linsley Hood, J. L., Modular pre-amplifier design, *Wireless World*, vol. 75 1969, pp.306-10.

High Fidelity Audio Designs, Ferranti Ltd 1967.

Thorig, J., High-quality pre-amplifier, *Mullard Technical Communications*, vol. 11, 1970, pp.153-5.

Cross references

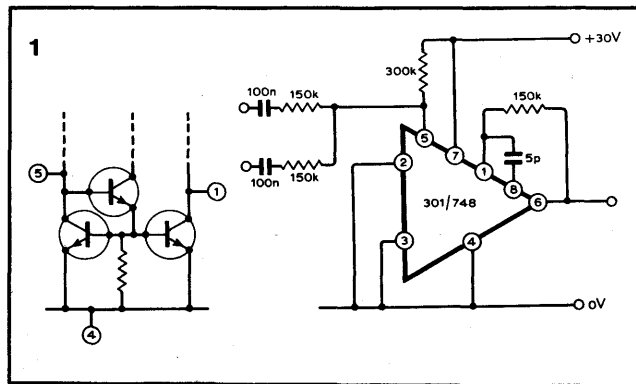
Series 5, cards 1, 4, 5, 8 & 11.



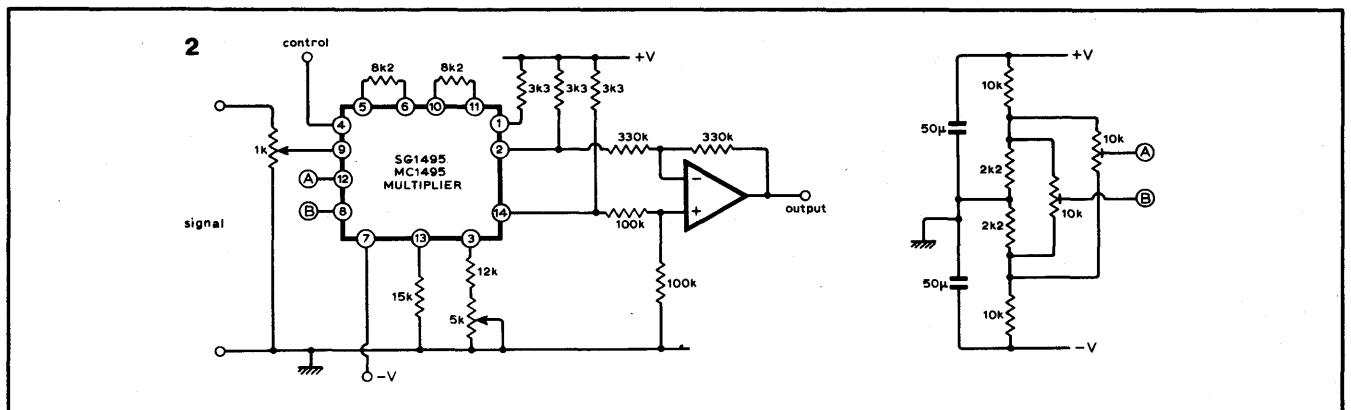
Audio circuits

1. Standard operational amplifiers have terminals other than the usual inverting/non-inverting inputs into which signals may be injected, though the terminals are designed to allow balance compensation etc. An example is shown where the op-amp is made to behave as a current-differencing amplifier (equivalent to LM3900). The input and feedback paths have no direct connection and single-ended supply operation is shown. An audio mixer is given as an example, but the technique can be extended to produce astable/bistable circuits, RC oscillators etc with the added option that the normal inputs are available for the injection of synchronizing pulses if required.

June, W. G. Op-amp in current differencing mode becomes a noninverting audio mixer, *Electronic Design*, vol. 22, 1974, p.130 (May 10).

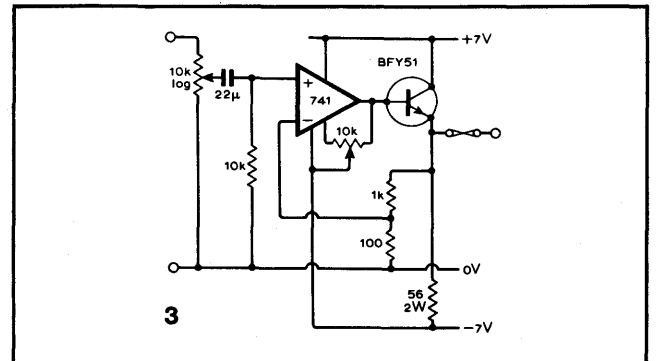


2. Analogue multipliers fed with a signal at one input and a single-polarity direct control voltage at the other give an output duplicating the signal in form but with a magnitude that can be reduced to zero for zero control voltage. If that control voltage is derived from a precision half-wave rectifier its value can be restricted to a given polarity, while depending on some combination of bias and control signals



(see reference article for details). The offset network shown is based on that of the i.c. manufacturer, and allows the output to be set as a true product of the two inputs. A similar circuit is used in the voltage-controlled filter.

Orr, T. and Thomas, D. W. Electronic sound synthesizer: part 2, *Wireless World*, vol. 79, 1973, pp.429-34.



3. A headphone amplifier has to accommodate a wide range of load resistances (e.g. 8 to 100 Ω), have a low output impedance for good damping, and should meet high standards in terms of frequency response noise etc since the headphones now available are themselves of very high quality. A simple circuit meeting these requirements takes a standard operational amplifier and adds a class A emitter-follower stage to enable the voltage gain to be sustained into loads as low as 8 Ω . A further reduction in cross-over distortion can be achieved by adding a 1-k Ω bias resistor from the op-amp output to the negative supply rail (i.e. operating output stage of op-amp in class A). The transistor operates at \approx 800mW dissipation with a quiescent collector current of 125mA and requires a heat sink thermal resistance of 50deg C/W.

Wherritt, P. Headphone amplifier, *Hi-Fi News*, vol. 19, 1974, pp.63-5 (March).

Wireless World Circard

Series 6: Constant-current circuits

Why aren't constant-current circuits used as much as constant-voltage circuits, when for every constant-voltage circuit there is a constant-current dual? For the most part, the reason is probably historical, as the following article points out, and electrical sources have for a long time been constant-voltage. The possibility of using constant-current circuits in place of those designed for constant voltage is interesting. If, for instance, inductors were inconvenient in a certain constant-voltage circuit, the dual would allow capacitors to be used instead, given a suitable constant-current supply.

These cards then give circuits for constant-current supplies, both a.c. and d.c., and one card deals with their use for measuring characteristics of semiconductor devices (card 11, useful for electroplating too).

An extremely simple circuit is given on card 3 for limiting current in the range 1 to 35 mA, together with data obtained as a result of measuring 20 samples of 741 and 301 op-amps in various configurations. The limiting obtained is only approximate and more precise methods of fixing current are given on other cards. How to use voltage regulators in constant-current circuits is explained in card 2, the current-mirror in card 4 and transconductance amplifiers in card 12. The well-known ring-of-two circuit (a.c. card 5, d.c. card 6) is adapted for low-voltage use in card 9.

- Hybrid constant-current circuit 1
- Constant-current use of voltage regulators 2
- Simple current-limiting circuits 3
- Current mirror 4
- AC constant-current circuits 5
- Ring-of-two reference 6
- Switching current regulator 7
- Thyristor control current regulator 8
- Low-voltage current regulators 9
- High-power current regulators 10
- Constant-current applications 11
- Constant-current amplifiers 12

Constant-current Circuits

For every circuit using a constant-voltage element or sub-section there is a dual circuit based on constant-current properties. That such circuits are less common and often misunderstood is partly for historical reasons, stemming from the lack of sources of electrical energy having constant-current characteristics. One cannot draw out of stores a "5A 250mV-hour battery". A battery that will sustain a constant current into an arbitrary load is not physically realizable, since the electro-chemical processes involved define the e.m.f., the current then being inversely proportional to resistance.

Either capacitors or inductors may be used for temporary storage of energy, but the cost and size penalties of the latter are considerable. Capacitors store charge, having a p.d. proportional to the stored charge, and sustain that p.d. to a first order against varying current drain; until the drain results in a significant loss of charge. Even more important is that the generation, transmission and transformation of a.c. by the Electricity Boards are all constant-voltage processes. After rectification, the only form in which d.c. power can be produced efficiently and with freedom from ripple is as a constant voltage.

Thus all common sources of electrical energy approximate to constant-voltage characteristics and the majority of electronic circuits have been designed for this mode of operation. It is a fascinating thought that there should be as many current-operated circuits as voltage-operated ones, though they may be unfamiliar in shape. For example transistors would have to be operated in series, carrying comparable currents in each device but with progressively increasing p.d.s moving from input to output in an amplifier, while the interstage coupling might be inductive. Conversely, where constant-voltage supply circuits use inductors to achieve particular effects, the corresponding circuits using capacitors would be attractive alternatives if constant-current supplies were available.

Each type of constant-current circuit seeks to achieve a constant current against variations in supply voltage, load resistance and ambient temperature as well as against component parameter changes. This will apply whether the supply is in the form of a direct voltage or an alternating current. In the former case, an intermediate step may be used in which the power is converted into a switched waveform before re-conversion to d.c., the method having high efficiency even when the load voltage is much less than the supply voltage (Fig.1). Alternatively the

d.c. may be used to power an amplifier which because of the design of its output stage or by virtue of the feedback employed, delivers a current to the load controlled primarily by some signal voltage or current (Fig. 2).

Purely d.c. systems may also fall into this category with the a.c. signal replaced by a direct voltage/current (Fig.3) which can be fixed or variable depending on the application. (Fixed if a constant current is to be forced in a zener diode to define its operating point, variable where used to plot the characteristics of a transistor.) In addition, current control can be achieved by devising a two-terminal circuit to be interposed between source and load (Fig.4). If the circuit has a high dynamic resistance the current is then stabilized against supply and load changes. To apply such a circuit to a.c. supplies involves a number of difficulties, not least that such constant-current action is achieved only

for that part of the cycle for which the input amplitude is in excess of some minimum value, typically 2 to 10V. It becomes particularly important to distinguish the parameter of the output whose constancy is being maintained. The peak value will be held constant by a two-terminal device having infinite slope resistance for amplitudes of input above the minimum. In many cases it may be necessary to rectify the applied voltage so that the circuit deals with a single polarity.

Two further parameters of interest are the r.m.s. and mean-rectified output currents. For both, the rise in current during the pre-limiting region as the input voltage increases causes a rise in the area under the current graphs, i.e. in the mean/r.m.s. current. If the two-terminal network is arranged to have a negative-resistance characteristic then the current can fall back during input peaks, offsetting the tendency for the mean/r.m.s.

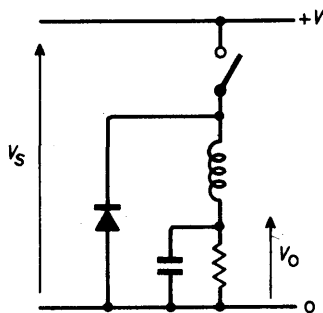


Fig.1. In trying to achieve a current constant against variations in supply voltage, load resistance, ambient temperature, and component parameters an intermediate step is sometimes used where power is switched before re-conversion to d.c.

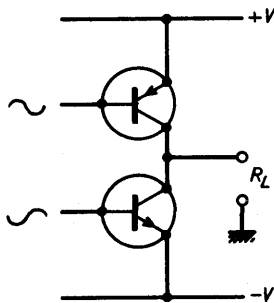


Fig.2. Another method of controlling load current.

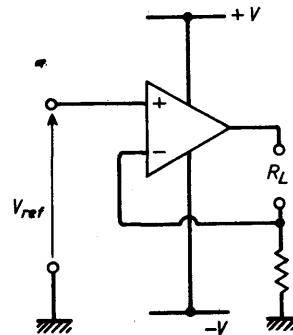


Fig.3. Technique of Fig.2 can be used with an a.c. signal replaced by a d.c. level.

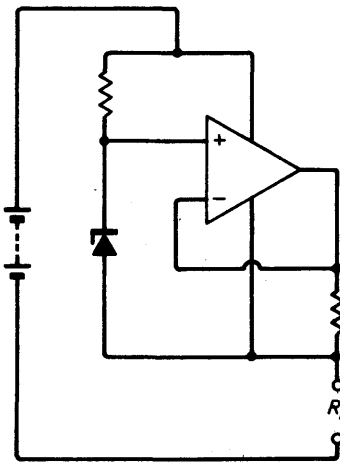


Fig.4. Current control can also be achieved by using Fig.3 as a two-terminal circuit between source and load.

currents to rise (Fig.5). A different value of negative resistance is required for the mean and r.m.s. conditions and it is further dependent on the input waveform. The method has the advantage that it operates on the instantaneous value of input, though methods based on thermistors and thermocouples might be used to monitor r.m.s. current via thermal effects. The necessary feedback would be more easily applied via conventional regulator circuitry and would involve thermal time delays that would not cope with input/load transients.

In the majority of these circuits the reference determining the current will be a voltage such as that developed across a zener diode. Where lower stability is adequate the p.d. across a forward-biased silicon p-n junction has advantages.

The voltage, or some function of it, appears across a resistor defining the current in that resistor. If the load is placed in series with that reference resistor, or in some other circuit path carrying a related current, then load current is fixed. Operational amplifiers have one output terminal committed to ground potential. If the generator representing the output has to appear in series with the reference resistor and load to define the current then a conflict appears (Fig.6). To achieve a current flow to ground the reference resistor (and with it the reference

voltage and its associated circuitry) would have to float. As this is inconvenient, circuits based on conventional operational amplifiers may be limited to constant-

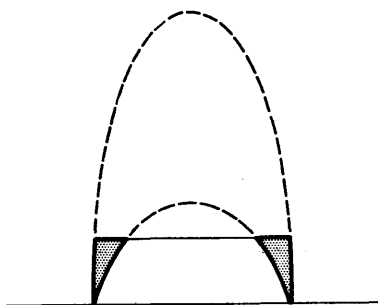


Fig.5. Negative-resistance causes current to fall back during input peaks, offsetting tendency for mean/r.m.s. currents to rise in the pre-limiting period.

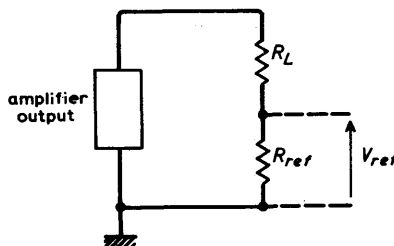


Fig.6. Circuits based on op-amps are generally limited to the case where loads do not require a ground connection.

current operation only with those loads not requiring a ground connection.

Circuit configurations are possible in which negative and positive feedback can be combined to raise the output resistance to very high values. Penalties include relatively poor stability of this output resistance and difficulties in achieving high output currents. This last demand is frequently met by adapting existing voltage regulators with a reference resistor at the normal output voltage terminals and the true load in series with it as outlined above.

Finally, the problem of controlling alternating current may be tackled in a different way by means of thyristor switches. These can be fired at appropriate points on the input waveform such that the mean current in the load is controllable. As the thyristor behaves as an almost perfect switch, no control is exercised over the instantaneous value of current. A filter provides a feedback voltage proportional to the mean current and controls the phase angle of the firing circuit. This phase angle control is quite distinct from the frequency/pulse-width modulation methods that are inherent in the switching amplifiers described earlier, and filtering of the output waveform would not normally be applied. The method would be suitable for such applications as battery charging where the current waveform is uncritical.

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This is the first ever Wireless World Annual. It's got pages of features covering all aspects of electronics and communications – new and established techniques, some practical, some theoretical – all written to the high standard you'd expect from Wireless World. Contents include: A General Purpose Audio Oscillator by L. Nelson Jones (a constructional project specially commissioned for the annual); Constructional Design for a Small Boat Echo Sounder by John French; Scientific Calculations with an Arithmetic Calculator by R. E. Schemel. There is also a reference section packed with useful information.

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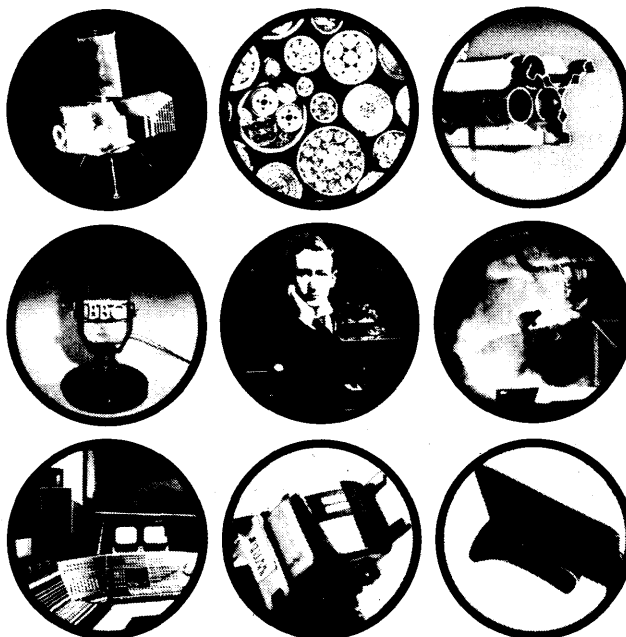
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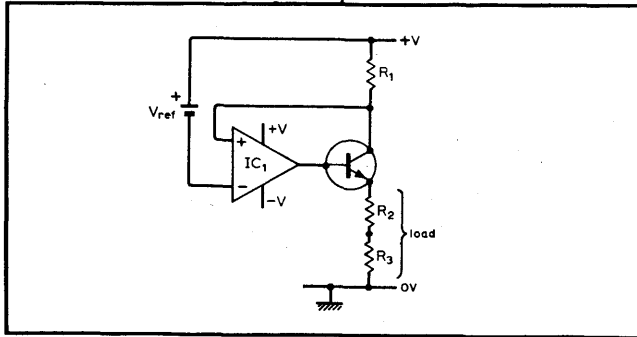
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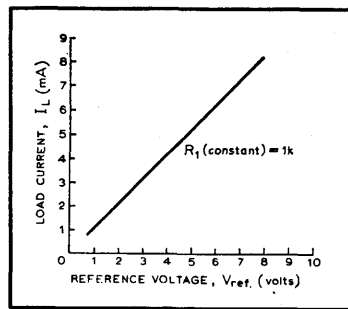
Hybrid constant-current circuit



Typical data

Supply: $\pm 15V$
 IC₁: 741; Tr₁: BFR41
 R₁: $1k\Omega \pm 5\%$
 R₂: $680\Omega \pm 5\%$
 R₃: $100\Omega \pm 0.05\%$
 Maximum current: 8mA.
 Note: Careful layout required to avoid r.f. oscillation. Absolute measurement of current obtained using accurate digital voltmeter across R₃. Variation of load

current with reference supply shown opposite.



Circuit description

This circuit permits high currents through the load (R₂ + R₃ in series), depending on the current capability of the bipolar transistor used. Negative feedback is applied via the operational-amplifier IC₁, the feedback being applied to the non-inverting terminal and being derived from the collector of transistor Tr₁, where inversion has occurred. Load current is essentially defined by V_{ref}/R_1 , because the potential difference between inverting and non-inverting inputs of the operational amplifier when the gain is high, is very small. This reference voltage, symbolised by an ideal battery, may simply be a reverse biased zener diode in series with a resistor connected across the d.c. supply, the inverting input being connected to the junction. This has the disadvantage of being uncompensated for temperature variations. If the zener diode has a positive temperature coefficient, this can be offset by connecting a forward-biased silicon diode with a negative temperature coefficient in series. Such a combination is available in a single package to provide a temperature-compensated zener diode.

If the current through R₁ increases, the potential difference

across R₁ increases, and the voltage applied at the non-inverting terminal decreases. This change is amplified by the operational-amplifier, and hence the base drive to Tr₁ is reduced, tending to compensate the original increase of the collector current which is approximately equal to the load current. As the gain of IC₁ is high, the input current demanded by this operational amplifier is extremely small, and the feedback also increases the effective output impedance of Tr₁.

Component changes

- IC₁: LM101, Tr₁: 2N2219, Tr₂: 2N3456.
- Supply: useful range down to $\pm 9V$. Typically variations of current better than 0.05% over this range, when V_{ref} is independent of the supply.
- If oscillation exists, connect a capacitor across R₁.
- Useful range of R₂: 330 Ω to 3.3k Ω . At 2mA load current variations less than 0.05%.
- At 2mA, variations are less than -2% with BFR41 h_{fe} in the range 90 to 220.
- Absolute measurement of current through R₁ and emitter current indicated a variation of around 1.5%.

Circuit modifications

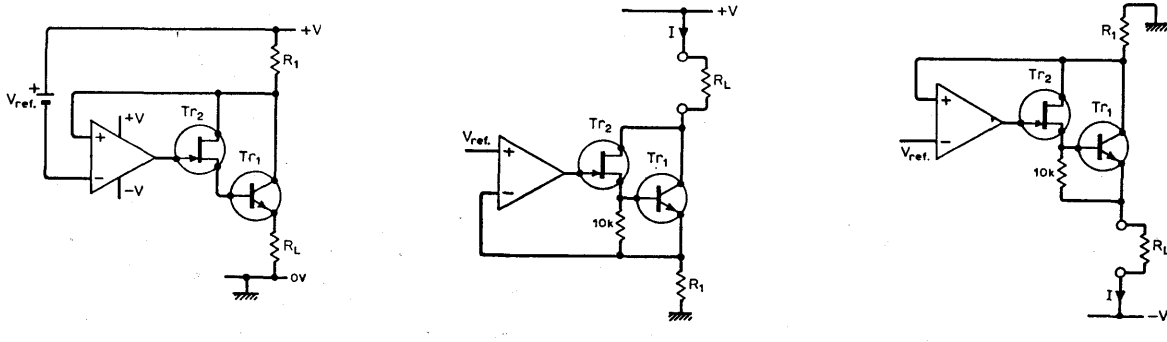
Current through R₁ is defined by V_{ref} in circuit shown left. However in this circuit, the current shunted from the collector to the non-inverting input of the operational-amplifier is considerably less than the original circuit, as the output current demanded from the op-amp is only the gate current of the f.e.t. Tr₂. The f.e.t.-bipolar compound pair has a much higher current gain and the load current is more nearly equal to that defined by V_{ref}/R_1 .

- Use f.e.t. 2N5457 to drive the bipolar transistor. Absolute measurement of current through R₁ and emitter currents of Tr₁ now indicate a variation of less than 1%. R₁: 1.1k Ω , R₂: 100. V_{ref} adjusted to give load current of 2mA. R₃ varied from 4.7k Ω (max) down to 10. Current change within 0.01%.
- Alternative arrangement of feedback connection shown centre and right. Circuit in centre uses the output stage as a non-inverting follower allowing feedback to be returned to the inverting terminal of the op-amp. This arrangement is sometimes known as a current sink. Circuit right shows the corresponding current source. This may have both the reference voltage and reference resistor returned to ground or the positive supply rail with the load returned to the negative rail for increased load potential difference.

Further reading

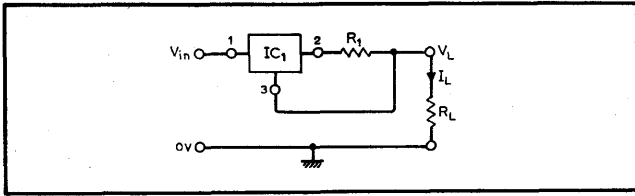
National Semiconductor Linear Applications AN-20. Hart, B., Current generators using unipolar-bipolar transistor hybrids. *Electronic Applications* 1966 vol. 27, pp. 30-7. Silicon Zener Diode and Rectifier Handbook, Motorola.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 2

Constant-current use of voltage regulators



Typical performance

V_{IN} : +15V
 V_L : 3V
 IC: LM309H
 R_1 : $245\Omega \pm 1\%$
 R_L : $120\Omega \pm 5\%$
 I_L : 25mA

For 1.5V pk-pk input ripple at 100Hz, load current ripple is approx. $16\mu A$ pk-pk.
 Dynamic output res: $90k\Omega$
 Dynamic to static output resistance ratio: ≈ 220
 For $25mA < I_L < 200mA$, I_L changes less than 3% for a 100% increase in R_L .

Circuit description

A very simple constant-current generator can be produced by placing a sufficiently large resistance between a constant voltage source and a load. This leads to a requirement of very high source voltages to supply constant currents of only a few mA. This simple approach is normally unacceptable. However, a constant-voltage regulator can be made to provide a constant current into a load, at reasonable voltages, while only carrying a relatively small standing current. The diagram above shows a monolithic voltage regulator connected as a two-terminal constant-current generator. This regulator was designed primarily as a fixed 5-V voltage regulator to supply the widely varying currents in logic circuitry. In the constant-voltage mode, R_1 would be set to zero and terminal 3 connected to ground instead of the output terminal. The circuit thus provides a regulated output voltage between terminals 2 and 3. Inclusion of R_1 between these terminals as shown ensures that it receives a constant voltage from the regulator and therefore carries a constant current which is supplied to the load resistance. (The stability of R_1 determines the stability of I_L .) The load will also carry the quiescent current from terminal 3 but this will normally be much smaller than the current in R_1 . This quiescent current places a lower limit on the available output constant current. The voltage regulator chip incorporates a temperature regulator to provide thermal, rather than current, protection. This technique allows a considerable increase in the maximum allowable output current, the device being protected against almost any overload condition.

Component changes

Useful range of V_{IN} + 6 to +35V.

I_L (min) $\approx 10mA$: lower limitation due to quiescent current at regulator terminal 3.

I_L (max) $\approx 200mA$: power dissipation limitation of 2W in regulator without heat sink.

For I_L values of 50, 100 and 200mA typical values of R_1 with $V_L = 3V$ are 109, 51.35 and 25.2Ω respectively.

If regulator is placed some distance from the d.c. supply filter, a capacitor of about $0.1\mu F$ may be required between terminal 1 and ground to prevent h.f. oscillation.

For higher output currents, up to about 1A, the LM309H can be replaced by an LM309K.

Circuit modifications

Any voltage regulator that can sustain a constant load voltage at a high current compared with its standing current may be used as a constant-current generator. Circuit shown left is a standard form of voltage regulator using Tr_1 and Tr_2 as a long-tailed pair with Tr_3 and Tr_4 forming a Darlington-connected output transistor. The long-tailed pair compares the reference voltage from the zener diode with the output voltage across a dummy load R_2 . If the voltage regulation is good and R_2 is constant then the current in it is constant. The current in the real load R_L is this current plus the currents in the long-tailed pair and reference diode, both of which can be made very much less than the dummy load current. If the "free" collector of Tr_3 and Tr_4 is accessible in the voltage regulator, R_1 may be placed between it and the positive supply, although R_L will not then be referred to ground.

Another floating-load constant-current generator is shown, middle, which applies the principle of series feedback. The p.d. across R_3 is a defined constant voltage and so also is the current in it. This current is virtually identical with that flowing in R_L . Amplifier could be a Darlington-connected pair.

Existing voltage regulators, even of the poorest kind, can be used to provide a constant current, one example being shown right. The zener diode fixed the p.d. across the emitter resistor R_4 and hence the current in R_L . This circuit suffers from the usual problems of matching up the temperature coefficients of the zener diode and transistor.

Further reading

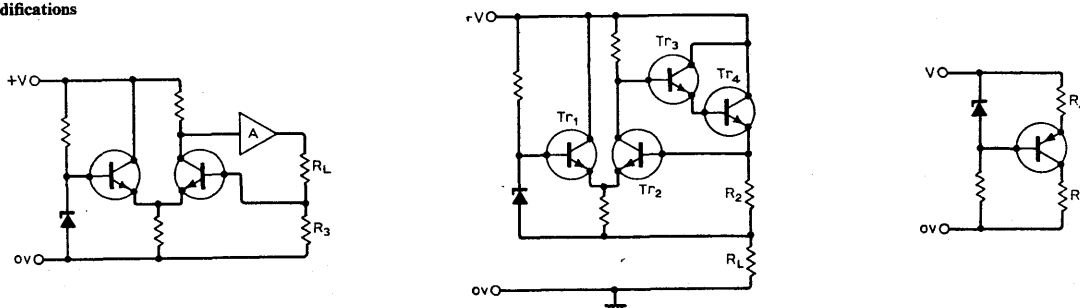
Linear Applications Handbook, National Semiconductor, AN 42-1 to 42-6, 1972.

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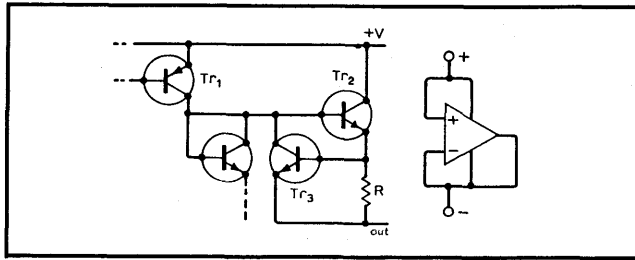
Series 6, cards 13, 10 & 11.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 3

Simple current limiting circuits



Typical performance

IC: N5741V (Signetics)

Supply voltage: 10V

Current: 26-30mA

Voltage for limiting: 7-9V

10 samples of other manufacturers 741/748

i.c.s gave current range 20-35mA.

10 samples from three manufacturers 301 i.c.s gave currents of 15-25mA, but included devices requiring only 2-3V to achieve limiting.

Circuit description

Many i.c. op-amps have protection circuits at their output which limit the current that can flow, even into a short-circuit of the output to either supply line, and regardless of the condition of the input terminals. The current is not defined as precisely as with the other constant-current circuits described on these cards; the limiting action is only intended to be approximate, and generally uses the base-emitter junction of a transistor as the sensing element (e.g. with Tr_3 as in a section of an i.c. shown above). Transistor Tr_3 is one of the output transistors and if the output current flowing through R tries to exceed the value at which the V_{be} of Tr_3 reaches 0.5V, Tr_3 comes into conduction, diverting the base current supplied by Tr_1 and preventing further increase in output. In general, the limit current falls with increasing temperature because the V_{be} of Tr_3 required for conduction falls, and the resistance of R increases with temperature. Such a mechanism is thus not adequate for precision constant current action but can offer good rejection of supply variation including ripple. If an i.c. op-amp having such limiting has its output shorted to one supply line and the inputs connected to the supply lines, in the sense that causes the output to try to drive towards the opposite line, the limiting mechanism comes into play and the complete circuit may be used as a two-terminal device. Placed between source and load, the load current is limited typically to 12-30mA depending on amplifier type for any p.d. across the amplifier above some minimum voltage (5-9V). The max p.d. across the amplifier must not allow the device dissipation to be exceeded, though self-heating minimizes the dissipation by reducing the current.

Component changes

● With output open-circuit the circuit may also draw constant current but of much smaller magnitude. Similarly, connecting output to opposite supply and/or reversing input terminals brings different sections of the circuit into action, i.e. several different current limits can be obtained.

● With typical device from N5741V range, six configurations were tested, as below, with minimum voltage of 8V throughout; tests carried out at 10V and resulting current limits from 0.85 to 30mA obtainable from single device:

inv. input	non-inv. input	output	current (mA) at 10V
+	-	+	30
-	+	-	29
+	-	-	12
+	-	o/c	1.4
-	+	+	0.9
-	+	o/c	0.85

Current reduction 20% for temperature increase of 50 deg C.

Circuit modifications

● The basic idea of using a transistor to monitor the p.d. across a current-carrying resistor is also applied in voltage regulators to limit the output current even into a short-circuit load. Here, Tr_3 deprives Tr_2 of base-current, monitoring the p.d. across an external resistor R_2 . This allows boosting of the output current via external transistor Tr_1 , a variable R giving control of the current limit.

● Limiting by sensing of the collector current of the output stage is also possible. The nature of the drive circuit is often such that a loss of, say, 1V in the collector circuit does not further increase the minimum supply voltage. As shown, Tr_1 is a constant-current stage biased by D_2 acting as a high-impedance load for the error amplifier (not shown). As the output current increases so does the p.d. across R_2 bringing D_1 into conduction and diverting current from Tr_1 i.e. limiting base current of Tr_2 .

● In principle simple limiting circuits may be added to any voltage regulator. Shown is a method by which base current is diverted from the series pass transistor by Tr_3 which senses the p.d. across R_4 . In this case it is the total current that is limited i.e. load current plus circuit quiescent current.

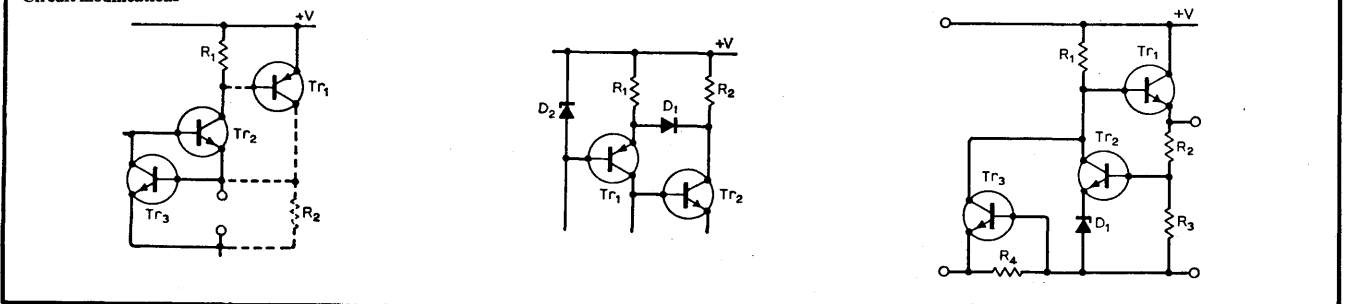
Further reading

Patchett, E. N., Automatic Voltage Regulators & Stabilizers, Pitman 1970, pp. 364-6.

Cross references

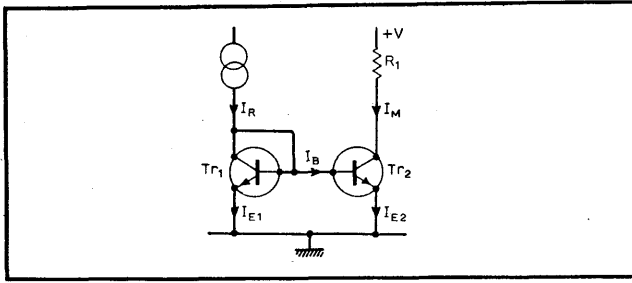
Series 6, cards 2, 5 & 9.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 4

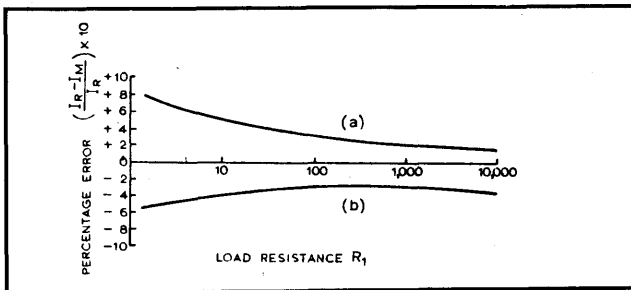
Current mirror



Typical performance

Supply: +6V
 Tr₁, Tr₂, part of CA3046
 R₁: 0-10kΩ decade resistor, ±0.05%
 I_R: 0-5mA from commercial current generator, ±0.05%
 I_M: Calculated from voltage reading across R₁ using five-digit voltmeter
 Dynamic output

impedance: 2MΩ at 50μA
 Curves below show percentage variation of 'mirror' current to reference current for the basic (a) enhanced (b) current mirror circuits, for currents in the range, 1μA to 5mA.
 Product of I_MR₁ maintained constant.



Circuit description

Circuit configuration is known as a 'current mirror' and is widely used in integrated circuits. If the two transistors Tr₁ and Tr₂ are considered identical so that the base-emitter voltages are the same, then to a first order the collector currents will be the same. Transistor Tr₁ acts as a diode whose forward voltage between base and emitter defines the base-emitter voltage of transistor Tr₂. If Tr₂ has a high current gain, then the reference current I_R will be approximately equal to the collector 'mirror' current I_M.

$$I_R = I_B + I_E = I_{E2}/(1 + \beta) + I_{E1} = I_{E2} \left(\frac{1}{1 + \beta} + 1 \right)$$

$$I_M = \alpha I_{E2} = \frac{\beta I_{E2}}{1 + \beta}$$

$$I_{E2} = I_R \cdot \frac{1 + \beta}{2 + \beta}; \therefore I_M = \frac{\beta}{1 + \beta} \cdot I_R \cdot \frac{1 + \beta}{2 + \beta} \approx I_R.$$

Hence if the reference current is fixed, the collector current of Tr₂ is fixed.

Discrete components are temperature sensitive and the circuit is not reliable with them. Closer matching of the transistor parameters and the facility of compensating changes due to temperature are available, when the transistors are produced on the same monolithic silicon chip. The circuit is thus often used in the reference stage for basic regulator circuits. Output impedance is approximately that of a common-emitter configuration; the effective resistance connected across base and emitter is the low dynamic resistance of Tr₁ connected as a diode.

Output resistance characteristic of this circuit is increased considerably by including a diode connected transistor in series with the emitter of Tr₂ as shown below (middle).

Component changes

- Dynamic output impedance reduces to 200kΩ for a current of 500μA, and 90kΩ for load current of 1mA.
- Percentage mirror current error is typically better than 2.5% for I_M = 500μA when R₁ is varied from 0-10kΩ without attempting to maintain V_{ce} of Tr₂ constant.

Circuit modifications

● Output impedance of the current mirror is increased by negative feedback via resistor R₂ (left) but its use should be restricted to currents in the microamp range.

● Higher output impedance obtained using the enhanced circuit shown middle. This requires about 1.2V minimum before control commences as the V_{be} of Tr₂ and Tr₃ must be overcome. The resulting transfer ratio of I_M/I_R can be shown to be (β² + 2β)/(β² + 2β + 2) indicating an improvement dependent on the β² term, the (2β + 2) term becoming insignificant for high-gain transistors.

● Current mirror, shown right, available within transistor package CA3084. This is a p-n-p version and illustrates the use of the current mirror in establishing multiple current sources. Diode D₁ is a transistor with its base and collector connected. The V_{be} values for each transistor are identical, and hence control of D₁ current ensures first-order constancy of currents in Tr₁ and Tr₂. In practice, the increased number of units of base current degrade the stability if too many stages are controlled.

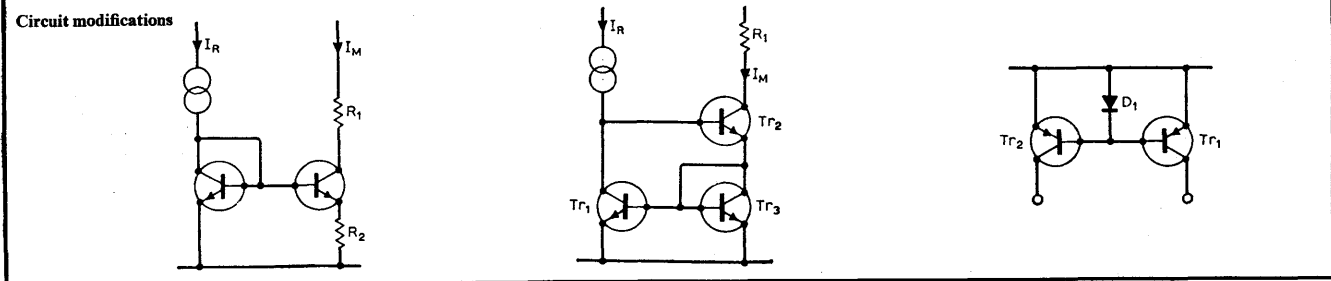
Further reading

Hart, B. L., Current generators, *Wireless World*, vol. 76, 1970, pp. 511-4.
 RCA Solid-State Databook, Series SSD-201, 1973.
 RCA Solid-State Databook, Series SSD-202A, 1973, pp. 325-76.

Cross references

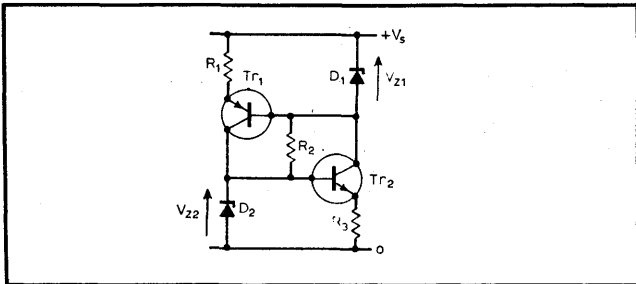
Series 6, cards 5 & 12.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 5

Ring-of-two reference



Typical performance

Minimum terminal p.d.

$$\sim V_{Z2} + V_{Z1} - 0.5V$$

Constant current

$$\sim \frac{V_{Z2} - 0.6}{R_3} +$$

$$\frac{V_{Z1} - 0.6}{R_1}$$

Tr₁: 2N3702

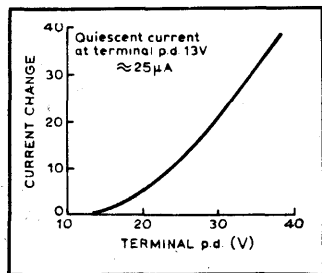
Tr₂: 2N3707

R₁, R₃: 470kΩ; R₂: ∞

D₁, D₂: Reverse biased base-emitter junction at planar transistor e.g. 2S512

Comparable results for

currents up to several mA. Self-heating effects significant at higher current.



Circuit description

The ready availability of two-terminal elements which can be placed in parallel with a load to make the load voltage stable is not matched by dual elements for sustaining constant load currents. Constant-current diodes are available but are no match for the variety and performance provided by zener diodes. Two problems have to be overcome in designing a two-terminal constant-current circuit. There will usually be two or more separate paths for current flow and they must either be separately constant or, if variable, such variations must be restricted to a low-current path. A second problem is that the minimum p.d. at which constant-current is achieved must be as low as possible, while the breakdown voltage should be high. The ratio of these p.d.s is one guide to the usefulness of the circuit and a ratio of 10:1 or greater is good. The upper voltage is fixed in the present circuit by the V_{cb} breakdown of the transistors and the lower voltage by the sum of the V_z values. The two current paths are separately constant and may be made equal or not as required. Diode D_2 maintains a

constant potential at the base of Tr_2 and hence a constant p.d. across R_3 ($V_z - V_{be}$). The resulting constant emitter current ensures that the collector current of Tr_2 and hence the current in D_1 are also constant. Similarly the p.d. across R_1 is defined ensuring the stability of current in D_2 . Thus each diode defines the current flowing in the other. The circuit is a form of complementary bistable and precautions must be taken to ensure that the on-state is the only practical one. This may be achieved by a starting resistor R_2 between the bases (or from Tr_2 base to +ve line for example).

Component changes

Tr₁, Tr₂: General purpose silicon e.g. n-p-n. types ME4103, 2N706, BFR41; p-n-p types 2N3702, ME0413, BFR81.

D₁, D₂: Zener diodes 2.7 to 12V. Low voltage units (2.7 to 4.7V) give minimum terminal p.d. and first-order compensation for V_{be} temperature drift. Higher voltage units increase dynamic resistance of circuit. Zeners of breakdown $\approx 6V$ have low temp. drift, and additional forward-biased diode in series gives temp. comp. (For very low voltage operation see card 9). Diodes need not have equal breakdown voltage. For low currents reverse breakdown in planar transistor base-emitter junctions offers good performance.

R₁, R₃: 330kΩ 1MΩ. At higher currents, self-heating effects vary current as terminal p.d. changes. At lower currents, low-leakage transistors used for Tr₁, Tr₂. Zeners may be replaced by reverse-biased base-emitter junctions of planar transistors (breakdown voltages typically 5 to 10V, fairly close tolerance for given device type).

R₂: Typically 330kΩ to 10MΩ. Use highest value that ensures self-starting. 1MΩ adequate with all except high leakage zeners.

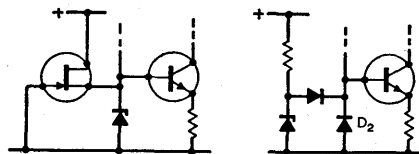
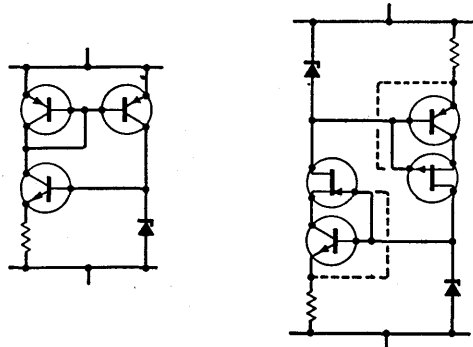
Circuit modifications

- To minimize the p.d. at which the circuit achieves constant-current operation, only one half of the circuit has a zener diode. The other half may have the zener replaced by any other element that sustains an approx. constant, p.d. against variation in current. A current mirror in one of its forms allows the circuit to function correctly for a terminal p.d. barely more than the zener voltage. Alternative circuits (card 4) can increase accuracy of current for small increase in minimum p.d.

- For highest dynamic resistance, each transistor may be replaced by cascode or similar circuits while retaining defined V_{be} characteristics of bipolar transistors. Alternative connection for f.e.t. gives higher dynamic resistance but version shown allows f.e.t. to operate with slight forward bias if required, increasing the current capability.

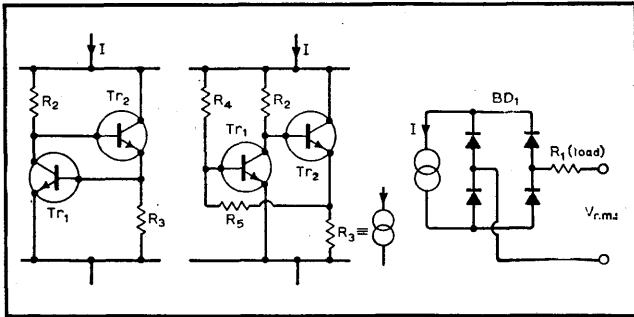
- Circuits are all bistable in form, with a possible non-conducting state. Any resistive start-up circuit degrades dynamic resistance. Use of junction f.e.t. with pinch off between V_{be} and V_z inhibits off-state without contributing current in one state. Identical zener diode with high resistance drive brings D_2 into conduction-preferred method in some i.c. regulators but current in R flows in load if used as two-terminal constant-current circuit.

Circuit modifications



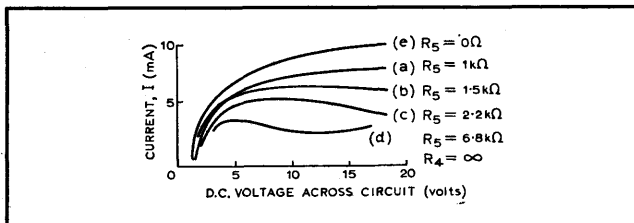
Wireless World Circard Series 6: Constant-current circuits 6

A.C. constant-current circuits



Typical performance

Tr_1, Tr_2 : BC125 R_5 : 1.5k Ω
 BD_1 : A154 Current constant at
 R_1 : 47 Ω ; R_2 : 3.9k Ω 5.8mA $\pm 1\%$ for direct
 R_3 : 100 Ω ; R_4 : 47k Ω voltage of 6 to 18V.



Circuit description

If a circuit can maintain a constant voltage across a resistor against changes in the supply voltage, then the current flow in this resistor is maintained constant. If this current is greater than any other current in the circuit, then the total current taken from the supply is reasonably constant. A simple circuit that attempts this has the base-emitter of Tr_1 in parallel with a 100 Ω resistor R_3 , maintaining a current through R_3 of about 6mA with the feedback loop closed via Tr_2 . Although the current in Tr_1 varies when the applied voltage varies, this current is appreciably less than that in Tr_2 , and so the dynamic impedance of the circuit used as a two terminal element is high. A more complex amplifier, e.g. a Darlington pair, in place of Tr_2 would allow the contribution to total current change, due to the current in R_3 , to be very small.

An alternative arrangement is to introduce R_4 and R_5 . If supply voltage increases, this potential divider increases p.d. across R_5 . The base potential of Tr_1 is substantially constant, and hence p.d. across R_3 must fall, and hence the current i.e. a relatively large increase in the current in R_2 (which is small)

is balanced by a small decrease in the relatively large current through R_3 . By suitable choice of R_4 , R_5 , the dynamic resistance can be controlled to be positive or negative, and with a critical value of R_5 is extremely high over a wide range of supply voltage. The operation of the circuit below 5V is non-linear.

When a.c. is to be applied, it may first be rectified so that the circuit sees a unidirectional voltage, but only the peak current can be controlled i.e. currents corresponding to voltages in excess of 5V. To control the r.m.s. value of current, and if the waveshape is unimportant, the negative resistance effect allows the current to fall during the peaks of the applied signal, compensating for the rise during the rest of the cycle. Adjustment is empirical and depends on waveshape, but offers a simple means of controlling current in a resistive load for heating, or the mean charging current in the battery.

Circuit modifications

- A high current gain in the output stage of the simple circuit, allows the bias current to be very small (left) and is therefore also suitable for high current circuits. Also Tr_1 had to act as both an error amplifier and reference against which the current is being compared i.e. the V_{be} of the transistor. To improve this, a zener diode may be added as reference with the transistor primarily performing the function of error amplifier.

- The bias current itself may be made constant if resistors are replaced by elements which are two-terminal constant-current devices (e.g. f.e.t.) which may itself be combined with a better amplifier such as an op-amp, to give improved overall stability.

- The control of alternating currents is possible where devices are available which may be made to directly accept signals of both polarities (right). One practical case is a junction f.e.t. in which a resistor-diode network attached to the gate allows interchangeability of source and drain, e.g. when supply to A is positive, D_2 conducts, clamping gate close to source voltage, and Tr_1 current is near maximum value and unvarient with respect to further increase in supply. As f.e.t.s have great variation in pinch-off voltage and 'on' current, equal resistors are connected into source and drain paths, to exercise control over the current.

Further reading

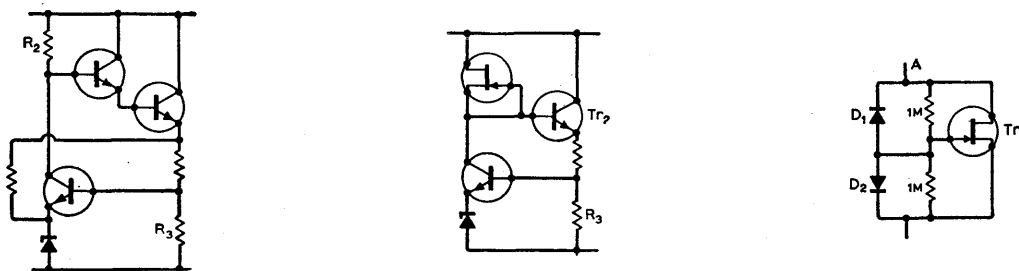
Williams, D. A., High-voltage constant-current source, *Wireless World*, Jan. 1972, p. 29/30.

Watson, G. Two transistors equal one, *Electronics*, 6 July, 1962.

Cross references

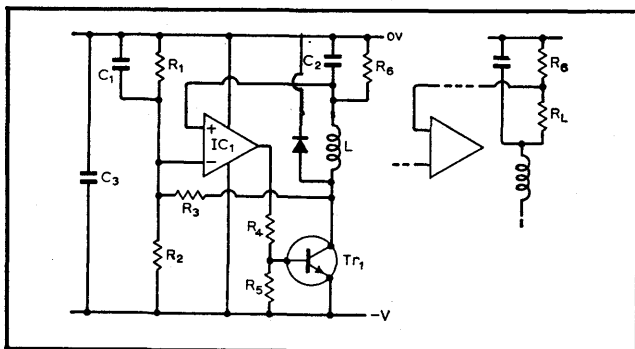
Series 6, cards 9 & 12.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 7

Switching current regulator



Typical performance

IC: 301
 Tr: TIP3055
 D: 1A 25V diode
 C_1 : 1nF
 C_2 : 22 μ F 6.3V Tantalum
 C_3 : 22 μ F 20V Tantalum
 R_1 : 1k Ω ; R_2 : 5.6k Ω
 R_3 : 470k Ω ; R_4 : 220 Ω
 R_5 : 150 Ω
 L: 5mH (Ferrite core)*
 For $R_6=2\Omega$, $V_s=-10V$
 load voltage: 1.2V,

supply current: 150mA
 switching frequency: 4kHz
 *See component changes
 ripple voltage: 100mV
 stability: output change
 < $\pm 4\%$
 for supply 5 to 20V
 output change
 < $\pm 1\%$
 for load resistance
 2 to 15 Ω .

Circuit description

The key difference between switching regulators and conventional types lies in the discontinuous operation of power stage which is isolated from the load by an LC network. The power transistor delivers current for short periods to the inductor and during its non-conducting period the current flow in the inductor is sustained through the diode. The resulting voltage step across the inductor (approximately equal to the supply voltage) defines the rate-of-change of current in terms of the inductance. If the period is short enough, the current is relatively constant, and together with the filtering action provided by the capacitor, the ripple voltage across R_1 can be small compared with its mean p.d. The circuit may be alternatively viewed as a simple astable in which the inversion due to the output transistor interchanges the functions of the op.amp. input terminals, while an LR circuit replaces the conventional CR version. Hysteresis provided by R_3 defines the pk-pk swing that will occur across R_6 . The smaller this hysteresis, i.e. the larger R_3 , the smaller the resulting ripple. This brings with it increased frequency of operation, as the rate-of-change of voltage is a function of L, C_2 , R_6 as outlined above. Mean level across R_6 is fixed by that across R_1 and is a fixed fraction of the supply voltage. In most applications this potential divider is replaced by stable reference voltage of suitable value (see cards 5, 9). As shown, the circuit acts as a

voltage regulator for a load at R_6 . To be used as a constant-current source the load may be placed series with the resistor across which constant p.d. is developed. Switching regulators may be driven by an external oscillator with the internal positive feedback eliminated.

Component changes

L: Frequency of operation is a compromise; too high and amplifier switching times limit performance, too low and increased inductance brings reduced efficiency because of winding resistance. Coils wound on ferrite rings/cores offer wide range of operating frequencies with minimum radiation of switching harmonics if shielded units used. Typical range 200 μ H to 10mH.

IC₁: Uncompensated op.amp. 748, etc. Possibility of 741,301 compensated amplifiers at low frequency with suitable choice of ferrite.

Tr₁: For currents < 500mA: BFR41, BFR50 with reduced efficiency; somewhat higher frequencies at moderate currents: MJE521.

R_1 , R_2 , R_3 : Set reference voltage/hysteresis. R_1 , R_2 replaced normally by separate reference circuit.

Circuit modifications

- To stabilize load voltage/current some stable reference voltage must be added. A simple circuit that allows operation down to very low supply voltages, tolerates high voltages and gives reasonable stability against temperature changes, matches the V_{be} characteristics of a silicon against a germanium transistor. Unselected units give a variation in reference voltage against supply of < 2% over the whole supply range of the regulator (e.g. 3 to 20V), and a typical temperature drift of < 0.1% per deg. C.

- Output current can be increased by replacing the drive transistor by any high gain combination such as the Darlington pair provided frequency is not too high (charge storage problems) and the increased losses due to saturation are acceptable. At low supply voltages the collector of the first transistor may be returned to the zero line.

- A positive voltage regulator using a standard i.c. is given in the first reference below. It operates at a higher switching frequency and contains its own voltage reference circuit. Pin 6 compares a portion of the output voltage with the internal reference, the error amplifier driving the transistor with positive feedback via pin 6 and defining the hysteresis.

Further reading

Designing Switching Regulators, National Semiconductor application note AN-2, 1969.

Nowicki, J. R., Power Supplies for Electronic Equipment, vol. 2, Leonard Hill, 1971, pp. 153-81.

Cross references

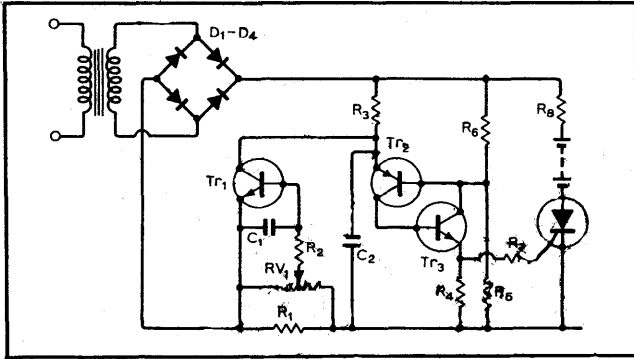
Series 6, cards 8 & 10.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 8

Thyristor control current regulator



Typical performance

T: 240V r.m.s. 50Hz primary
30V r.m.s. secondary
D₁ to D₄: 50V 1A bridge rectifier

Tr₁, Tr₃: BC 125

Tr₂: BC 126

Tr₄: 50V 1A (mean d.c.)

thyristor (2N1595 etc)

R₁: 12Ω; R₂, R₅, R₆: 10kΩ

R₃: 150kΩ; R₄: 470Ω

R₇: 100Ω; R₈: 15Ω

C₁: 470μF; C₂: 22nF

Supply: 200V r.m.s.

Battery terminal p.d.: 8V

Charging current set to:

50mA (mean)

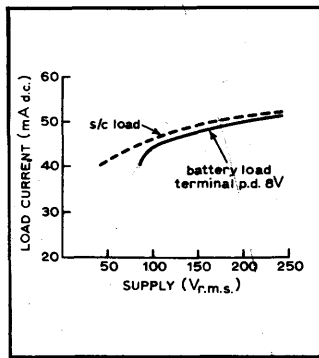
Change in current for

supply voltage $\pm 25\%$

$\approx \pm 4\%$

Change in current for terminal p.d. changed by

$\pm 2V \approx \pm 0.5\%$



Circuit description

The circuit consists of four sections: a full-wave bridge-rectified power supply; a thyristor in series with the load with the angle of conduction varying the mean load current; a pulse-generating circuit which delivers a series of pulses to the thyristor starting at a particular instant in each half-cycle and a current-sensing transistor that varies the pulsing circuit to control the mean current via the firing angle. Once the thyristor has fired, the remainder of the circuitry has no influence on the instantaneous current (determined only by the elements in series across the supply: R₁ thyristor, load, R₈). Any increase in the mean current causes the mean p.d. across R₁ to increase and via RV₁, smoothed by R₂, C₁, bring Tr₁ into conduction. This by-passes some charging current from C₂ delaying the onset of firing of the unijunction-equivalent composed of Tr₂, Tr₃, R₅, R₆ (see Series 3, card 4). The minimum p.d. wasted across current-sensing resistor R₁ need only be $\approx 0.6V$, giving good efficiency. Accuracy of control is limited by relatively low gain of control element, its temperature dependence, etc. Adding a zener diode in emitter of Tr₁ and dispensing with RV₁ would define control point more accurately at expense of increased voltage/dissipation in R₁.

Component changes

T, D₁ to D₄: Diodes must carry peak current much greater than mean current where conduction angles are small (high supply voltages, low load voltages) i.e. if mean load current is to be 1A peak currents might have to be $> 5A$. Similarly for transformer, thyristor.

R₁: At max. setting of RV₁, mean voltage across R₁ is 0.6V approx. and mean current = $0.6V/R_1$. Setting RV₁ to 50% doubles mean current, and p.d. across R₁, quadrupling power in R₁.

C₁: Smooths bias to Tr₁, 50 to 1000μF low-voltage electrolytic. R₂: Increased value allows lower C₁ for given smoothing but decreases accuracy of current. Typical range: 2.2 to 47kΩ.

R₃, C₂: To give free running frequency $\geq 100Hz$ so that firing can occur early in each cycle. R₃: 47 to 470kΩ; C₂: 10 to 100nF.

Tr₂, Tr₃, R₅, R₆: Can be replaced by single unijunction transistor e.g. 2N2646, 2N2160, etc. Any other general-purpose silicon transistors in place of Tr₂, Tr₃.

R₄, R₇: Reduce R₄ to 100 for some unijunctions. R₇ not critical.

Thyristor: Any medium sensitivity, low-voltage thyristor. For higher peak currents reduce R₁, R₈ proportionately. Resistor R₈ can be omitted if very high peaks can be tolerated by thyristor, load.

Circuit modifications

- The supply to the sensing/firing circuits may be limited and/or stabilized by a zener diode to improve control over the firing point, and to protect the circuitry when the thyristor supply is too great. For example, this would be necessary if constant-current action were desired directly from mains with no intervening transformer. Dissipation in R₅ would be high. In this, as in main diagram, a unijunction may be substituted for the complementary bistable.

- Where the circuit is to be used for battery charging, over-voltage protection might be desired. One possibility is to monitor the battery voltage directly (or better via an RC filter to eliminate spikes, as with R₂, C₁ over) using a zener diode or other suitable reference to define onset of conduction in Tr₄. The latter can then be used to raise the potential at the junction of R₅, R₆, delaying and eventually preventing firing. Addition of a series resistor R₉ to the junction of Tr₂ base/Tr₃ collector prevents excessive current flow via Tr₄, Tr₃.

- Alternative coupling methods including pulse transformation, light-emitting diodes, etc., may be used if thyristor is at an inconvenient potential relative to firing circuit.

Further reading

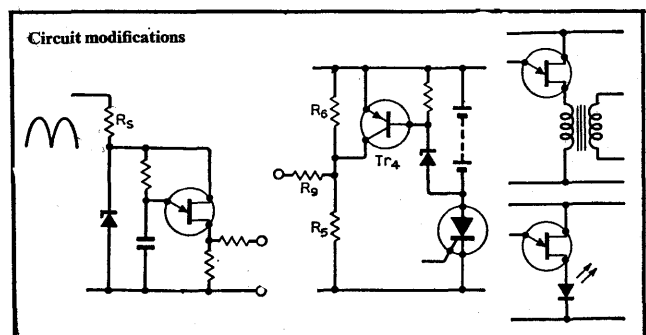
Low-cost constant-current battery charger with voltage limiting, *Semiconductors* (Motorola), vol. 3, 1972, no. 1, pp. 15/6.

400V constant-current source, *Electronic Circuit Design Handbook*, Tab. 1971, p. 298.

Nowicki, J. E., *Power Supplies for Electronic Equipment*, vol. 2, Leonard Hill, 1971, pp. 182-93.

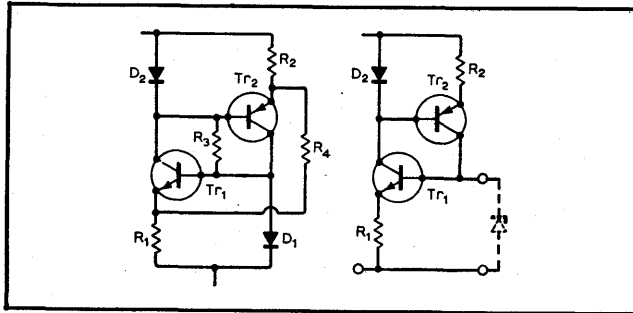
Cross references

Series 2, card 5. Series 3, card 4. Series 6, card 7.



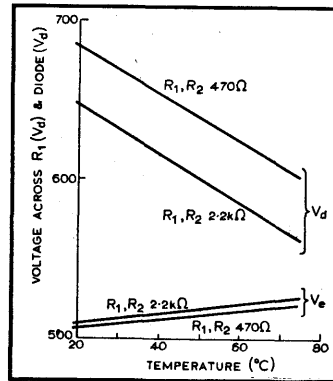
Wireless World Circard Series 6: Constant-current circuits 9

Low-voltage current regulators



Typical performance

D₁: 1S130
 D₂: 1S130
 Tr₁: 2N1308
 Tr₂: 2N404
 R₁, R₂: 220Ω
 R₃, R₄: ∞. Typically leakage current of Ge transistors sufficient for self-starting. To increase dynamic resistance R₄ may be in range 100R₁ to 1000R₁.



Circuit description

The ring-of-two reference (card 5) may be adapted for very low voltage applications by replacing the zener diodes by forward-biased silicon diodes or any other element having dynamic resistance less than static resistance ('amplified' diodes, asymmetric voltage-dependent resistors, gallium arsenide diodes, etc.). The transistors used must then have a V_{be} less than the diode forward voltage drop, and germanium devices are indicated for use with silicon diodes. For optimum temperature compensation with these devices, the p.d. across each emitter resistor should be around 420mV (a figure based on the junction properties of the devices). This is not always convenient to achieve, but stability of 0.1%/deg. C is normally possible. Leakage currents of the Ge transistors are enough to ensure start-up in most cases and R₃ may be dispensed with. Resistor R₄ may be added to neutralize the effect of R₃ if present, and if absent to control the dynamic resistance of the two-terminal circuit. It bypasses current around the transistors reducing the collector current in each, i.e. opposing the natural tendency for a slight increase in current as the terminal p.d. increases. Dynamic resistance may even be made negative and large if R₄ is reduced sufficiently though over a more limited range of supply voltages than normal. This circuit, as with related circuits on card 5, may be used to supply a constant

current to an external zener diode minimizing the total supply voltage required (as compared with its use as a two-terminal circuit interposed between supply voltage and load).

Component changes

D₁, D₂: Any silicon p-n junction including diodes (1N914, etc.) base-emitter junction of transistors (2N3707, BC125, BC126, ME4103, ZTX300, etc.) diode-connected transistor i.e. collector-base short.

Tr₁: n-p-n germanium transistor (OC139, 2N1302, 2N1304, 2N1306, 2N1308).

Tr₂: p-n-p germanium transistor (2N1303, -05, -07, -09, OC42, OC44) for optimum temperature performance with reasonably high gain transistors, diode/transistor combination should result in 400-450mV across emitter resistor.

Circuit modifications

- Diodes may be placed in one limb of the circuit, overcompensating the temperature induced change in Tr₁ V_{be} . By keeping R₁ and R₂ low, resulting decrease in the p.d. across R₁ is insufficient to compensate for the change in the V_{be} of Tr₂. Hence currents in the two limbs change in opposite senses and approximate cancellation is possible. Once this has been achieved, R₁, R₂ may be replaced by a single potentiometer, varying the total current while remaining approximately compensated.

- A different circuit using transistors of only one type is basically a voltage regulator defining the p.d. across a resistor whose current is larger than the remaining circuit currents (similar to card 2). Simplest version defines the current in terms of Tr₁ V_{be} and suffers from variation of current in R₁ as supply varies in addition to temperature dependence ($\approx 0.3\%/deg. C$).

- Replacing R₁ by a junction f.e.t. Tr₃ improves the constancy of current against supply voltage while the introduction of D₁ a germanium diode gives first-order temperature compensation.

- With the penalty of higher terminal p.d. better stability is given by the addition of zener diode D₁. Resistors R₂, R₃ compensate for current variations in R₁ by causing the p.d. across R₄ to fall as the supply voltage rises. Typically R₁ = 10R₄, R₂ = 100R₄, R₃ is varied to optimize slope resistance, but is in the region 0.5 to 5R₄.

Further reading

Williams, P., Low-voltage ring-of-two reference, *Electronic Engineering*, 1967, pp. 676-9.

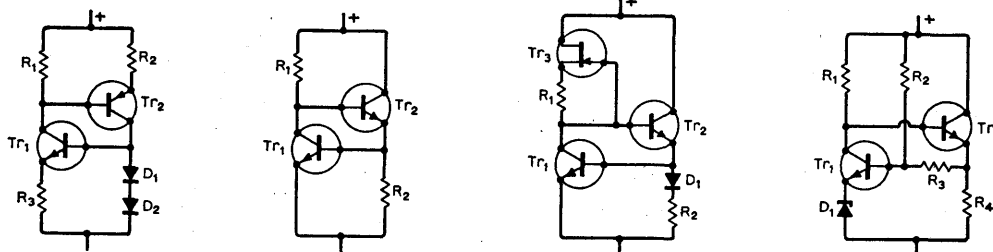
Verster, T. C., Temperature-compensated low-voltage reference, *Electronic Engineering*, 1969, p. 65.

Watson, G., Constant-current circuit, *Electronics*, 6th July, 1962, p. 50.

Cross references

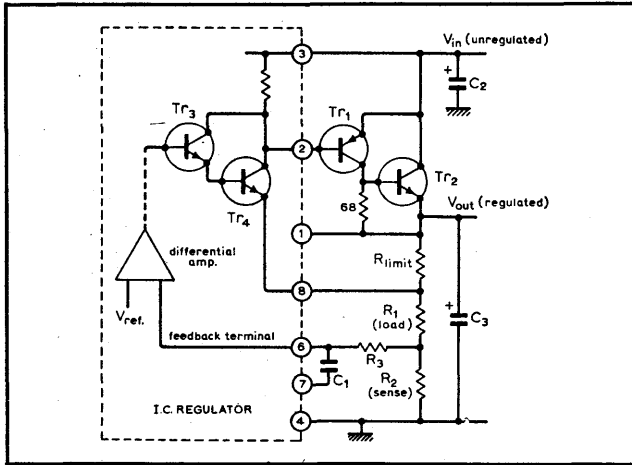
Series 6, cards 3, 4, 5 & 6.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 10

High-power current regulators

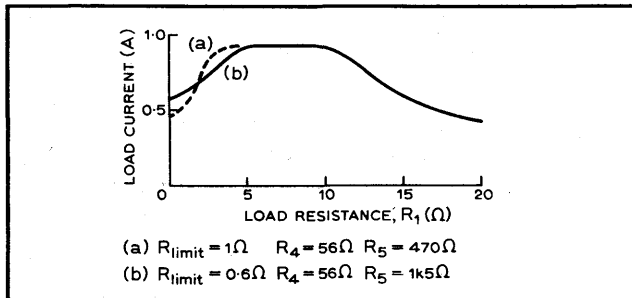


Typical performance

Load current: 0.9A
 Unregulated input: 13 to 20V
 IC: LM300
 Tr₁: BFR81; Tr₂: MJE521
 R_{limit}: 1Ω; R₁ (load): 10Ω
 R₂: 1.95Ω; R₃: 2.2kΩ

C₁: 47pF
 C₂: 1μF (tantalum)
 C₃: 4.7μF (tantalum)

Graph shows effect of foldback current limiting on output current when load R₁ is varied (see circuit over, left)



Circuit description

This is basically a series voltage regulator used as a constant-current source, where the maximum output current depends on the current gain and power rating of the series-pass transistors (Tr₃, Tr₄) connected as a Darlington pair. Further amplification, and thus a greater output current, is available by modifying this series element by connecting two discrete transistors Tr₁ and Tr₂ to give a compound emitter-follower.

The p-n-p/n-p-n combination is preferred for an improved temperature coefficient over a straightforward quad emitter-follower.

The essential function of this regulator is that some fraction of the output voltage (or a voltage due to load current through a resistor) is compared with a reference voltage developed within the i.c. regulator. If the output voltage changes, the error signal is amplified and used to compensate for the original change by modifying the drive to the compound emitter-follower. The internal reference voltage is approximately 1.7V, and hence the feedback sense voltage developed across R₂ must approach this value for the desired load current, thus defining R₂. The resistors across the base-emitter terminals of the external transistors cause the operating currents to be raised and improves the stability.

An arrangement for foldback current limiting is shown over (left) and is used to protect the regulator against the load going short-circuit, and limits the current to around 0.5A under this condition. Capacitor C₁ is a frequency compensation capacitor. The additional current gain necessary for the high current regulators may cause h.f. oscillation, eliminated by connecting a tantalum capacitor across the input and the output.

Component changes

R₁ varies from 1 to 10Ω, current variation within +0.1% over the full range. Regulator may be LM100 or LM305. Tr₁: 2N3055. Tr₂: 2N2905.

Parasitic oscillations can be suppressed by threading a ferrite bead over the emitter lead of power transistor Tr₂.

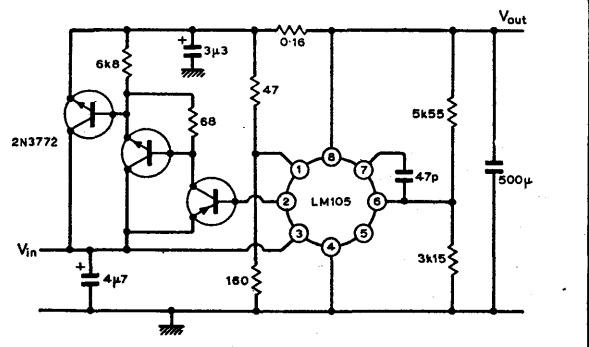
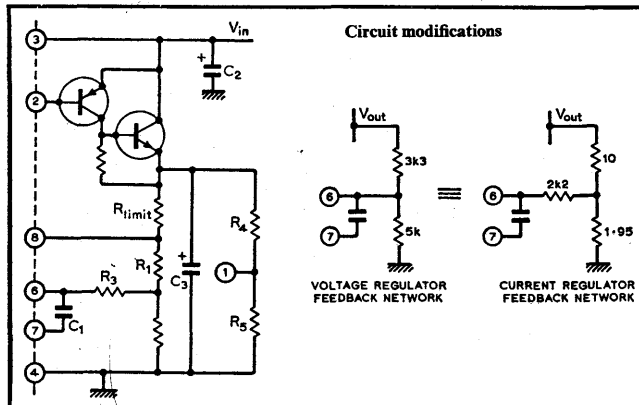
Basic voltage regulator normally has its output voltage set by connecting the tap on a potential divider to the feedback terminal. This explains the values shown for the voltage regulator divider, and need for R₃ when the i.c. is used as a current regulator, the network equivalents being shown over (middle).

Circuit modifications

- Foldback current limiting is achieved by connection of resistors R₄ and R₅ (left). This provides protection for the regulator against excessive power dissipation should the load short-circuit, and limits the current to about 0.5A. Limiting starts when the voltage across terminals 1 and 8 exceeds +0.4V, and depends on the potential differences across R_{limit} and R₄. This critical voltage increases the positive bias on a transistor which therefore conducts harder and steers current away from the first transistor of the series element, and hence the load current decreases.

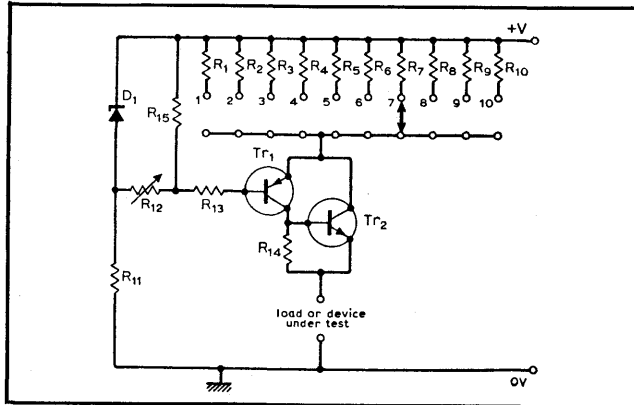
- Very high output currents can be obtained using LM105 or LM305 regulator, and an additional high power transistor. A typical arrangement is shown right to produce 10A, and with foldback current limiting. Input level should be > 9V.

Cross references Series 6, cards 2 & 7.



Wireless World Circard Series 6: Constant-current circuits 11

Constant-current applications



Typical performance

Supply: +12V

Tr₁: 2N3702

Tr₂: BFY50

D₁: HS7062

R₁: 560k Ω ; R₂: 270k Ω

R₃: 100k Ω ; R₄: 56k Ω

R₅: 27k Ω ; R₆: 12k Ω

R₇: 5.6k Ω ; R₈: 2.7k Ω

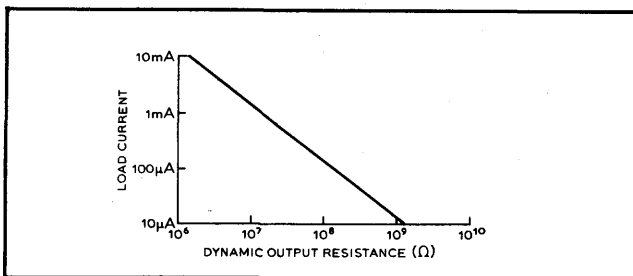
R₉: 1.2k Ω ; R₁₀: 560 Ω

R₁₁: 470 Ω ; R₁₂: 100 Ω

R₁₃, R₁₄, R₁₅: 1k Ω

I_{supply}: 14.5 to 24.3mA.

With load of 1k Ω all preset currents within +8% of nominal values and decade values, e.g. 10 μ A, 100 μ A, 1mA, 10mA within $\pm 1\%$ of each other. Dynamic output resistance/load current: see graph opposite.



Circuit description

A preset constant current may be used in many instrumentation applications in the same way as a preset voltage. Such a current generator may be used, for example, to test semiconductor devices such as diodes and zener diodes to obtain their current-voltage characteristics; in a zener diode the current may change by a factor or more than 100 with a corresponding voltage change of only a few percent. The circuit shown provides constant currents that are preset within the range 100 μ A (S₁ in position 1) to 10mA (S₁ in position 10), with an overall stability of less than 1% at any preset value. The accuracy of the preset currents is not so high as preferred-value 5% resistors were used, but can be improved by using selected values. For diode testing over a wide range of currents, the preset currents are chosen to be multiples of 1, 2, 5, 10 to allow rapid construction of a log-scale graph.

The zener diode D₁ sets the base potential of Tr₁ and hence the p.d. across its selected emitter resistor R₁ to R₁₀. Current in the selected resistor is therefore defined as is the current in the load or device under test. Transistors Tr₁ and Tr₂ form a complementary pair, the equivalent compound transistor having a current gain approximately equal to the product of the individual current gains and an input characteristic equivalent to that of Tr₁. The base current of Tr₁ is thus very much less than the load current so that the latter is virtually the same as that defined in the selected emitter resistor. By selecting the emitter resistor to be R₇, the load current can be set to be 1mA by adjustment of R₁₂. Constant currents of 10 μ A, 100 μ A and 10mA are then also defined to an accuracy, depending on the tolerances of R₁, R₄ and R₁₀ respectively.

Component changes

Larger values of constant current can be obtained by changing Tr₁ and Tr₂ to higher power transistors capable of handling the larger currents. The p.d. available at the load terminals can be increased by using a lower voltage zener diode for a given value of +V. The value of +V can be increased, provided that the breakdown voltage of Tr₁ and Tr₂ is not exceeded, to provide higher load voltages at defined currents. If the Tr₁ biasing network is replaced by a simple potentiometer between the supply lines a high output impedance is still obtained but the load current is less stable and the load p.d. will fall as the load current is increased by altering the potentiometer setting.

Circuit modifications

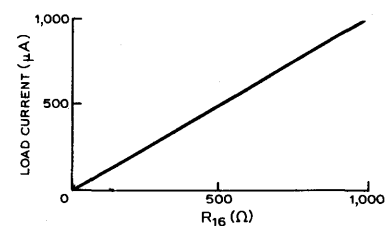
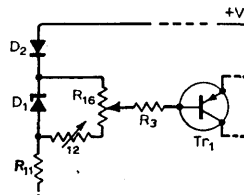
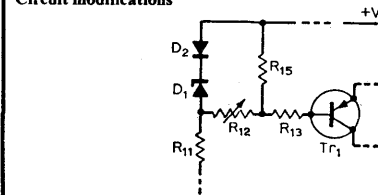
- Errors in the constant currents will be due to drift in the zener diode, drift in V_{be} of Tr₁ and the finite and variable current gain of the compound transistor. In the circuit discussed the zener diode is chosen for low slope resistance to limit dependence on supply voltage. If the circuit is operated from a stabilized voltage supply, the low slope resistance can be abandoned and the zener diode can be chosen to provide best temperature matching. A forward-biased junction diode can then be placed in series with a zener diode to provide temperature compensation for the drift in V_{be} of Tr₁ (see left), where D₁ could be a 5.6V zener and D₂ a BYX22-200.

- In addition to the preset constant currents it is often necessary to provide a current that may be accurately varied over a restricted range. This can be achieved by connecting a potentiometer of the calibrated multi-turn type across the zener diode as shown middle. A graph of the variation in load current achievable using S₁ in position 7 and a 1-k Ω potentiometer is shown right.

- As well as being used for measuring the characteristics of diodes and zener diodes, the unit described may also be used to measure loop resistance by monitoring the load terminal p.d. with a d.v.m. whilst feeding an appropriate constant current to the unknown resistance. By feeding a constant current to the emitter of a transistor and measuring its base current the d.c. current gain can be quickly found. Another application is in electrochemical plating.

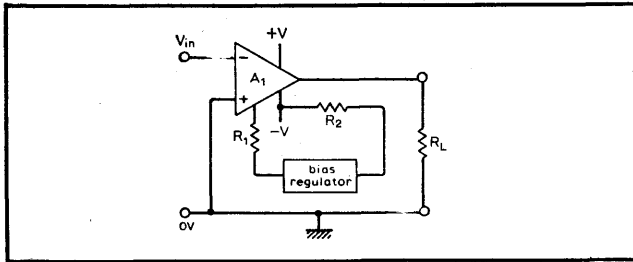
Cross references Series 6, cards 1 & 2.

Circuit modifications



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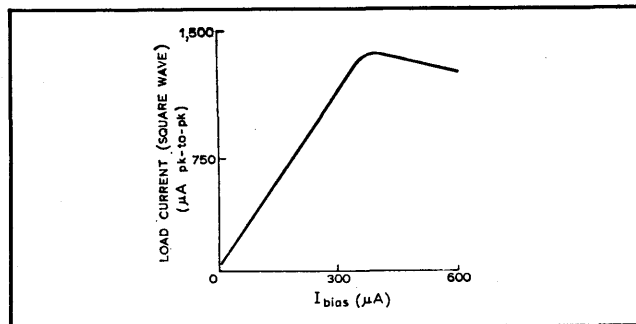
Constant-current amplifiers



Typical performance

Supplies: $\pm 6V$
 $A_1: \frac{1}{3} \times CA3060$
 (regulator is part of CA3060)
 $R_1: 53.7k\Omega \pm 1\%$ for $I_{bias}: 100\mu A$
 $R_2: 47k\Omega$ for I_{bias}

$\leq 100\mu A$; $R_L = 1k\Omega$
 Equivalent source resistance with $I_{bias} = 100\mu A$ is approx. $264k\Omega$
 i.e. load current changes by about 4% for a 1000% increase in R_L .



Circuit description

A class of monolithic amplifiers is now available called operational transconductance amplifiers. This type of amplifier is a novel circuit having similar general characteristics to an operational voltage amplifier except that its gain is better described in terms of a transconductance rather than as voltage gain. Its open-loop voltage gain is equal to the product of its transconductance and the load resistance it feeds.

In the circuit A_1 is one of three transconductance amplifiers in a single package together with a bias regulator. The regulator is supplied from the $-V$ rail through a resistor R_2 and each of the class-A push-pull transconductance amplifiers are biased independently by a suitable resistor R_1 . The transconductance of the amplifier is controlled by the bias current i.e. by the value of R_1 . For a given input voltage between the inverting and non-inverting inputs the output current is defined by the bias current which can be varied over a wide range.

While the amplifier can be used in its linear mode with various

feedback arrangements, the open-loop circuit shown above can deliver a square wave current to the load resistance. The peak-to-peak amplitude of the square wave is under the control of the bias current. As the amplifier has a high output impedance, it may be thought of as being a generator of a current square wave having a definable and constant peak-to-peak value. The circuit can supply an output of around 1V pk-pk into loads of around $10k\Omega$ with an equivalent source resistance of about $260k\Omega$, provided V_{in} is large enough.

Component Changes

Useful range of supply: ± 2.5 to $\pm 7V$
 Maximum differential input voltage: $\pm 5V$
 Maximum d.c. input voltage: $+V$ to $-V$
 Useful range of bias current approx: $10\mu A$ to $2mA$
 Maximum bias regulator input current (total for 3 amplifiers): $-5mA$
 Useful frequency range for square wave output current is typically $120kHz$.

Circuit modifications

An amplitude-modulated constant-current source is obtained if the modulating voltage source is connected as a floating source in series with R_1 or as a grounded source to the bias terminal through a resistance of the order of $100k\Omega$. In the first arrangement 100% amplitude modulation of the output square wave is obtainable, whereas the latter connection provides about 30% modulation depth using a 12V pk-pk sine wave source.

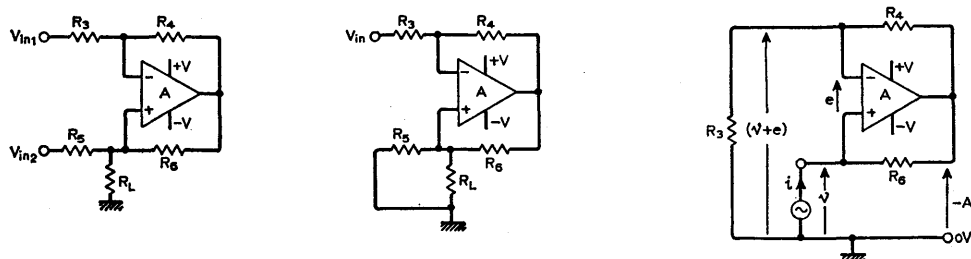
Circuit left shows the general form of a circuit, known as the "Howland" circuit, which provides a constant current into the load by virtue of the fact that A , R_4 and R_5 act as a negative impedance converter. As shown, V_{in2} must supply the short-circuit load current, therefore the circuit is often used in the form shown centre. The high output impedance available at the load terminals can be seen by reference to the diagram on right where R_L has been replaced by a voltage source, V_{in} has been set to zero and R_5 temporarily removed, for analysis.

The output impedance at the load terminals is $Z_o = Z_p // R_5$ where $Z_p = v/i$. For simplicity, let $R_3 = R_4 = R_5 = R_6 = R$, then $-Ae = 2(V + e)/R$. Hence $e = -2V/(A + 2)$ and $i = (V + Ae)/R = v - [A \cdot 2V/(A + 2)]$. Thus $Z_p = V/i = (A + 2)R/(2 - A)$ and $Z_o = Z_p // R = R(A + 2)/4$. Therefore, as $A \rightarrow \infty$ $Z_o \rightarrow \infty$ and a constant current may be fed to R_L .

For an operational amplifier of the 741 type, $A = -jA_o f_o / f$ where A_o and f_o are typically 10^5 and $10Hz$ respectively. In this case $Z_o \approx -jA_o f_o R / 4f$ or $Z_o \approx -j\omega C$ so that Z_o consists of a capacitor $C \approx 2/\pi f_o A_o R$. For $R = 10k\Omega$, $C \approx 64pF$. Thus, the constant load current will be 3dB down w.r.t. its low frequency value at $f = 1/2\pi C R_L \approx A_o f_o R / 4R_L \approx 250R/R_L$ (kHz) for a 741-type operational amplifier.

Cross references Series 6, cards 4 & 6.

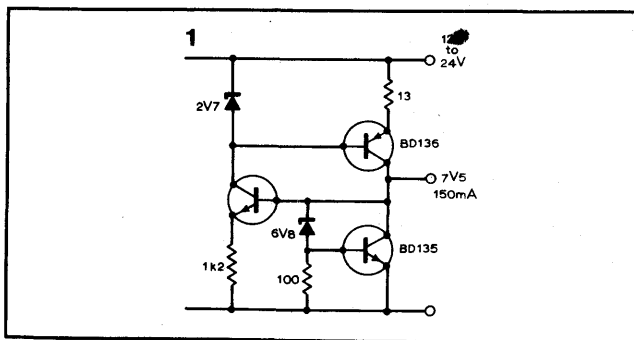
Circuit modifications



Constant-current circuits

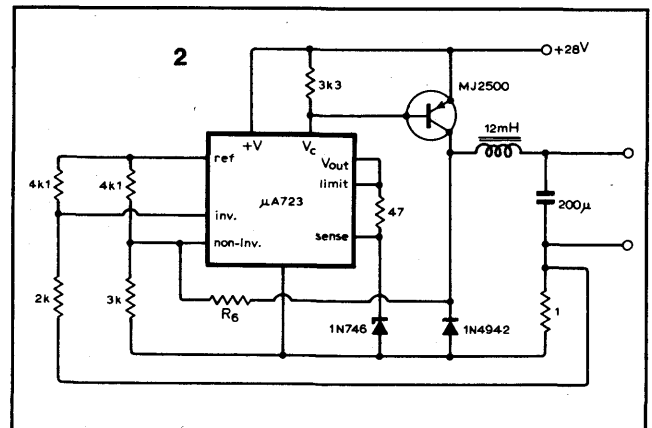
1. The ring-of-two reference has a complementary pair of transistors driving constant currents into two voltage reference elements such as zener diodes. At higher current levels, one of the zener diodes is replaced by a zener/transistor combination acting as a shunt regulator in which the zener carries only the base current of the n-p-n transistor. Although the circuit as shown is for the regulation of an output voltage, such circuits can be used as two-terminal current regulators i.e. the dual of power zener diodes. The minimum terminal p.d. can be reduced by replacing the 6.8V zener diode by a forward-biased diode or, with further-reduced stability, replacing zener BD135 and 100Ω resistor by a single 15Ω resistor.

Hibbert, G. Avoiding power-supply hum, *Wireless World*, 79, 1456, p.515, (Oct. 1973).



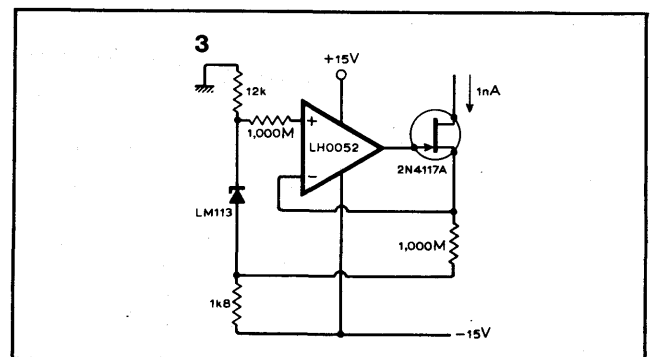
2. Monolithic voltage regulators are readily adaptable to act as switching regulators or to provide a constant current action; the circuit shown combines these functions. The circuit is a form of astable in which R_6 is selected to give a hysteresis of 28mV at the non-inverting input. The internal reference voltage of 7.15V is reduced by the potential divider to give 3V at the non-inverting input i.e. the hysteresis is 1% of the mean value. The network feeding the inverting input sets the p.d. across the 1Ω resistor to 1V, and the mean current in the load to 1A. The switching mode ensures high efficiency regardless of load p.d. making it a suitable circuit for battery charging of one or more cells. The LC circuit at the output filters the switched output for the power transistor and the switch frequency may rise to 20kHz.

Summer, S. E. Switching regulator produces constant current output, *Electronics*, March 7, 1974, p.114.



3. To control very low currents with high accuracy both the sensing and control elements must have input currents much less than the current to be controlled. For currents at the nA level this implies input currents of order 10pA if accuracies of 1 or 2% is required. The circuit shows one way of achieving this in which a f.e.t. input operation at amplifier compares the p.d. across a low-voltage reference diode with that across a very-high-value resistor. If the f.e.t. is selected for minimum gate leakage current the drain and source currents are equal and the output current is defined at about 1.2nA for the given diode.

Underwood, R. K. New design techniques for f.e.t. op-amps, National Semiconductor application note AN-63, 1972, p.9.



Wireless World Circard

Series 7: Power amplifiers

For the most part, these cards discuss circuits for audio-frequency amplifiers, including direct-coupled circuits, though r.f. amplifier circuits are included; video amplifiers are to be treated separately. The article gives a good introduction to power amplifiers and their configurations. Efficiencies and power output for different kinds of output stages are derived for class A and B circuits, and the brief survey concludes with a discussion of class C and D amplifiers for completeness.

Two class A and two class B circuits are described for output powers up to about 5 watts. The general-purpose wideband amplifier, with its 10-octave bandwidth up to 16MHz, is a useful circuit to have to hand. The other r.f. one—a class C circuit providing 1.4 watts at 7.2MHz—is more specialized and it was not possible to give design procedure on the card for other levels and frequencies (but refer to further reading).

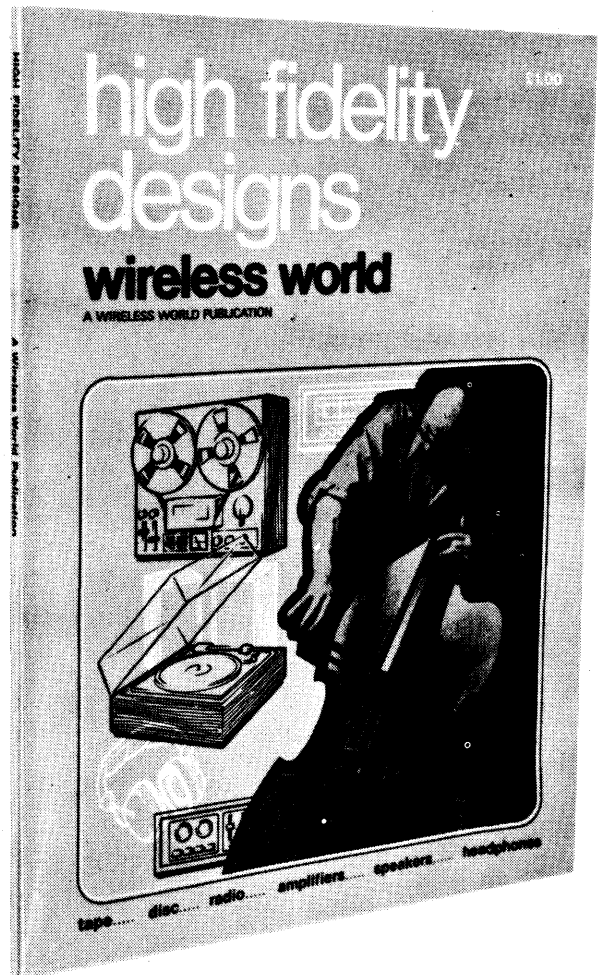
The broad usefulness of the set is aided by cards describing a pulse amplifier, a pulse-duration modulated output stage, bridge amplifier techniques and a 100-volt amplifier circuit.

Basic power amplifiers	1
Servo amplifier	2
Pulse buffer amplifier	3
Push-pull class A power amplifier	4
High-voltage amplifier	5
Class C power amplifier	6
Bridge output amplifiers	7
Class B quasi-complementary output	8
Broadband amplifier	9
Class A op-amp power booster	10
DC power amplifier	11
Class D switching amplifier	12

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Power amplifiers

All amplifiers are power amplifiers in that the power delivered to the load is greater than that drawn from the source. Few are power amplifiers in the same sense that an operational amplifier with feedback may be said to be a voltage amplifier or a current amplifier. Thus in Fig. 1 the load voltage is defined for a given input signal and the load power is proportional to the conductance of R_L . For Fig. 2 feedback defines the load current while the corresponding power developed in the load is proportional to the resistance of R_L .

This suggests that as many power amplifiers use shunt-derived feedback to define their output characteristics, they could properly be regarded as voltage amplifiers which just happen to be capable of delivering large powers to a load of sufficiently low resistance. Operating these amplifiers from a constant supply voltage in the usual way fixes an upper limit on the load voltage. Practical imperfections in the transistors together with current limiting resistors or other protective circuitry reduce this upper limit but still leave the peak output voltage broadly proportional to the supply.

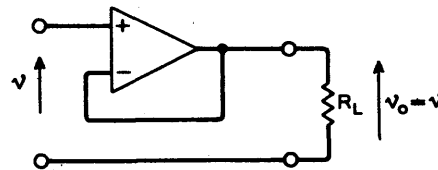
Output power depends equally on the maximum current that can be supplied to the load. The mean value of this power over a complete cycle for a sinusoidal output voltage and resistance load is $\hat{V}\hat{I}/2$, where \hat{V} and \hat{I} are the peak instantaneous values of voltage and current, and as $\hat{V} = \hat{I}R_L$ and $V_{rms} = \hat{V}/\sqrt{2}$, alternative expressions are $V_{rms} \cdot I_{rms} = \hat{V}^2/2R_L = I^2R_L/2 = V_{rms}^2/R_L = I_{rms}^2R_L$.

At this point, you might be asking: "what about r.m.s. power?". This, unfortunately, is on a level with the equivalent enquiry after the well-being of the workers. It can be answered in various ways none of which are useful. To interpret it properly it must be realized that, while the power developed in the load varies from instant to instant, it is the mean or average value of that power that determines, for example, the loudness of the sound produced by a given loudspeaker. The r.m.s. value of the power can be defined mathematically in the same manner as the r.m.s. value of the voltage, but it has no comparable physical significance, i.e. it would require the instantaneous power to be squared, integrated to obtain the mean value of the "power squared", and then the square root taken of the result. The confusion arose because mean power happens to equal the product of r.m.s. voltage and r.m.s. current for certain specified conditions which commonly occur. Hence the "r.m.s." term has become firmly attached to the power measurement itself particularly in the advertising for audio equipment. It should be detached.

While the voltage term in the output

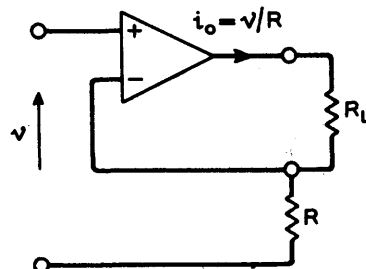
power is fixed by the supply voltage, the current term is a property of the amplifier. Consider first the amplifiers based on a single transistor as shown in Fig. 3. In each case the transistor is assumed to be dealing with an a.c. signal which has equal positive and negative magnitudes, e.g. a sine wave. Thus the transistor must be biased to some quiescent voltage/current setting which will allow positive and negative-going output swings. If distortion is to be avoided the transistor must remain conducting throughout the cycle, i.e. neither the current nor the p.d. across the transistor shall fall to zero. This mode of operation, class A, may be defined in terms of the "angle of conduction", being the full 360° of the cycle. In class B each device conducts for precisely 180° or half the cycle and in class C conduction is for <180°.

In Fig. 3 (a) if the direct current is permitted to flow in the load, equal positive and negative excursions occur for a voltage across the transistor equal to half the supply voltage (assuming an ideal transistor). Thus the peak of the a.c. component of load voltage is $\hat{V} = V/2$. Hence the a.c. power in the load is $(V/2)^2/2R_L = V^2/8R_L$. In the absence of signal the current drawn from the supply is $V/2R_L$ giving a supply power $V^2/2R_L$. This shows an efficiency of 0.25. The quiescent power splits 50/50 between transistor and load resistance.



$$P_o \propto 1/R_L$$

Figs. 1 and 2. In a voltage amplifier and a current amplifier, load voltage or current is defined for a given input signal and load power is proportional to the conductance of R_L (Fig. 1 above) or the resistance of R_L (Fig. 2 below), suggesting that power amplifiers using shunt-derived feedback could be looked on as voltage amplifiers, but capable of handling large powers.



$$P_o \propto R_L$$

It is possible to do still worse. In Fig. 3 (b) the load is capacitively coupled to the amplifier to eliminate the direct current in the load. A collector resistor is still required to allow the flow of current in the transistor, establishing the quiescent conditions for class A operation. Now the total a.c. power is split between R and R_L and the maximum efficiency is reduced to 0.125.

The situation can be retrieved if the collector resistor can be replaced by some constant-current stage as in Fig. 3 (c). The positive peak current in the load can then equal the quiescent current even when the collector approaches the supply voltage (assuming a constant-current stage that can function with a p.d. falling towards zero). Hence the load can have a maximum current swing simultaneously with a maximum voltage swing. In Fig. 3 (b) when the transistor current falls to zero, R and R_L are effectively in series and the p.d. across R limits that across R_L .

The constant-current stage may consist of an inductor whose reactance is high compared to the resistance of the load at all frequencies of interest; an ideal transformer that also allows the use of arbitrary load resistance by proper choice of turns ratio; a transistor biased to deliver a constant current. The disadvantage of (d) and (e) is that the amplitude-frequency response is limited unless bulky and expensive inductor/transformers are available. They do offer the possibility of higher efficiency than any of the other circuits. For example, Fig. 3 (f) allows the peak current in the load to equal the quiescent current, and the peak voltage to equal the transistor quiescent voltage, i.e. half the supply voltage for the best operating conditions. Thus load power is $(I_{dc}V/2)/2$ while supply power is $V \cdot I_{dc}$, giving an efficiency of 0.25 bringing the efficiency back to the level of Fig. 3 (a) but with the d.c. component removed from the load. In Fig. 3 (d), the peak current in the load is still equal to the quiescent current, but the inductance allows the collector voltage to swing positive with respect to the supply line as the transistor current falls—a load peak voltage equal to the supply voltage being available. The a.c. power in the load then becomes $V_{dc}/2$ for the same supply power $V I_{dc}$, and maximum efficiency is 0.5. This is the highest efficiency possible in class A and the transformer-coupled circuit of Fig. 3 (e) has the same capabilities. It is common for practical circuits using small transformers to have efficiencies in the region of 0.25 to 0.4.

The low efficiencies attained by these single-device circuits lead to the investigation of multiple transistor circuits. Simply operating transistors in parallel may increase the quiescent power they can dis-

sipate and hence the available output power, but the method offers only second-order improvements in efficiency by reduction of saturation voltage etc. Before turning to other classes of amplifier, consider the natural extensions that are possible of the circuit of Fig. 3 (f). Replace the transistor used as a current source by one receiving a signal as in Figs 4 (a) to (d). The signals to the two transistors depend on the configuration used, but the aim in each case is to cause one transistor to decrease its current by the same amount as the other increases it, still assuring class A operation for each transistor individually. In this way the peak current in the load may equal twice the circuit quiescent current: at the point when the current in one transistor reaches zero the other has doubled. The peak load voltage remains at half the supply voltage when the output is biased to the supply mid-point for maximum undistorted output.

So far the term "matching" has been avoided. For low-level amplifiers, input impedance is frequently matched to that of the source; this is the condition for minimum noise. To maximize the power gains of following stages the output and input impedances may be matched, i.e. made equal. This remains the practice in r.f. amplifiers, but at lower frequencies deliberate mismatch is more common as it allows for better control over the gain.

A fallacy that is based on the experience with these low-level stages, and derives from the maximum power-transfer theorem is often extended to power amplifiers. For each of the class A stages described earlier there is a value of load resistance that maximizes the output power without clipping the waveform peaks. As shown for each individual case there are separate limits for both peak voltage and peak current, and the optimum load will have a resistance given by the ratio of these peaks. This load resistance has nothing to do with the output resistance of the transistor.

Consider Fig. 5 which represents the operation of Fig. 3 (d). Draw the load line representing the load resistance through the quiescent point; the maximum output power without distortion is achieved for the slope giving symmetrical voltage and current swings in the positive and negative directions. Such a line intersects the V, I axes at $2V_c$ and $2I_c$ respectively and the slope of the line is the same as that joining the V_c, I_c points on the axes. This optimum load is thus confirmed as depending on the quiescent conditions with the slope of the tran-

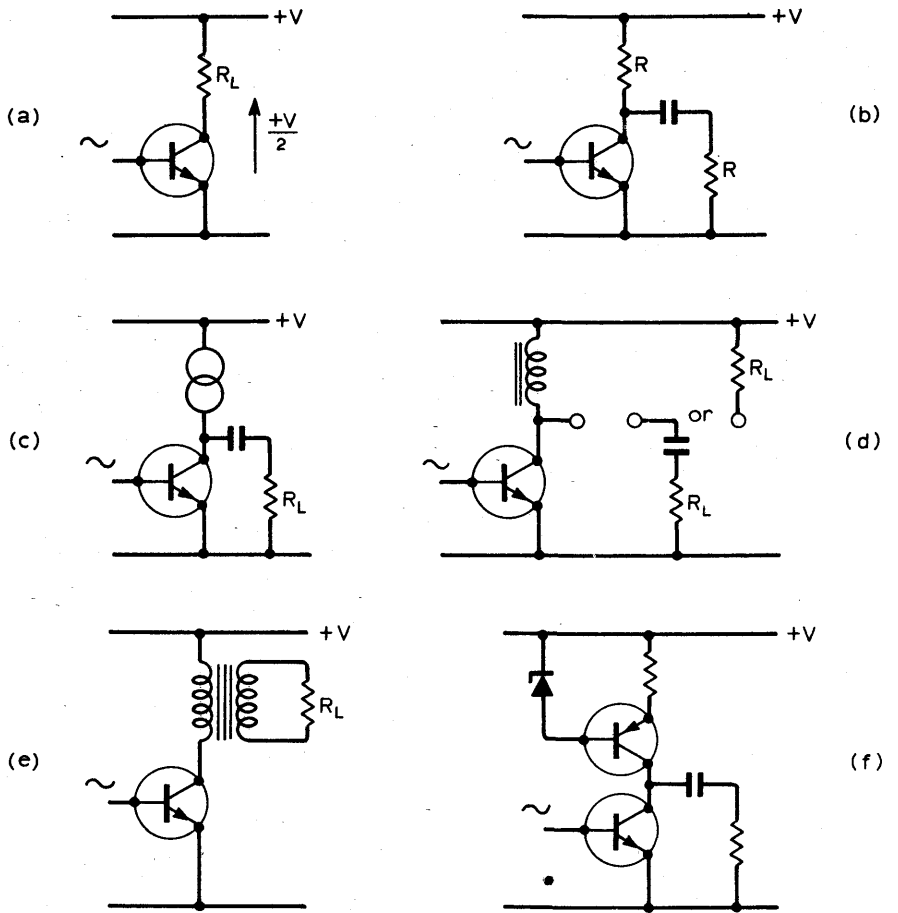


Fig. 3. In a resistive-load class A amplifier (a) efficiency is only 25%; with a capacitively-coupled load efficiency could be 12½% for equal values of resistor shown (b). Replacing collector resistor by a constant-current circuit (c) means the peak load current can equal the quiescent current. Examples of constant-current circuits are a simple inductor (d) or transformer (e) both giving maximum possible efficiencies of 50%, or at (f) using an additional transistor, efficiency 25%.

sistor characteristics (the true output resistance) playing no part. Life is rarely that simple in practice, and the results are modified by saturation effects as well as by the various non-linearities, but not sufficiently to disturb the principle, which applies equally to the circuits of Fig. 4.

Hence for Fig. 4 (c) when used as a class A amplifier the quiescent current (I_s) may be calculated from the supply voltage (V_s) and the intended load resistance. Peak voltage in load is $V_s/2$; peak current in load is $2I_s$; therefore optimum load resistance is $V_s/4I_s$. Resulting mean load power is $(V_s/2)2I_s/2 = V_s I_s/2$, corresponding to an efficiency of 0.5 for ideal transistors.

These circuits are not restricted to class A operation. Indeed they are more commonly used as push-pull class-B amplifiers in which the bias network (not shown) is adjusted to bring both transistors to the edge of conduction. Each transistor then conducts during one half-cycle, there being no quiescent current. There is no comparable limit to the peak current that may be provided; class B simply demands that conduction takes place in a device for 180° in the cycle. A limit will be imposed in any particular design by the current/power limitations of the transistors/power supply. In principle any basic design for a class B power amplifier using configurations such

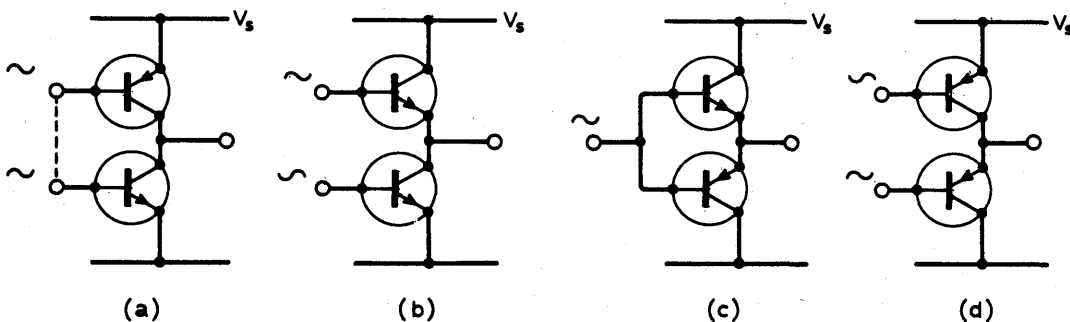


Fig. 4. Using a second for signal handling enables the peak load current to equal twice the quiescent current, with efficiencies of up to 50%.

as those of Fig. 3 (c) and (d) may be extended to higher current levels by replacing the output transistors with Darlington pairs, complementary pairs etc. Thus 100W and 100mW amplifiers may be surprisingly similar in configuration. At high power levels the importance of protection and of maintaining stable bias leads to the addition of circuits monitoring and/or controlling the current in the output stage.

To minimize the distortion due to device non-linearity at low currents (cross-over distortion) the bias networks are set to provide some quiescent current, setting the operation intermediate between true class B and class A—often called class AB and further subdivided into AB₁, AB₂ according to the fraction of the cycle for which each device is non-conducting. The design of low-distortion power amplifiers is a highly specialized subject that will warrant separate treatment in a later series though outline design procedures and practical examples of simple and economical circuits are given in this series of Circards.

Quiescent power in class B is zero. Maximum output power with ideal transistors (Fig. 6) is $\hat{V}^2/2R_L$ where \hat{V} is $V_s/2$. Therefore $P_{L,max}$ is $V_s^2/8R_L$.

Under these conditions, the mean current drawn from the supply is $V_s/2\pi R_L$. This is because the current is drawn from the supply only during the positive half-cycle; the negative half-cycle results in charge being withdrawn from the large coupling capacitor, which charge is restored during the next positive half-cycle. The mean power drawn from the supply is $V_s I_{dc} = V_s^2/2\pi R_L$ and the corresponding efficiency is

$$\frac{P_L}{P_s} = \frac{V_s^2 \cdot 2\pi R_L}{8R_L \cdot V_s^2} = \frac{\pi}{4}$$

or 78.5%.

As the load power is proportional to the square of the output voltage while the supply power is proportional to the voltage it follows that efficiency is proportional to output voltage. It is also true that at some intermediate level of output, the load power having fallen faster than the supply power, the power in the transistors passes through a maximum. For sine wave drive the maximum dissipation in each transistor is at an output voltage where \hat{V} is V_s/π , and the dissipation is then about one fifth of the maximum output power derived above, i.e. a 10W amplifier could theoretically be constructed using a pair of transistors with power ratings of only 2W each.

Class C amplifiers are normally restricted to tuned amplifier systems, conduction taking place for only a small part of a cycle, and the recovery of a sinusoidal waveform then demands a high-Q circuit. The exception is the power-control circuits such as controlled rectifier and triac circuits in which only the mean value of voltage/current/power is of interest and waveform shape is non-critical. These differ from conventional class C circuits in that the conduction has a controlled starting angle but always finishes near the end of a half-cycle; class C r.f. amplifiers are biased such that

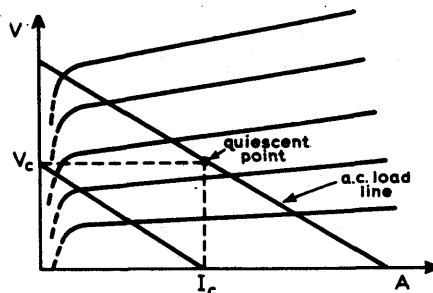


Fig. 5. Optimum load for maximum signal handling is represented by load line cutting axes at $2V_c$, $2I_c$, for Fig. 3 (d) or (e) and depends on the quiescent condition and not transistor output resistance. In general optimum load can be calculated from peak voltage/peak current.

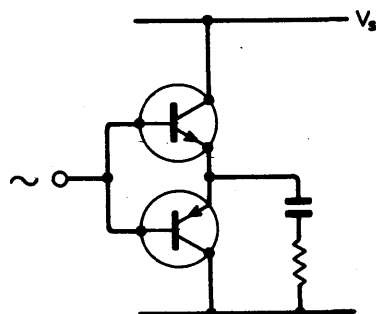


Fig. 6. In this typical class B stage, text shows maximum efficiency to be 78.5%.

the conduction angle is symmetrical about the peak of the drive waveform.

A further difference is that the power control devices are operated as nearly as possible as perfect switches, while at the high frequencies normally associated with class C stages, a very detailed design procedure is required to cope with transistor parameters. This will normally include complex conjugate matching to source and load, to optimise performance. Efficiency can exceed that for class B, though power losses in the passive components involved in the tuning/matching processes are inevitable. A further application of class C power amplifiers is in frequency multiplication where the output circuit is tuned to a harmonic of the input frequency. These aspects are germane to more detailed later studies of r.f. circuits.

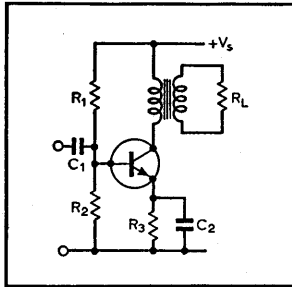
Class D is the generic term for switching circuits in which the active devices are switched multiply in and out of conduction during a single cycle of the input signal. They are also power realizations of pulse modulation systems, the theory for which can be used to determine the spectral components of the output. As one example, circuits such as those of Fig. 4 may have their signal drive replaced by high-speed pulse waveforms whose widths are modulated by the received signal. If the load is fed via an LC filter, the pulse frequency terms can be removed and the output transferred to the load is proportional to the signal.

For ideal transistors the switching pro-

cess allows for zero power dissipation; at all times either the transistor p.d. or its current tend to zero. If the unwanted terms are to be well outside the band of frequencies that it is required to amplify, then the pulse frequency may have to be so high that serious power losses occur during the instants of switching, while charge storage in either of the transistor base-emitter junctions can lead to excessive current spikes through the series path then provided by the two transistors. The principle is more readily applicable to small servo-motor systems than to audio amplifiers as the electro-mechanical properties of the load do not require very high switching frequencies. In some cases efficiencies may exceed 90% with 100% as the theoretical upper limit.

In all the above circuits an ever-present problem is that of protecting both the circuit and its load from excessive current flow. Much time and energy is expended on systems for protection against faults, but inevitably accidents happen, so often after some improvement or embellishment has been added. One can only wonder if such thoughts may have been in the mind of William Wordsworth when he wrote
I have submitted to a new control;
A power is gone which nothing can restore;
A deep distress hath humanized my soul.
A cry from the heart that will speak to all designers of power amplifiers.

Basic power amplifiers



Typical data

Supply: +15V
 Tr: BFY50
 R_1 : 1.2k Ω
 R_2 : 120 Ω
 R_3 : 10 Ω
 T: 3.25:1 turns ratio
 Quiescent current: 70mA
 Output power into 25 Ω
 load: ~400mW for 10%
 distortion

Class A

The classic transformer-coupled class A amplifier has been superseded for most purposes, but may still be applied where good isolation is required between source and load, or where the optimum impedance for maximum undistorted output is very different from the load impedance. Resistors R_1 and R_2 fix the base potential of Tr_1 provided the current through them is much greater than the base current. This base current is the required collector quiescent current divided by transistor h_{FE} . These parameters fix the value of $R_1 + R_2$ by the approximate relationship $R_1 + R_2 = h_{FE} V_s / m I_C$. The value of m the ratio of divider current to base current, is a compromise between stability and wasted power. Typically $m = 5$ to 20.

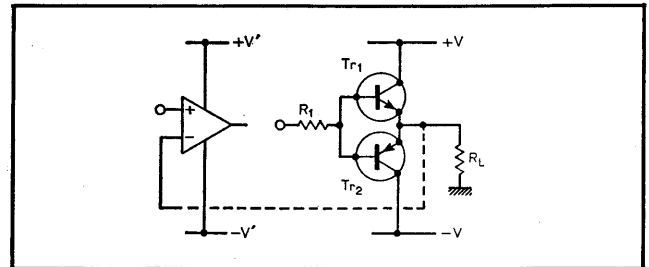
Emitter current (and hence I_C) is defined because the p.d. across R_3 equals the p.d. defined across R_2 minus the V_{be} of Tr_1 . For silicon transistors this is 0.6V and is stable to within 10 or 20% for most transistors under most operating conditions. The resulting p.d. across R_3 is again a compromise between high values for better stability and low values for minimum wasted power – not less than 0.5V and not greater than say 20% of supply voltage as a guide for power stages. Capacitor C_2 decouples R_3 to prevent negative feedback within the required frequency range. As R_3 may be a low resistance, C_2 must then have high capacitance.

Class C

The basic principle behind class C amplifiers is simple, the efficient realization difficult. The transistor conducts only on positive peaks of the input signal with the RC time constant determining the angle in the cycle for which conduction continues, the base-emitter of the transistor acting as a diode and allowing C to charge during the peak. The current in the output circuit is then in the form of pulses of current of which the fundamental term flows in the load if the LC circuit resonates at the fundamental frequency. A high-Q circuit ensures that the harmonics are sharply attenuated giving good output waveform simultaneously with high efficiency. A wide range of load and source impedances can be accommodated by introducing suitable LC networks at input and output (see card 6).

Class B

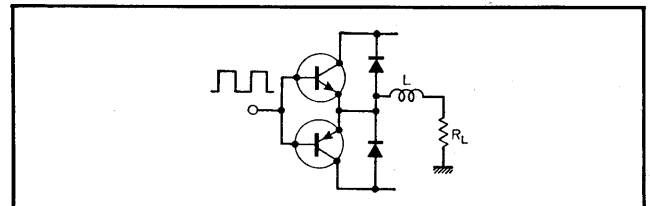
The complementary pair of transistors acting as emitter followers comprise the basic class B push-pull stage. Transistor Tr_1 conducts during the positive half-cycle and Tr_2 during the negative half cycle. For input voltages close to zero neither transistor conducts as each requires a finite base-emitter voltage for conduction (~0.5V for silicon devices). Non-linearities at low-levels make direct voltage drive at the bases unattractive, with the resulting cross-over distortion being very apparent in badly designed amplifiers of this type. If the output stage is included within



Supplies $V = \pm 6V$,
 $V' = \pm 15V$
 Tr_1 : BFR41
 Tr_2 : BFR81
 IC_1 : 741
 R_1 : 470 Ω

R_L : 15 Ω
 Output power 0.92W at:
 64% efficiency
 Output voltage swing:
 10.5V pk-pk for $\pm 6V$
 supply

the feedback loop of a high-gain amplifier the negative feedback reduces the distortion very considerably. At high frequencies the falling gain of the op-amp prevents the feedback from being fully effective and the crossover reappears. Voltage-gain as shown is unity, but standard feedback networks may be used to obtain any desired voltage gain. Output may be increased to 1.75W into 8 Ω but heat-sinking is then advisable. If the objectionable audio effects of crossover are to be minimized biasing networks are inserted between the transistor bases.



Class D

In the class D amplifier, one or more transistors act as switches, connecting the drive point of an LR series circuit to the supply lines. This delivers a square wave to the LR circuit and provided the reactance of the inductor is high at the switching frequency there is little output. If the duty-cycle of the input waveform is altered the output will have a mean level which is a function of the duty cycle. A frequency lower than that of the basic switching frequency is used to modulate the pulse-width/position of the square wave generator and the low voltage is then a function of that signal voltage. For ideal transistors there is no power lost at the switching frequency and the overall efficiency can approach 100%. Diodes clamp the output voltage to the supply lines. The drive voltage must be large enough to saturate the transistors.

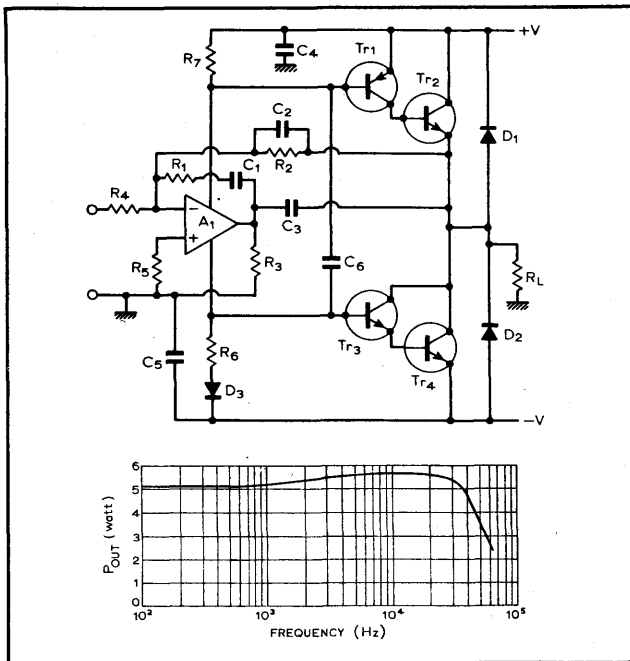
Further reading

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 Stewart, H. E., Engineering Electronics, Allyn & Bacon 1969, pp.589-642.
 Birt, D. R., Modulated Pulse Amplifiers, *Wireless World*, 1963, pp.76-83. (Also subsequent articles and letters.)

Cross references

Series 7, cards 4, 5, 9, 10, 11 (class A), 2, 3, 7, 8 (class B), 6 (class C), 12 (class D).

Servo amplifier



Typical performance

Supplies: $\pm 15\text{V}$, 235mA
 Quiescent current:
 $\pm 1.8\text{mA}$
 A₁: 741
 Tr₁: BFR81
 Tr₂, Tr₄: TIP3055
 Tr₃: BFR41
 R₁, R₂: 15k Ω
 R₃: 47 Ω ; R₄, R₅: 4.7k Ω

R₆, R₇: 180 Ω
 C₁: 2nF; C₂: 100pF
 C₃: 4.7nF; C₄, C₅, C₆:
 470nF
 D₁, D₂: SP2; D₃: 1N914
 R_L: 18 Ω
 Rise-time $\approx 30\mu\text{s}$ (4.8V
 pk-pk at 1kHz)
 V_{in}: 2.07V r.m.s. with-
 out clipping

In servo systems a servoamplifier is needed when a high-power load must be driven from a low-power source. Amplifier A₁ acts as a see-saw amplifier having its gain determined by R₃/R₄ which can be adjusted to accommodate a wide range of input signal levels from a transducer. With no input signal, the output power transistors are virtually cut off, the only drain from the supply being the quiescent current of the operational amplifier (around 2mA). Hence the base-emitter junction of Tr₁ is forward-biased by only about 350mV due to the p.d. across R₇. The base-emitter junctions of Tr₃ and Tr₄ would be forward-biased to a smaller extent unless R₆ was greater than R₇. However, including D₃ and making R₆ = R₇ produces the desired bias with D₃ providing some temperature compensation for the base-emitter voltage of Tr₄. The amplifier has a class B push-pull output stage so that a bipolar input signal produces class B currents in its supply leads. These currents are used to provide the base drive to the compound power transistors which supply the load currents to R_L in push-pull. Transistors Tr₃ and Tr₄ form a Darlington pair while Tr₁ and Tr₂ are its complementary equivalent. The Darlington configuration is used to provide high current gain to ensure that the load current is much larger than the amplifier's quiescent current. To guard against instability, R₁ and C₁ provide feedback around the operational amplifier and R₃ and C₃ provide feedback around the power stage. Bandwidth of the amplifier is controlled by C₂R₂ time constant which can be held fixed when the gain is varied by R₂, if C₂ is also

adjusted. Diodes D₁ and D₂ protect the output transistors against breakdown when the load is highly inductive.

Component changes

Useful range of supplies: ± 6 to $\pm 18\text{V}$.

Output power and efficiency fall as supply voltage is reduced: typically P_{out} is 0.8W and efficiency is 65% with $\pm 6\text{V}$ at 1kHz. With maximum drive, P_{out} falls as R_L increases: for supplies of $\pm 15\text{V}$, typically, P_{out} is 12.6W for R_L = 6.8 Ω and P_{out} = 3.8W for R_L = 25 Ω . Total harmonic distortion falls as drive increases: typically 0.45% for V_{in} = 2.8V and 5.3% for V_{in} = 150mV (supplies $\pm 15\text{V}$, R_L: 18 Ω and f = 1kHz sine wave).

Circuit modification

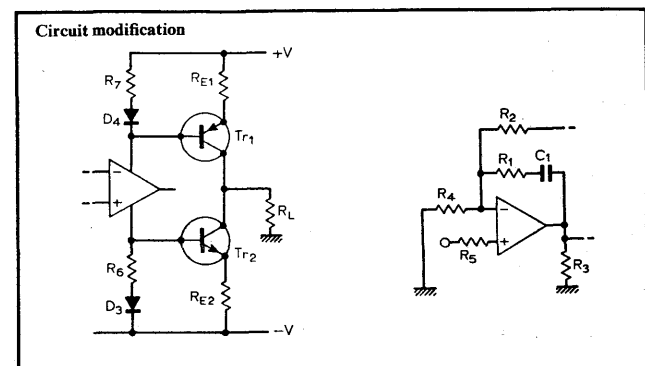
- The Tr₁-Tr₂ and Tr₃-Tr₄ Darlington pairs in the output stage may be made single n-p-n and p-n-p transistors. Ideally, these transistors should have high current gains to provide a peak load current that is significantly in excess of the quiescent current in the amplifier. They also need to have a higher power rating and the combination of high power, high current gain and wide bandwidth is not an easy specification to meet at low cost. The use of single BRF81 and BRF41 transistors provides a reasonable compromise.
- A modification which can improve stability while allowing some quiescent current in the output stage, i.e. biasing in class AB, is obtained by including resistors in the equivalent emitters of the drive transistors, increasing the p.d. across R₆ and R₇ and/or placing a diode in series with R₆ and R₇. The resistors in the emitters can be selected to provide the required quiescent current. (See circuit left.)
- In principle, any other feedback configuration may be used; for example taking the input signal to the non-inverting input of the operational amplifier and grounding the input end of R₄ converts the feedback to a series-applied form with the accompanying increase in input impedance. (See circuit right.) The operational amplifier may be supplied with differential input signals if desired.

Further reading

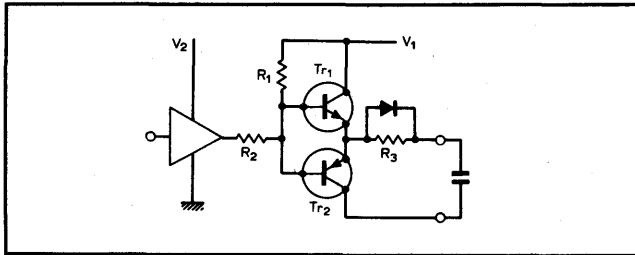
Campbell, D. L. & Westlake, R. T., Build a high-current servoamplifier with i.cs, *Control Engineering*, December 1969, pp.91-4.
 Garza, P. P., Getting power and gain out of the 741-type op-amps, *Electronics*, 1 Feb., 1973, p.99.

Cross references

Series 7 cards 1 & 12.
 Series 2 card 4.
 Series 4 card 8.



Pulse buffer amplifier

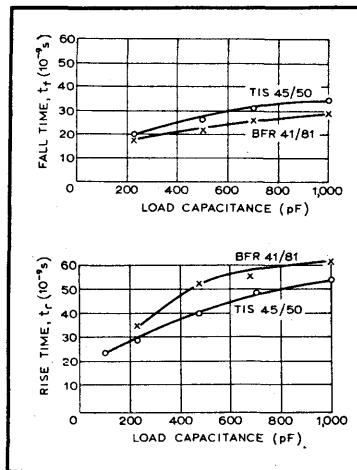


Typical performance

V_1 : +14V; V_2 : +5V
 Tr_1 : TIS45; Tr_2 : TIS50
 IC_1 : 1/6 SN7406
 R_1 : 470; R_2 : 100 Ω
 R_3 : 10 Ω
 D_1 : PS101; C_1 : 680pF
 Input pulse height: 4V
 Duration: 600ns
 P.R.F.: 50kHz
 Rise time: 20ns

Output pulse: rise time 49 ns; fall time 32ns; pulse height: $\approx V_1$ (Rise and fall times measured between 10% and 90% levels). Variation of rise and fall time with several capacitive loads shown right

Some small distortion effects on input drive pulse were not apparent on the output pulse.



Circuit description

The complementary symmetry output stage commonly used in class B amplifiers is equally applicable to pulse outputs. The problem here is that using only a single transistor in the output will only allow any capacitive load to have either a fast rise time or a fast fall time, but not both. Or if the output stage is operated in class A, it needs a quiescent current greatly in excess of the charging current required by the capacitor to achieve a high rate of rise and/or fall. The class B push-pull stage shown has Tr_1 driving the capacitor in the positive direction when a positive-going edge is applied at the base connection, while Tr_2 drives the capacitor in the negative direction. Rise and fall times are now determined by the current flow in the capacitor, which on the positive-going edge is limited by the base current that can be supplied by R_1 as D_1 is allowed to conduct. On the negative-going edge, current

through R_2 is significantly greater and could cause excessive current flow in Tr_2 but the diode is reverse biased and R_3 takes the place of limiting action previously provided by R_1 . It is not possible in a simple circuit of this kind to choose a simple bias network for R_1 and R_2 which would give the same bias drive current in both directions.

IC_1 is an open-collector high-voltage output device which pulls the potential at the bases of Tr_1 and Tr_2 to a low value when in conduction, and when out of conduction allows the bias to rise towards V_1 via R_1 .

Component changes

- Transistors Tr_1 and Tr_2 can be replaced by BFR41 & BFR81 or BC125 & BC126 with poorer rise and fall times. Typical comparison

	rise time (ns)	fall time (ns)
TIS45/50	12	12
BC125/126	28	14
BFR41/81	38	15

- For each capacitor value, overshoot on leading and trailing edges of output pulse is approximately 25% of pulse level.

- Resistive load: 100 Ω , V_1 : 14V, V_2 : +5V; output pulse excursion is from 1.6 to 12V.

Pulse width: 6 μ s. Useful frequency range 3 to 100kHz. Corresponding mean current from supply 1.5 to 30mA d.c.

- IC_1 : SN75451A or SN7407 for greater output voltage levels and faster rise times.

Circuit modification

- Rise and fall times for the circuit left are given centre. The lower level of drive pulse from the i.c. is approximately zero and hence pulse rise times will be slightly larger than in the original circuit.

- An alternative arrangement is shown right. If the drive voltage goes positive, the Zener diode transfers current to the base of Tr_2 which brings Tr_2 into conduction, clamping the output to the negative supply rail, with very small saturation effects. Conversely, if the output swings negative Tr_1 conducts and clamps the output to the positive rail, i.e. the peak-to-peak output swing into the load is almost equal to the supply rail values.

Further reading

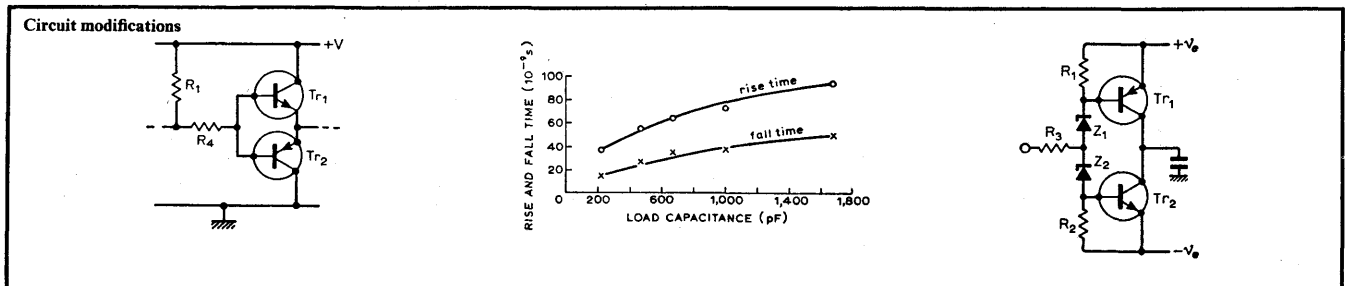
Texas Instruments Technical Seminar 1972, m.o.s. memory drivers.

SGS-Fairchild, Industrial Circuit Handbook, 1967, p.38.

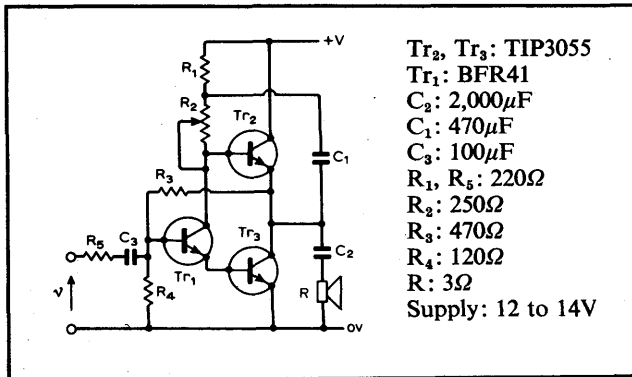
Williams, P., Voltage following, *Wireless World*, vol. 74, 1968, pp.296.

Cross references

Series 6, cards 1, 2 & 8.



Push-pull class A power amplifier



Typical performance

For supply of 13V, quiescent current of 950mA, max. output for 3% distortion is 12V pk-pk into 5 Ω (3.6W). Mean current falls to 820mA at max. output. Full power bandwidth: 20Hz to 100kHz. Hum and

noise: 80dB below full output. Quiescent current: 1.25A @ 13V. Output power: 5W into 3 Ω @ 5% t.h.d. Distortion: <1.1%, 1W into 3 Ω , 100Hz to 10kHz. Voltage gain \sim -2. Input impedance \sim 250 Ω .

Circuit description

Class A push-pull amplifiers have at least two active devices in the output stage, and each device should operate under the same quiescent conditions. A drive circuit using one or more devices provides antiphase signals to the output pair which should have matched parameters. Thus a minimum of three transistors is called for and more are commonly required. By using current phase-splitting, a simple circuit results which still gives adequate efficiency and distortion figures. The key feature of the circuit is that the current in R_2 remains constant throughout the a.c. wave form while its d.c. value can be adjusted to set the desired quiescent current. Bootstrapping via C_1 ensures that any increase in the collector potential of Tr_1 is transferred via the emitter follower action of Tr_2 to reappear at the junction of R_1 and R_2 . Hence the change in p.d. across R_2 approaches zero except at very low frequencies where the reactance of C_1 becomes significant. As there is no change in R_2 current, any increase in Tr_1 current increases the base current of Tr_3 while reducing the base current of Tr_2 by substantially the same amount. Accurate current phase-splitting together with matched current gains of Tr_2, Tr_3 keep the distortion low. Overall negative feedback via R_3 defines

the output quiescent voltage as a multiple of the base voltage of Tr_1 (\sim 1.3V) and the ratio R_3/R_4 scales this base voltage up to half the supply voltage, i.e. the output transistors operate with equal V_{ce} as well as equal I_c .

Component changes

Tr_2, Tr_3 : Power transistors with closely matched h_{FE} at operating current. Quiescent power (at least twice max. output) determines types and heat sinks.

2N3055 for $P_o > 5W$. MJE521 for $P_o > 1W$.

BFY50, BFR41, etc., for $P_o < 1W$.

Tr_1 : BFY50, BFR41, 2N3053 for most applications.

C_2 : Reactance $< R_L$ at lowest freq. Typically 200 to 5000 μ F.

C_1 : Reactance $\ll R_1$ at lowest freq. Typically 100 to 500 μ F.

R_1, R_2 : Set output current $V_s/2(R_1 + R_2) \approx 2I_s/h_{FE}$. One resistor made variable to adjust mean current. Typical range 100 Ω to 1k Ω (higher values for low-power circuits).

R_5 : Sets voltage gain $\sim -R_3/R_5$ and input resistance $\sim R_5$.

R_3, R_4 : Set output voltage (quiescent) to $\sim 2V_{be} [(R_3/R_4) + 1]$. Current in R_3, R_4 to 5 to 20 times base current of Tr_1 . Typical values R_4 : 100 to 500 Ω . R_3 : 300 Ω to 3k Ω .

Circuit modification

- Open-loop gain of the original circuit is low and feedback that can be used may not reduce distortion sufficiently. Simple bias circuit leaves the output at a fixed multiple of V_{be} rather than at the supply centre point, i.e. resistors require readjusting for different supply volts. Adding Tr_4 increases open-loop gain, allows 100% d.c. series-applied feedback and has input feedback and load all referred to same supply line. This eliminates bootstrap capacitor provided speaker can tolerate direct quiescent current of driver stage. For output at midpoint of supply $R_6 \approx R_7$. Voltage gain $\approx (R_3/R_5) + 1$. Reactance of $C_2 \ll R_5$ at lowest frequency of interest. Typically R_4, R_5 : 1 to 10k Ω , R_6, R_7 : 20 to 200k Ω . (circuit left)

- For higher input impedance, input potential divider may be bootstrapped. Interchanging locations of R_5, C_2 allows R_6 to be bootstrapped, almost doubling input impedance (centre)

- Quiescent current depends on current gains of Tr_2, Tr_3 . By monitoring circuit mean current and using result to control drive current Tr_1 , mean current can be made constant, e.g. for Tr_5 a germanium transistor, D_1 a silicon diode, mean p.d. across R_1 is controlled at 0.4V. (right)

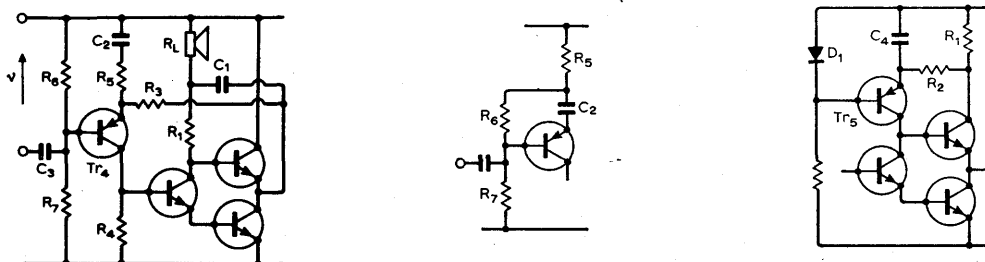
Further reading

Linsley Hood, J. L., Simple class A amplifier, *Wireless World*, vol. 75, 1969, pp.148-53.

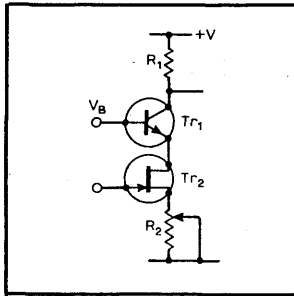
Markus, J. (ed.), Improving signal transfer in Electronics Circuits Manual, 1971, p.19.

Allison, W., Self-biasing class A power amplifier, *Wireless World*, vol. 78, 1972, p.577.

Circuit modifications



High-voltage amplifier



Tr_2 : 2N3810
 R_1 : $10k\Omega$; R_2 : 456Ω
 Quiescent voltage: 52V
 V_B : 10V
 Input signal: 7V pk-pk
 Output voltage: 84V pk-pk
 Gain constant up to 20kHz
 Variation of output with R_2 not decoupled shown opposite.
 Effective output impedance of transistor configuration $5M\Omega$.

Typical performance

Supply: +100V
 Tr_1 : MJE340

Circuit description

The characteristics required by an amplifier may include high voltage gain and in some applications the ability to withstand high output voltages simultaneously. Such a combination is not available within a single device, but the circuit shown arranges that the necessary input impedance gain characteristics are obtained by Tr_2 and the high voltage characteristics by Tr_1 . The input characteristics aimed at were that the device should behave with a defined gain, so that the whole system could be considered equivalent to a valve. Transistor Tr_2 is thus a field-effect transistor whose gain is controlled by the quiescent current, which may be set by R_2 . The drain of Tr_2 feeds into the emitter of Tr_1 whose base is maintained at a constant potential, just high enough to ensure that Tr_2 has a quiescent voltage that is above its pinch-off value. The bias voltage should be obtained from a low impedance circuit. Hence Tr_2 is operating into a low impedance, while Tr_1 is virtually a common-base stage and has thus the highest voltage rating that it could possibly have. The current at the collector of Tr_1 is essentially the same as the emitter current as the current gain from emitter to collector is nearly unity. There is no significant Miller/Blumlein effect between the collector of Tr_1 and the gate of Tr_2 as the voltage swing at the collector is isolated from the gate of Tr_2 . The capacitance between Tr_1 collector and base is now effectively a capacitance to ground rather than to the input of the amplifier. However this capacitance still affects the output characteristics, as it is

in parallel with R_1 for a.c. and determines the bandwidth of the amplifier. The problem is more severe than in many low-voltage amplifiers because R_1 will have a much higher value for a given quiescent current because the p.d. across it may be in excess of 100V. This is the usual penalty to be paid for a high-voltage gain, i.e. the associated high load impedance will have a longer time constant for a given capacitance. The voltage rating is close to the V_{CE} breakdown of Tr_1 .

Component changes

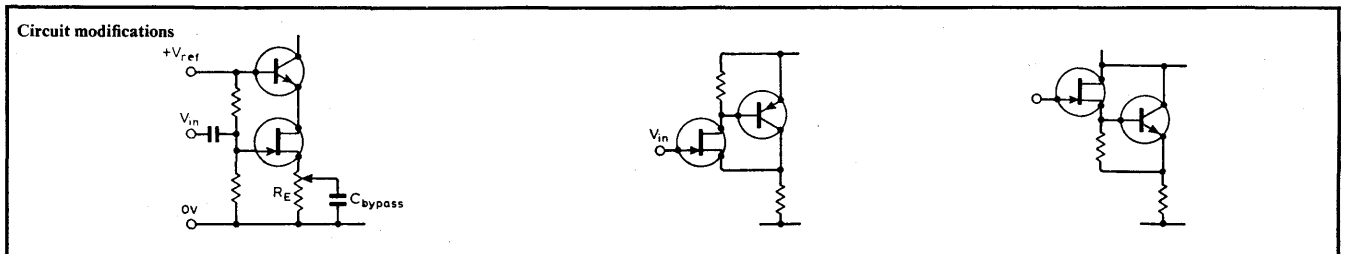
- Decouple R_2 with $150\mu F$ capacitance to retain g_m of the combined transistors. Output 82V pk-pk for an input signal of 2.4V pk-pk. Low frequency cut-off then 5Hz.
- Range of V_B 8 to 11V – value not critical, with no significant effect on performance.
- Supply may be increased up to 300V with appropriate changes in R_1 and R_2 to control quiescent voltage. Typically (i) supply: +200V, V_Q : 110V, R_2 : 122Ω , V_{in} : 7V pk-pk; V_{out} : 180V pk-pk; R_1 : $10k\Omega$. (ii) supply: +300V, V_Q : 150V, R_2 : $1.5k\Omega$, R_1 : $68k\Omega$, V_{out} : 275V pk-pk.
- Increase of +V from 100 to 200V, maintaining circuit resistors constant reduces h.f. cut-off by approximately 20% indicating that this is dependent more on external components rather than operating conditions.

Circuit modifications

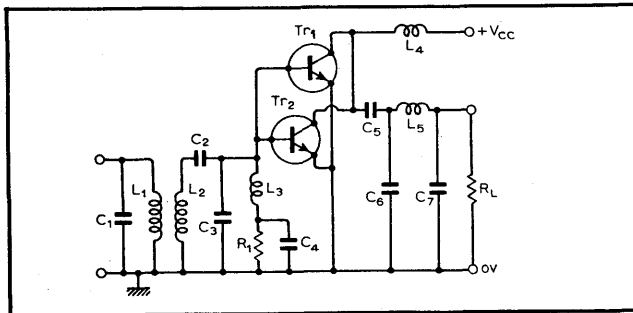
- F.E.T. g_m is controllable by varying negative feedback. A wide range of control is possible with the circuit shown left. Because the gate is positive, R_E can be large for chosen quiescent value of drain current, the feedback being varied via C without then altering the d.c. state of the circuit.
- F.E.T. g_m can be boosted by adding a p-n-p bipolar transistor to achieve a complementary pair (centre), or an n-p-n transistor for a Darlington pair (right), as the output impedance may be considered to be approximately $1/g_m$. The effective output impedance is less than that of the f.e.t. alone.

Further reading

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 High voltages switched with a single transistor, 400 Ideas for Design, Vol. 2, 1971, Hayden.

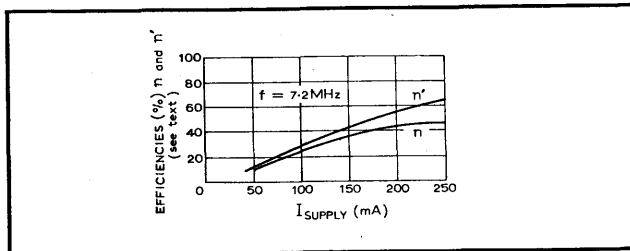


Class C power amplifier



Typical data

Supply: 12V

Tr₁, Tr₂: BFR41R₁: 100Ω; R₂: 50Ω
(carbon)C₁: 180pF; C₂: 360pFC₃: 47pF; C₄: 10nFC₅: 500pF; C₆: 190pFC₇: 805pF; L₁: 2.7μHL₂: 2.16μH; L₃: 2.38mHL₄: 230μH; L₅: 1.51μH

Circuit description

Many class C amplifiers find application in the v.h.f. and u.h.f. bands, special transistor fabrication techniques being used to optimize their performance. For correct design it is necessary to establish a suitable model for the transistor behaviour under class C conditions, some manufacturers providing the appropriate data. In general, this data is not available for class C designs operating at frequencies lower than about 10MHz, so that a successful circuit normally results from a breadboard version using variable capacitors. The circuit shown above was produced on this basis where C_2 , C_3 , C_5 , C_6 , C_7 were originally fixed capacitors 'padded' out with variables. Source and load resistance were 50Ω and the output power obtained at 7.2MHz was 1.41W with a drive signal producing 250mA supply current. Overall efficiency was only 47% (see graphs) but taking account of the d.c. drop (3.52V) across the r.f. choke L_4 efficiency rises to 66.5%. Hence L_4 should have low resistance, but its effect is less noticeable at lower currents. Transistors Tr₁ and Tr₂ were general-purpose transistors connected in parallel to reduce dissipation problems. The tuned networks in the input and output circuits should match the source to the transistors and the transistors to the load for maximum power transfer. Careful layout is essential and the circuit can easily oscillate as L_3 , L_4 and the collector-base capacitance of the transistors form the basic arrangement of a Hartley-type oscillator.

Component changes

The circuit can operate over a limited frequency range and a wide range of supply voltages and power levels provided the input and output networks are re-adjusted to cater for the changing values of transistor input and output resistance and capacitance.

Alternative general-purpose transistors can be used, such as BFY50.

Single transistor can be used when reduced power is acceptable.

Input transformer can be dispensed with if alternative input and output networks used (see over).

Circuit modifications

Correct design procedures for class C r.f. power amplifiers tend to be highly analytical due to the need to consider the correct choice of input and output coupling networks, their working Q-factors, degree of harmonic rejection, possible causes of spurious oscillation and the d.c. operating conditions. For a successful design the impedances at the transistor input and output terminals must be known under the desired operating conditions. Use of small-signal parameters leads to considerable errors in a class C design as the voltage and current swings are so large in such a power amplifier. When class C transistor data is available it is normally provided in the form of equivalent parallel input resistance and reactance and parallel output capacitance as a function of frequency and power output. The equivalent parallel output resistance is given approximately by $R = V_{cc}^2 / 2P_{out}$. Even with this data available a choice must be made from the large number of possible input and output coupling networks. Often a T-configuration is suitable for both networks as shown left. These networks complex-conjugate match the source to the transistor and the transistor to the loads. Both networks introduce losses due to component imperfections. Choice of the working Q-factors is a compromise between losses in the coupling networks, their selectivity and realizable component values. If the loaded-Q is high the capacitors will be small, the selectivity will be high but the losses will be large. A low working Q-factor implies the opposite. When the available data is correctly interpreted it will normally still be necessary to tune the amplifier for optimum performance, for example by adjustment of C_1 to C_4 . Complete design procedures are given in the first three references.

Further reading

Motorola, application note AN-282: Systemizing r.f. power amplifier design, 1967.

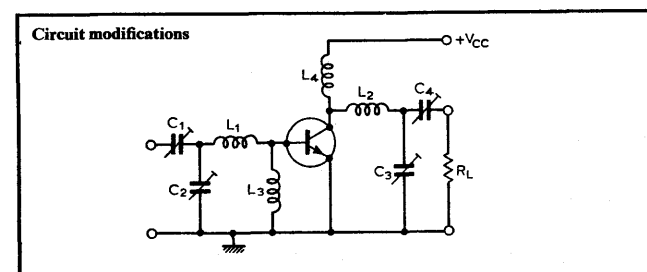
Hilbers, A. H., On the input and load impedance and gain of r.f. power transistors, *Electronics Applications*, vol. 27, 1967, pp.53-60.

Mulder, J., On the design of transistor r.f. power amplifiers, *Electronic Applications*, vol. 27, No. 4, pp. 1967, 155-171.

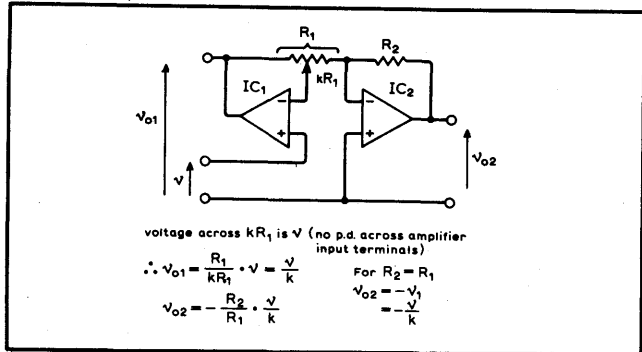
Markus, J. (ed.), 8-MHz, 3-W amplifier, in *Electronic Circuits Manual*, 1971, p.15.

Cross references

Circard series 7, card 1.

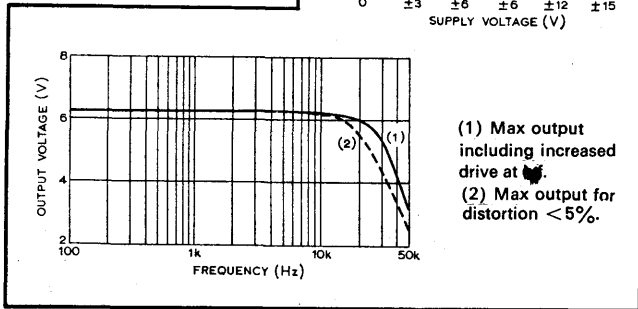
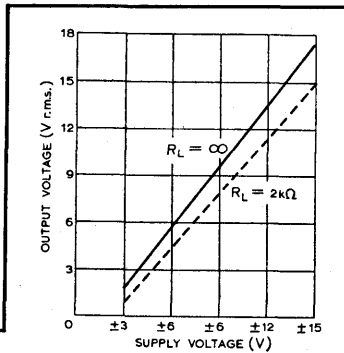


Bridge output amplifiers



Typical data

IC₁, IC₂: 741
 R₁: 10kΩ pot
 R₂: 10kΩ
 Supplies: ±15V
 R_L: 2kΩ
 Output voltage: 15V
 r.m.s. into 2kΩ (17.5V
 r.m.s. o/c) for $k = 0.1$ to
 1.0 at 1kHz.



Circuit description

Most power amplifiers have a single-ended output, delivering to the load a voltage whose peak-to-peak value is at most equal to the total supply voltage. If transformers/inductors are allowed such single-ended stages may produce peak-to-peak output voltage swings of up to double the supply voltage, but only if the transistor breakdown voltages are equally high. The economic and performance limitations imposed by transformers point to the need for an alternative output configuration for increased output voltage swing. If the load is taken between the outputs of two amplifiers delivering inverted outputs of equal magnitude, then the load voltage being the difference between the two has twice the magnitude of each separately. The method is illustrated using standard operational amplifiers, but is applicable to amplifiers at all power levels, where the constraint of a grounded load need not be met. This particular configuration offers the advantage that a single

potentiometer controls the gain of both channels. The exact balance is adjusted if required by setting $R_2 = R_1$. Equal magnitudes of output are ensured for this condition assuming ideal amplifiers because the two resistors carry equal current while their junction is a virtual earth point. A further advantage of this circuit is the high input impedance. As only one amplifier has a common-mode signal, the amplitude response differs somewhat, but the difference is only significant at those frequencies where the characteristic of each amplifier has departed significantly from the ideal. Slew-rate limiting, an output circuit phenomenon, determines the highest frequency at which large output voltages are obtainable with low distortion.

Component changes

- Replace amplifiers by any compensated type (307, etc.); alternatively use uncompensated types (748, 301, etc.) with appropriate compensation capacitor (reduced compensation possible with increased gain leading to higher slew rate).
- Resistor values non-critical but $R_1 = R_2$ gives push-pull output (circuit usable as phase-splitter for succeeding stages). Resistor R_2 may be made adjustable to take up tolerances if outputs are required to be given ratio, leaving tapping point on potentiometer to vary total gain. Typical values for R_1, R_2 ; 1kΩ to 250kΩ. Higher values lead to offset, drift and additional h.f. limitations; lower values absorb too much of the available output current.
- If unity gain is sufficient, IC₁ may be replaced by voltage follower, R_1 replaced by fixed resistor.

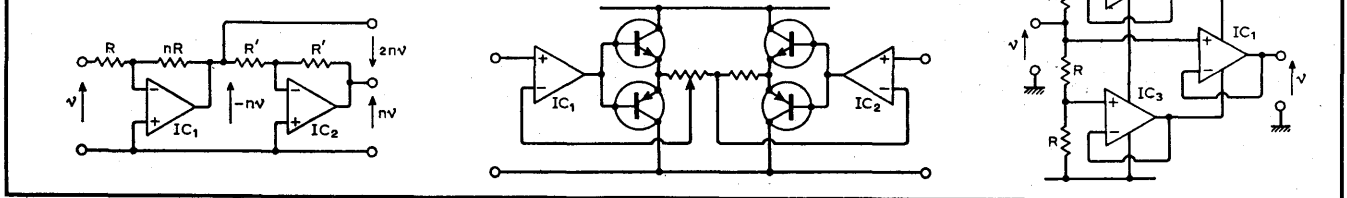
Circuit modifications

- Using two separate inverting amplifiers, with second set for a gain of -1, control over both outputs is obtained by varying the gain of the first. As both are used as virtual-earth stages feed-forward compensation may be used to obtain stable performance with considerable increase in slew-rate and cut-off frequency.
- Current capability of the output stages can be increased by any of the ways suggested on the cards describing class B/class A amplifiers. The simplest addition is a pair of complementary emitter-follower combinations. Output current capability may be increased by one or two orders of magnitude, but the output voltage swing is slightly reduced because of the base-emitter p.d. of the transistors. Crossover distortion may be minimized by the addition of diode/transistor biasing networks to the transistor base circuits. (middle)
- An alternative to the bridge circuit for increased voltage swing is the principle of supply bootstrapping of which this is one version. (right)

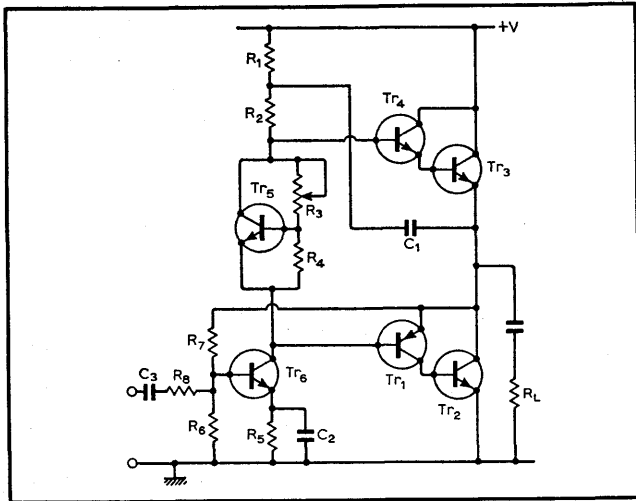
Further reading

Greiter, O., Transistor amplifier output stages, part 1, bridge circuits, *Wireless World*, vol. 69, 1963, pp.17-20.
 Del Corso, D. & Giordana, M., Simple circuit to double the output-voltage swing of an operational amplifier with increased slew rate, *Electronics Letters*, vol. 8, pp.151/2.

Circuit modifications



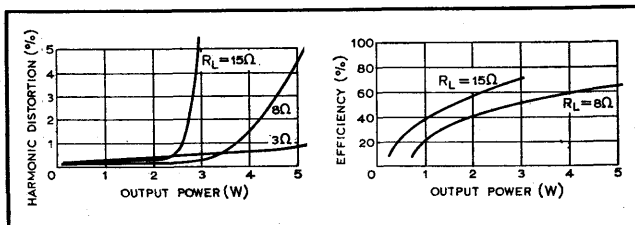
Class B quasi-complementary output



Typical performance

Supply: +20V
 Tr₁: BFR81; Tr₂, Tr₃: TIP3055
 Tr₄, Tr₅, Tr₆: BFR41
 R₁, R₂: 1.5kΩ; R₃: 1kΩ
 R₄: 470Ω; R₅: 330Ω
 R₆: 1.8kΩ; R₇: 8.2kΩ
 R₈: 1kΩ; R_L: 8Ω
 C₁: 100μF; C₂: 22μF;
 C₃: 10μF
 Main d.c. output: 10V
 Input signal: 2.6V pk-pk

Output signal: 6.7V pk-pk
 Output power: 5.4 watts
 Harmonic distortion: 5.8%
 Quiescent current: 0.41A
 Graphs of harmonic distortion and efficiency versus output power for loads of 15Ω and 8Ω shown below



Circuit description

This is a circuit of a class B push-pull amplifier in which transistors Tr₃ and Tr₄ complement the pair Tr₂ and Tr₁. To use n-p-n transistors in the output stage for economy, the configurations of the two sections are different, i.e. Tr₃ and Tr₄ are connected as a Darlington pair and Tr₁ and Tr₂ as a complementary pair. They receive essentially the same a.c. drive, but with the bases separated by Tr₅. Tr₃ and Tr₄ conduct

for positive-going output signals and Tr₆ supplies base current drive to Tr₁ and Tr₂ for negative-going output signals. Transistor Tr₅ is used in the so-called amplified diode configuration in which the potential difference between the bases of Tr₁ and Tr₄ is set as a multiple of the V_{be} of Tr₅ by the potential divider R₃, R₄, i.e. R₃ can be adjusted to give the desired quiescent current in transistors Tr₂ and Tr₃. A forward bias is available which may allow the transistors to conduct to a small extent, just sufficient to minimize the crossover distortion that can never be entirely absent. Transistor Tr₆ is an inverting amplifier with overall negative feedback through R₇, the values of R₆ and R₇ determining the d.c. output potential in conjunction with R₅. Because R₅ is decoupled, the a.c. properties of the arrangement are determined by the ratio of R₇ to the source resistance. Resistors R₁ and R₂ are centre-tapped and this point is taken to the output via C₁, which bootstraps R₂ so that the current through it remains constant throughout the cycle of output voltage swing.

Circuit modifications

- To avoid dangerous overcurrent in either of the output stage transistors, the current may be limited by adding series resistors R_e between the emitters and the output terminal (left).
- Middle circuit shows an alternative arrangement, adding transistors Tr₇ and Tr₈. These are normally non-conducting except under overload conditions, i.e. as the output current increases the voltage drop across R_{e1} or R_{e2} causes Tr₇ or Tr₈ to turn on and divert the base current available to Tr₄ or Tr₁, limiting the output current to V_{be}/R_e.
- Alternative configurations for the output stages are shown right (i) requires low and high power n-p-n and p-n-p transistors to make up the Darlington pairs, the minimum p.d. between input and output circuits being twice the V_{be} of a single transistor, (ii) uses complementary equivalent pairs with only one base-emitter path between input and output. Each pair comprises two inverting stages with 100% series-applied negative feedback giving unit gain.

Component changes

Adjustment of R₃ to avoid just visible crossover distortion gives a quiescent current of 7mA.

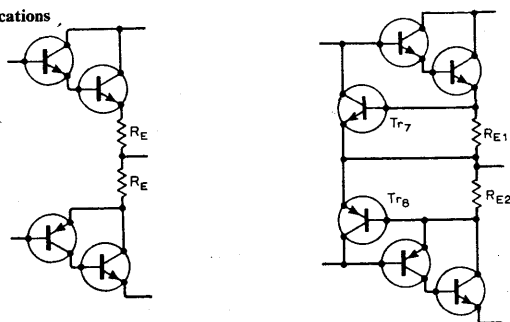
Further reading

- New uses for the LM100 regulator, National Semiconductor application note AN8-7.
- Grebene, B., Analog integrated circuit design, Van Nostrand 1972, pp.163-7.
- Amplifier efficiency (Letters), *Wireless World*, vol. 75, 1969, p.381.
- Hartz, R. S. & Kamp, F. S., Power output and dissipation in class B transistor amplifiers, RCA publication AN-3576. (Also in publication SSD-204A, p.594.)

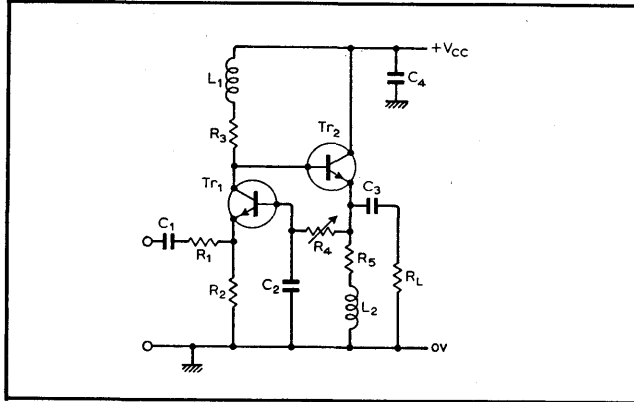
Cross references

Series 7, cards 1, 2 & 3.

Circuit modifications

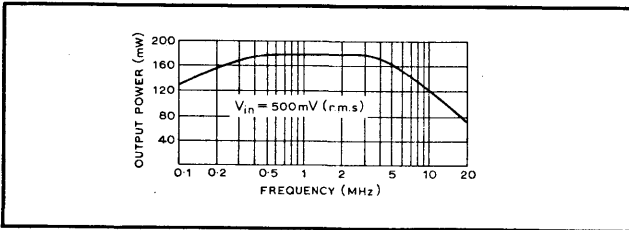


Broadband amplifier



Typical performance

Supply: +20V, 118mA
 Tr₁: BFR41; Tr₂: BFY50
 R₁: 33Ω; R₂: 150Ω
 R₃: 220Ω; R₄: 22kΩ
 R₅: 120Ω; R_L: 50Ω
 (carbon)
 L₁: 1.7μH; L₂: 220μH
 Power gain ≈ 14dB
 3dB bandwidth 64kHz to
 16MHz



Circuit description

In many applications the transfer of power to a load at maximum efficiency is not the primary consideration. Often, power gain is required for small input signals over a wide frequency range without introducing significant intermodulation and harmonic distortion. The common-base stage offers the best linearity of voltage gain against collector current, the latter changing in sympathy with this input signal. The emitter follower, while not providing voltage gain, gives a current gain of the same order as a common-emitter stage and is therefore very useful for transferring power to a load. To obtain this transfer with little distortion, it is necessary to operate the emitter follower at a relatively high quiescent current even for quite small input signals. The circuit uses a common-base stage feeding the load via an emitter follower. To maximise the gain-bandwidth product, Tr₁ and Tr₂ operate in regions where their current gain is much smaller than their normal current gains. The input resistance of the common base stage is inverse to its quiescent current, so that a high current allows the amplifier's input resistance to be matched to that of the source by a suitable choice of R₁. Resistor R₃ is determined from the required voltage gain (A_v) for equal source and load resistances $R_3 \approx A_v h_{te2} R_L / (A_v + h_{te2})$. Inductor L₁ is included to offset the capacitive loading due to Tr₂ and strays to maintain the gain at high frequencies. To deliver as much output current to R_L as possible at high frequencies choke L₂ is included in series with R₅.

Component changes

With V_{cc} (min) = +5V, V_{in} (max) ≈ 140mV r.m.s., supply current is 30mA, and P_{out} ≈ 11mW.

Tr₁ and Tr₂ can both be BFR41.

Tr₁ can carry a much smaller quiescent current, using for example an ME4103, with increased values of R₂, R₃ and R₄. R₁ can be increased or decreased to allow matching to source resistances greater or less than 50Ω respectively.

Circuit modifications

If the input signals are very small, output powers of around half a watt can still be obtained over a wide bandwidth by cascading a pair of amplifiers of the type described. When the gain-bandwidth product of the amplifier is not the most critical requirement and a higher efficiency is needed, the quiescent current in Tr₁ may be drastically reduced. Resistors R₂ and R₃ would then need to be increased, with a corresponding increase in R₄, if this is to be the means of controlling the quiescent operating conditions. The lower Tr₁ current may be chosen to make the natural input resistance of the stage, in the absence of R₁, the value required to match the source.

● Input resistance may be defined using shunt-applied feedback, as shown left, where the emitter of Tr₁ is d.c. or a.c. grounded, the feedback is not decoupled and the voltage gain is determined by the ratio R_A/R_B. The input resistance is largely that of R_A except at high frequencies where the feedback falls and the impedance at Tr₁ base must be considered.

● Inclusion of R₆, as shown right, may be applied to both the previous circuits to allow an output to be taken from the collector of Tr₂. To maximize the signal swing in the collector circuit of Tr₂ the bias network must be readjusted to leave a small voltage at Tr₂ emitter, say by reducing R₄ and R₂ in the original circuit. The output resistance is approximately R₆; this stage is therefore convenient for feeding directly into any other low impedance stage, such as that left, with R_A removed. This mismatch can often be of advantage in extending the bandwidth of the amplifier.

Further reading

Hirst, R., Wideband linear amplifier, *Wireless World*, vol. 75, 1969, pp.168-70.

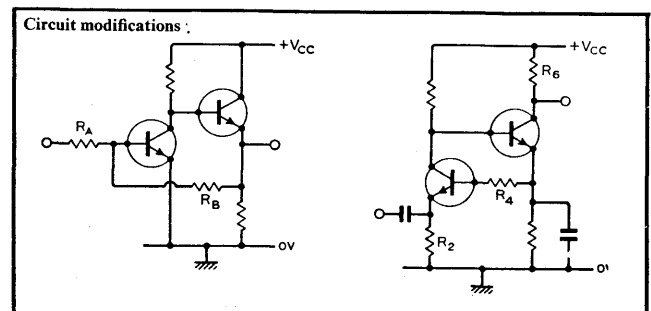
Meindl, J. D. & Hudson, P. H., Low-power linear circuits, *IEEE Journal of Solid-State Circuits*, vol. 1, 1966, pp.100-11.

Lo, A. W. (and others), Transistor for Electronics, Chapter 9, Prentice-Hall, 1955.

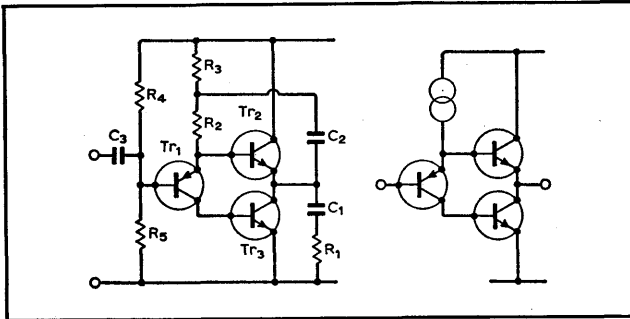
Griffiths, H. N., Simple wideband amplifier, *Wireless World*, vol. 75, 1969, p.478.

Cross references

Series 7, cards 1, 4, 5 & 10.



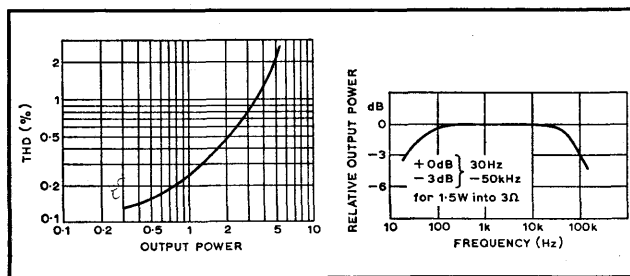
Class A op-amp power booster



Typical performance

Tr₁: BFR81
 Tr₂, Tr₃: TIP3055
 R₁: 3Ω; R₂: 250Ω pot.
 R₃: 220Ω; R₄, R₅: 10kΩ
 C₁: 2,000μF; C₂: 470μF
 C₃: 10μF
 Supply voltage: 12V
 Quiescent current:
 1.25A (set by R₁)

Output power for 1%
 t.h.d. into 3Ω load: 4.2W
 (supply current falls to
 1.05A at full output).
 Output voltage swing to
 within about 0.7V of
 supply lines for 3Ω load
 and about 0.15V for 15Ω
 load.



Circuit description

Available operational amplifiers have limited output currents, but may have a voltage swing approaching supply values. The circuit shown is class A buffer amplifier of unity voltage gain which may be added to such amplifiers to increase their output current to 1A or more. In addition the circuit is a very simple version of the voltage follower, having a low d.c. offset between input and output, a voltage gain very close to unity and a high input impedance. With the bootstrap technique applied the amplifier is capable of driving low load resistances to within <1V of each supply line.

If a constant current flows in R₂, then as the base potential of Tr₁ increases the emitter current of Tr₁ decreases and with it the collector current.

This fall is fed to the base of Tr₃ causing it to conduct less,

while the fall in emitter current releases more of the constant current in R₂ to flow in the base of Tr₃. Provided the current gain of Tr₁ is reasonably high, the magnitudes of the base current charges in Tr₂, Tr₃ are equal but the signs are opposite. This represents an approach to ideal current phase-splitting. The constant current in R₂ is provided by the bootstrap capacitor C₂, such that any change in the potential at the base of Tr₂ is coupled via the follower action to the positive end of R₂, i.e. with no resulting change of p.d. across R₂ in the ideal case. Resistors R₂ or R₃ require to be variable to set the output current and stability of that current then depends on h_{FE} variation in Tr₂, Tr₃. The base-emitter p.d.s of Tr₁, Tr₂ substantially cancel, as they can readily be chosen for junction area ratios matching the quiescent current ratios. As a class A amplifier, maximum theoretical efficiency is 50%. At full output the load power may approach 40% of supply power in practice, but the quiescent power is somewhat higher than the supply power at full load.

Circuit modifications

- The good d.c. offset characteristics allow the amplifier to be used as a voltage follower with d.c. coupling to the load. Bootstrapping should be retained unless the amplitude response is required to extend to d.c., as it swings the junction of R₂, R₃ above the supply on positive signal swings. Hence it can drive Tr₂ base far enough positive to saturate Tr₂ hard making maximum use of available supply voltage. If the load is to be a.c. coupled but may carry a small quiescent current, the load resistance R₁ may replace R₃. (left)

- Any other constant-current circuit may replace the bootstrap arrangement, e.g. a f.e.t. either with gate strapped to source as shown or with a resistor in the source lead to define some lower value of current. (middle)

- Although the distortion of the buffer stage above is low, the addition of a high voltage gain amplifier such as an op-amp can increase the voltage gain to (R_B/R_A) + 1 while providing sufficient overall feedback to make distortion very low. The wide bandwidth of the buffer stage together with its unity gain minimizes the risk of instability at high frequencies. Should this be troublesome an op-amp with external compensation may be used with increased compensation capacitor. (right)

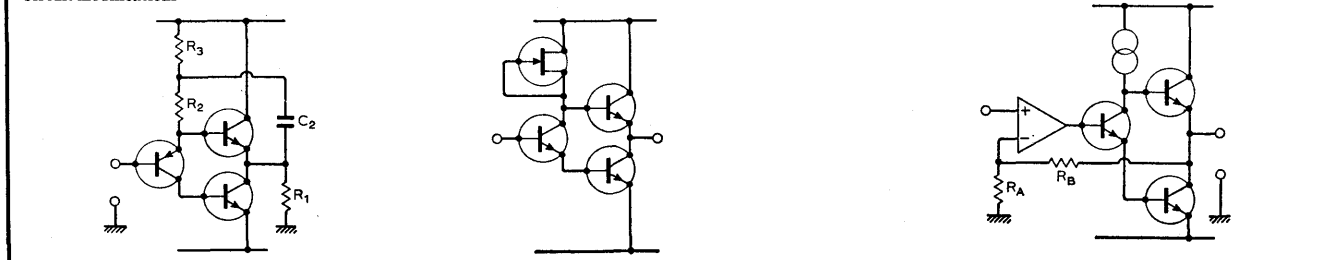
Further reading

Belcher, D. K., Inexpensive circuit boosts op-amp output current, 400 Ideas for Design, vol. 2, Hayden, pp.1-2.
 Bloodworth, G. G., D.C. amplifier with unity voltage gain, *Electronic Engineering*, 1965, pp.112-4.
 Electronic Circuit Design Handbook, Current boosters for i.c. op-amps, Tab, 1971, p.161.

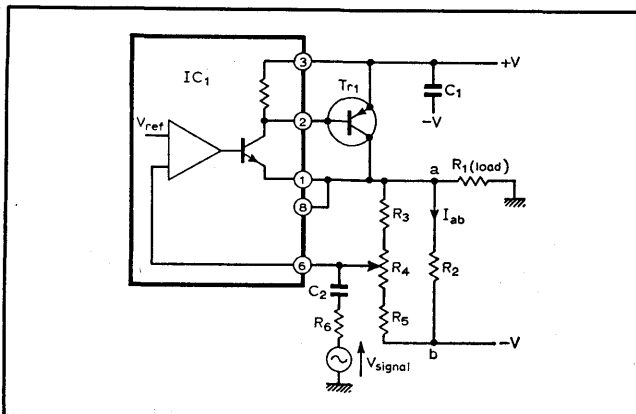
Cross references

Series 7, cards 2, 4 & 8.

Circuit modifications

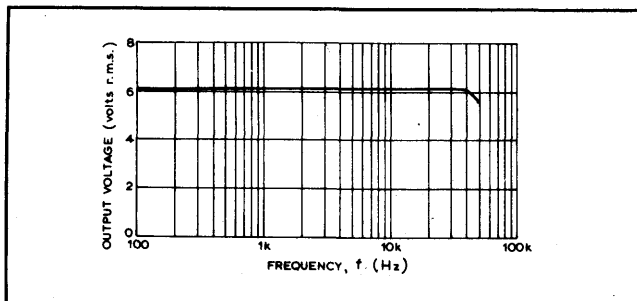


D.C. power amplifier



Typical performance
 IC₁: LM305 or LM100
 Tr₁: MJE271
 Supplies: ±15V
 R₁, R₂: 150Ω; R₃: 15kΩ
 R₄: 1kΩ; R₅: 1.8kΩ
 R₆: 10kΩ
 C₁: 1μF (tantalum);
 C₂: 10μF

Input signal: 2.8V r.m.s.
 at 100Hz
 Maximum output voltage before symmetrical clipping 6.2V r.m.s.
 Output power: 250mW
 Harmonic distortion:
 0.35% at 1kHz, and
 0.32% at 20kHz.



Circuit description

This circuit uses a voltage regulator i.c. package to supply an output stage Tr₁ where the amplifier is to be used primarily with a unipolar signal, though it can also be interpreted as a class A output stage which can be a.c. coupled to a load. The i.c. regulator contains its own reference voltage and a separate feedback point (terminal 6) which allows the potential at the collector of Tr₁ to be set to some stable value which is a multiple of the internal reference voltage, that multiple being set by R₃, R₄ and R₅ the quiescent current in Tr₁ is then set by the bias resistor R₂ in conjunction with this predetermined voltage.

An a.c. signal can be superimposed at pin 6 via C₂ and R₆ the circuit then behaving as a see-saw amplifier, as the reference voltage leaves the feedback terminal 6 as an a.c. virtual earth. For d.c. purposes, the circuit may be treated as series applied feedback. The peak current in the load is limited to a fraction of the quiescent current for negative excursions; as the voltage goes negative the p.d. across R₂ falls and with it the current through R₂. The current in R₁ for this voltage excursion can never exceed R₂ even when the transistor current falls to zero in the positive direction; however, much greater currents can be provided through Tr₁. The amplifier is thus an inefficient class A amplifier whose effectiveness can be improved by replacing R₂ by a constant-current stage which can sustain a given peak current in R₁ almost equal to the quiescent value, even for large voltage excursions in the negative direction. Capacitor C₁ is used to suppress h.f. oscillation and a low inductance type must be used.

Onset of slew-rate limitation occurs at 70kHz for an output signal level of 16V pk-pk when the signal level is reduced to 3 to 5V pk-pk by reducing the input signal. Voltage gain is flat up to 100kHz, with 3dB fall-off occurring about 250kHz.

Circuit modifications

- Resistor R₂ is replaced by the Baxandall constant-current circuit shown left Tr₁: BFR81, Tr₂: TIP3055, R₇: 18Ω, R₈: 3.9kΩ. This permits a much greater input signal level before peak clipping occurs. Resistor R₇ is chosen for approximately a 100mA constant quiescent current in the path a-b (about 1.8V is available at terminal 6). If the transistor Tr₁ output current is 200mA pk-pk, then the output current swing in load R₁ is twice that for the case when R₂ is 150Ω with the same quiescent current. A comparison of instantaneous currents for the two possible circuits between a and b is tabled below.

- The regulator may be replaced by the operational amplifier emitter-follower circuit, shown right. To maintain the d.c. stability of the output, the non-inverting terminal must be connected to a suitable stable reference voltage. If the d.c. power supply is stabilized, then this may be a tapping on a potential divider connected across the supply. For minimum drift, the effective resistance seen at both input terminals of the op-amp should be comparable.

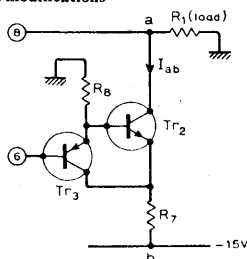
Further reading

New uses for the LM100 regulator, National Semiconductor application note AN-8, 1968.
 Amplifier efficiency (Letter), *Wireless World*, vol. 75, 1969, p.535.

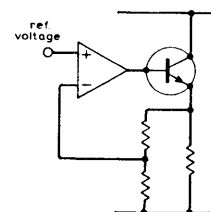
Cross references

Series 3, card 8.
 Series 7, card 12.

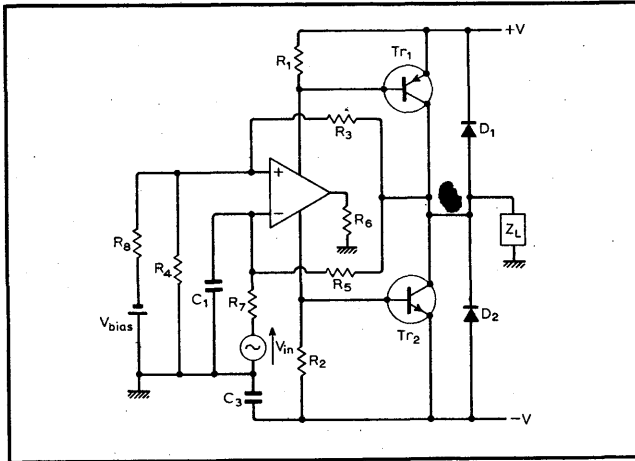
Circuit modifications



current through Tr ₁ (mA)	I _{ab} (mA)	load current (mA)	circuit element in a-b
100	100	0	R ₂
200	150	+50	
0	50	-50	
100	100	0	constant current source
200	100	+100	
0	100	-100	



Class D switching amplifier



Typical performance

A₁: 301
 Tr₁: 301
 Tr₂: BFR41
 D₁, D₂: SD2
 Supply: ±10V (4 to 15V)
 R₁, R₂: 180Ω, R₃: 10kΩ
 R₄: 100Ω, R₅: 4.7kΩ
 R₆: 470Ω, R₇, R₈: 1kΩ
 C₁, C₂, C₃: 100nF
 Z_L: 1mH (*r* = 0.9Ω)
 +15Ω

V_{bias}: -640mV to set mean load voltage to zero with V_{in} = 0.

Switching frequency: 27.8kHz, max 40 kHz.
 With V_{in} = 0, supply current is ±20mA; with V_{in} = 3.4V pk-pk 100Hz; current is ≈ 130mA; power in 15-Ω load ≈ 1.66W; residual "carrier" ≈ 300mV across 15Ω; overall efficiency 64%; output stage efficiency ≈ 76%; 3-dB bandwidth ≈ 600Hz. With rectangular input at 100Hz, output rise and fall times ≈ 600μs.

Circuit description

Basically, the circuit is an astable oscillator, generating a squarewave that is used to drive a complementary pair of output transistors into conduction on alternate half-cycles of the squarewave. The output transistors thus switch the voltage to the load at a frequency that is much higher than that of the signals to be amplified. The squarewave generator is designed around the operational amplifier A₁ which uses positive feedback via R₃ and R₄. The periodic time of the squarewave fed to R₆ depends on the time constant R₅C₁ if R₇ is much greater than R₅. To obtain a realistic switching frequency with reasonable components and also to obviate the need for large input signals a compromise must be made in the value of R₇. Current in R₆ flows alternately in R₁ and R₂ producing p.d.s across these resistors that are sufficient to switch on Tr₁ and Tr₂ respectively. The signal applied to R₇ causes the mark-to-space ratio of the output waveform from the astable to vary in sympathy with the instantaneous value of V_{in}, so that the mean value of the voltage applied to the load also varies

directly with the input signal. If the load impedance has an external filter, or is by its nature self-filtering such as with a motor, then the power drawn from the amplifier at the switching frequency is low and the useful signal power in the load will be high.

If the mark-to-space ratio of the squarewave generated by the astable is not unity with V_{in} = 0, it can be made so by a suitable choice of the bias supply and R₈. Diodes D₁ and D₂ protect Tr₁ and Tr₂ against breakdown when the load impedance is highly inductive.

Circuit modifications

- The bias source to set the mark-to-space ratio of the squarewave to zero can be obtained by a potentiometer connected between ground and the appropriate supply line.
- While an inductor is normally used in series with a resistive load to filter out the h.f. squarewave, any suitable low-pass filter can in principle be connected between the junction of Tr₁ and Tr₂ collectors and R₁. Another possible method is to connect a capacitor in parallel with the inductive smoothing choke so that it is resonant at the switching frequency. For example, with a choke of 1mH and *r* = 1Ω, a parallel capacitance of 16nF would be resonant at switching frequency of 40kHz. At signal frequencies less than about 500Hz, the impedance of this tuned network is inductive, having a maximum impedance of about 3Ω.
- The complementary pair of transistors forming the output stage can be replaced by a bridge-type network as shown left. The four transistors are fed with complementary pulse-width-modulated squarewaves which cause the transistors to be switched on and off in pairs. With Tr₁ and Tr₄ on current flows in the load in one direction and is reversed when Tr₂ and Tr₃ are switched on.
- Another practical form of bridge output stage is shown right using a pair of voltage comparators to generate the complementary pulse-width-modulated switching waveforms. The bridge of power transistors is connected across a single-ended supply. Component details are given in the first reference.

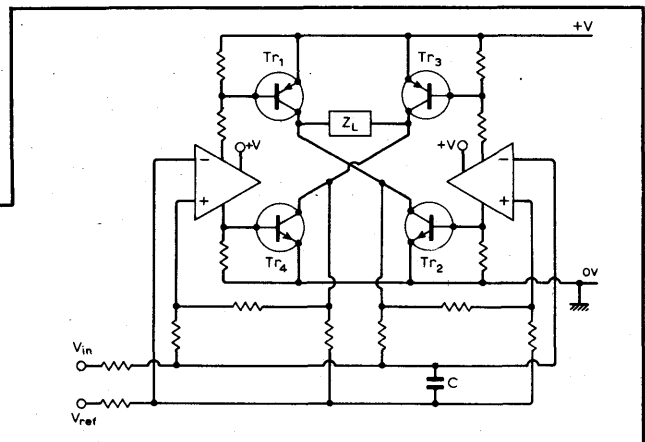
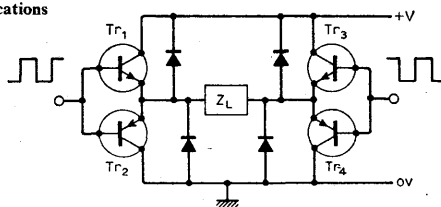
Further reading

National Semiconductor, data sheet and application notes on the LM311 voltage comparator, 1970.
 Camenzind, H. R., Modulated pulse audio and servo power amplifiers, International Solid-State Circuits Conference, University of Pennsylvania, Philadelphia, 1966, pp.90/1.
 Meidl, J. D., Micropower circuits, Wiley, 1969, pp.61, 64 & 65.

Cross references

Series 7, cards 1 & 2. Series 2, card 4. Series 3, card 1. Series 4, card 8.

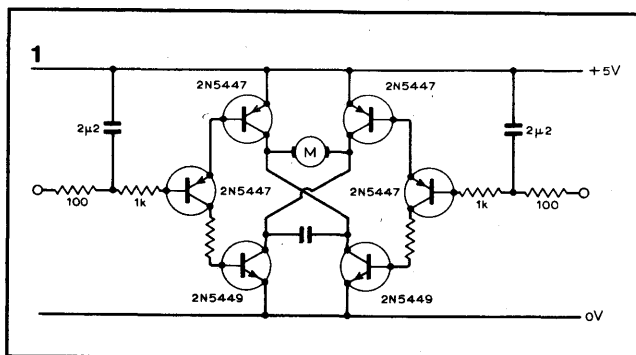
Circuit modifications



Power amplifiers

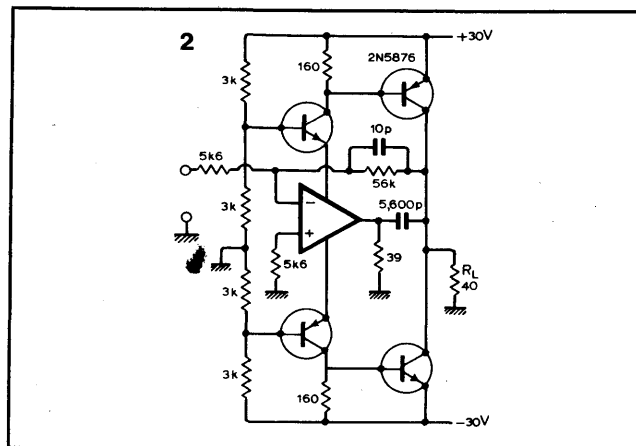
1. A bridge output configuration is very convenient for driving small motors in servo systems. It avoids the need for a centre tapped supply and with it the possibility of unbalance in the supplies which can cause unequal response in the two directions. The inputs can be driven from t.t.l. open-collector gates, though decoupling between the motor and logic-circuit supplies is desirable. The servo system is of the pulse proportional type with the transistors in the switched mode. The dissipation in each transistor is low and small plastic encapsulated types are adequate. For higher motor supply voltages, special open-collector devices are available with voltage ratings to 15V (e.g. SN7401A).

Bessant, M. F. Multi-channel proportional remote control, *Wireless World*, vol. 79, 1973, pp.479-82.



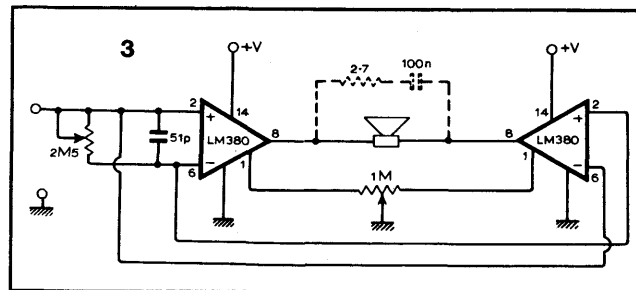
2. The use of the supply leads to an operational amplifier as signal output connections allows anti-phase current drives to a complementary pair of power transistors. By adding cascode-connected transistors the voltage limits of the op-amp can be adhered to while allowing any desired supply voltage. The cascode transistors experience the difference between supply and op-amp voltages while the output stages in this circuit have to cope with the full 60V. As shown, the peak power in the load can be as high as 22W and the output can swing to within 0.5V of either supply line. Amplitude response is flat to 30kHz.

Garza, P. P. Getting power and gain out of the 741-type op-amp. *Electronics*, vol. 46, 1973, p.99, (Feb. 1)



3. Recent monolithic audio power amplifiers have more and more of the required bias circuitry incorporated within the i.c. In some cases the only external components are a gain control and one or two components to ensure stability with particular loads (low load resistances in this circuit). In addition supply decoupling (with 0.1µF disc ceramic capacitor close to i.c.) may be needed if the supply is more than 2 to 3in from amplifier. The LM380 input is d.c. coupled, has internal feedback to set the gain and requires only a single polarity supply with the signal ground-referred. The bridge condition as drawn permits the output power to be doubled (limited by thermal ratings) and needs no coupling capacitor to load provided 1MΩ offset control sets outputs to same d.c. level.

Byerly, J. E. and Kooi, M. V. LM380 power audio amplifier, National Semiconductor application note AN-69, 1972, p.6.



Wireless World Circard

Series 8: Astable circuits

Almost every possible kind of astable circuit is covered in series 8, probably as a result of the interesting unified approach discussed in the article. The oscillator is depicted as a bridge network (Fig. 5) with one bridge arm providing positive feedback, the other negative feedback, the amplifier input differentially connected between the two centre points of the arms, and the output feeding the bridge to sustain oscillation. This approach allows one to see how the six configurations of Figs. 6 and 7 are really variants of the basic bridge circuit. Practical circuits based on Figs. 6 and 8 are given on card 5.

Five cards show other kinds of astable circuit constructed from logic elements, one being a hybrid in that discrete components are used to provide a constant charging current to cross-coupled monostables (card 11). A single-capacitor astable using t.t.l. NOR or NAND gates as inverters is useful up to frequencies of a few megahertz (card 7). Two other single-capacitor circuits, the c.m.o.s. inverter circuits of card 1 and the current-switching emitter-coupled circuit of card 9, both allow voltage control of frequency.

Whilst some logic astable circuits have poor frequency-supply voltage stability, most discrete-component circuits have a stability of the order of 1% per volt, and three have stabilities of less (0.46% card 10, 0.3% card 3, 0.2% card 5).

- Complementary m.o.s. astable circuit 1
- RTL astable circuit 2
- Complementary astable circuit 3
- TTL Schmitt astable circuit 4
- Operational amplifier astable circuit 5
- Astable blocking oscillator 6
- TTL dual inverter astable circuit 7
- Coupled logic gates astable circuit 8
- Emitter-coupled astable circuit 9
- Discrete-component Schmitt astable 10
- Dual-monostable astable circuit 11
- Astable circuit with f.e.t. 12

Astable circuits

Many circuits generating periodic rectangular waveforms depend on changing the charge on a capacitor. This rate of change of charge is frequently determined by an RC circuit, which inherently produces exponential waveforms when connected to a voltage source (Fig. 1). Such waveforms may be used to control the switching on and/or off instants of an active device.

An astable multivibrator using discrete

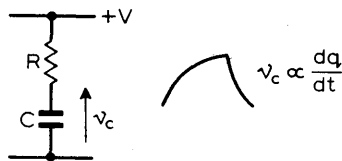


Fig. 1. Series RC produces an exponential waveform across the capacitor.

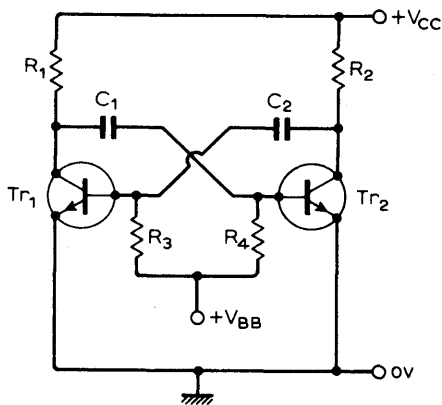


Fig. 2. Basic astable multivibrator. Variation of V_{BB} changes the period.

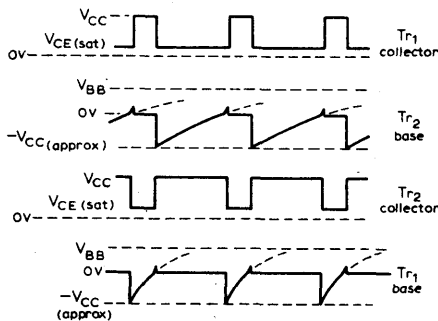


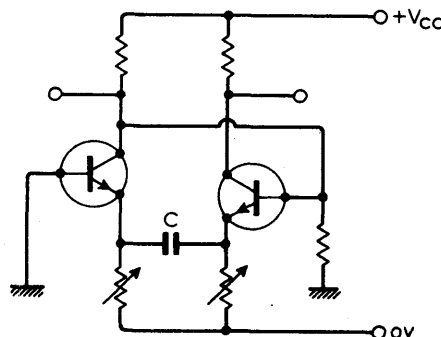
Fig. 3. Waveforms encountered in the circuit of Fig. 2, where C_1R_4 is longer than C_2R_3 .

components is a commonly occurring example of a circuit employing this principle (Fig. 2). This uses two transistors, cross-coupled via capacitors C_1 and C_2 . Normally this circuit has no stable states (they may be called quasi-stable states), but switches alternately from the one state of Tr_1 saturated, Tr_2 off, to the state of Tr_2 saturated, Tr_1 off. For a given value of V_{CC} , the rate of switching depends on the time constant C_1R_4 , C_2R_3 , and on the potential to which the base resistors R_3 and R_4 are returned. The more conventional configuration will return R_3 and R_4 to the $+V_{CC}$ rail to give a period independent of rail voltage.

Fig. 3 shows that the output rectangular waves available at the collectors are in anti-phase, and their mark-to-space ratio may be varied by adjusting the C_1R_4 and C_2R_3 time constants. Some astables are attractive due to the small number of components required, but they might be considered to have certain disadvantages. The use of a single CR network gives an output waveform having a non-unity mark-to-space ratio which may be difficult to closely control. Also a second anti-phase output is not available. A circuit using a single capacitor that overcomes these objections to an extent is the emitter-coupled astable of Fig. 4. In this arrangement, two emitter resistors are used to allow independent adjustment of the mark and space times of the two output waveforms.

Another group of astable circuits apparently different from each other fit into the form of the general bridge network shown in Fig. 5. The amplifier block is provided with two external

Fig. 4. A single-capacitor astable multivibrator providing mark-to-space ratio adjustment and antiphase outputs. Circuits using one capacitor are normally less flexible.



networks. One network, providing negative feedback, ensures that the d.c. conditions in the amplifier are such that it sits near the middle of its operating region where the gain is high. The other network, which provides positive feedback, forces the amplifier to switch between two distinct states. This amplifier block may comprise either a single differential-input amplifier or two single-ended types. In

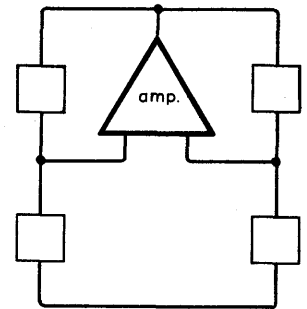


Fig. 5. Basic diagram of the group of oscillators using both positive and negative feedback. The circuit is seen to be a bridge network with a sustaining amplifier.

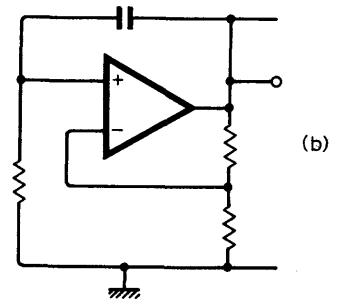
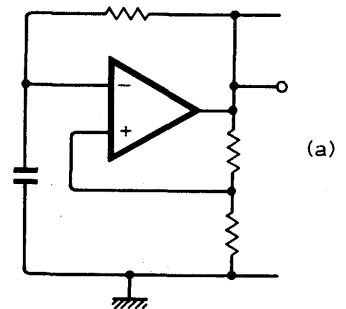


Fig. 6. Two possible combinations of feedback types using a single differential amplifier.

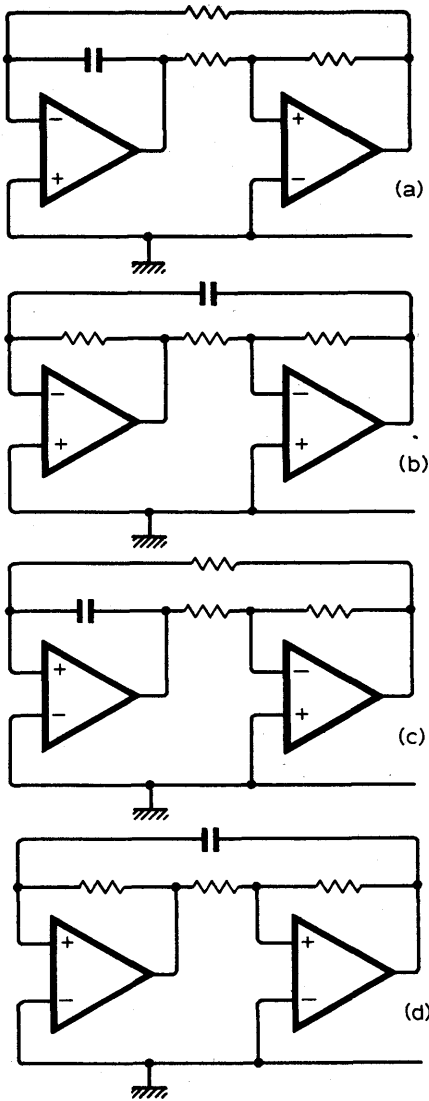


Fig. 7. Four combinations of RC feedback and amplifiers used in the single-ended mode.

conjunction with the different combinations of feedback components, this leads to the apparently different configurations shown in Figs. 6 and 7. These circuits have been redrawn in Figs. 8 and 9 respectively to show their relationship to the bridge arrangement. Circuit 7(d) does not operate as an astable because it will permanently latch into one state, since positive feedback is applied across each single-ended amplifier.

The output waveform from an astable multivibrator is not necessarily rectangular. A self-retriggering action can be obtained by using an electronic switch to discharge a capacitor when its potential exceeds a preset d.c. bias which holds the switch open.

To provide circuits compatible with logic levels, many astables operate in their saturated mode to give well-defined voltage limits. This is achieved at the expense of the switching rate, but may be avoided by the use of current mode switching techniques.

Other applications demand long periods with accurately defined transition times, and this requires high stability passive components. Reasonably small value but stable capacitors can be used in conjunction with field-effect transistors to provide the long time constants necessary. It should be remembered that if an astable multivibrator has good frequency stability, it may prove difficult to synchronize it from an external source.

The reader will have noted that inductive timing elements are conspicuous by their absence. The reason behind this is that, in comparison with capacitors, inductors tend to be more costly and physically larger. One type of inductively coupled circuit worthy of mention is the astable blocking oscillator, which is capable of producing output pulses having a very small mark-to-space ratio.

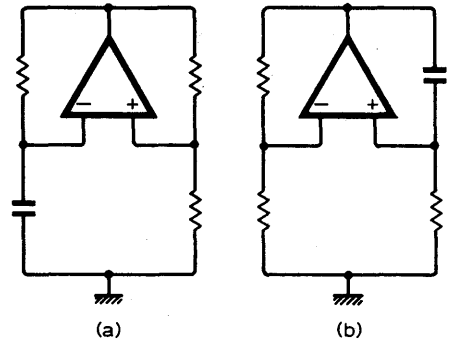


Fig. 8. Circuits of Fig. 6 redrawn to show that they conform to the general bridge circuit of Fig. 5.

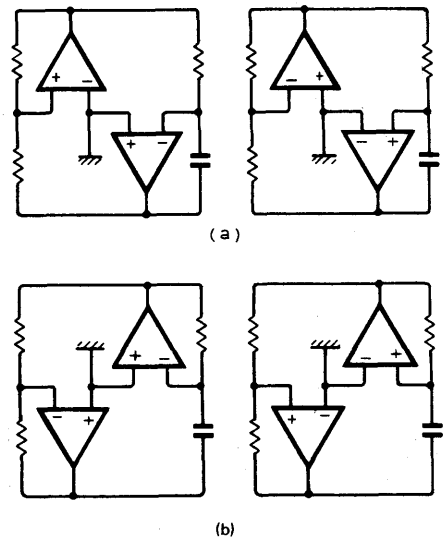
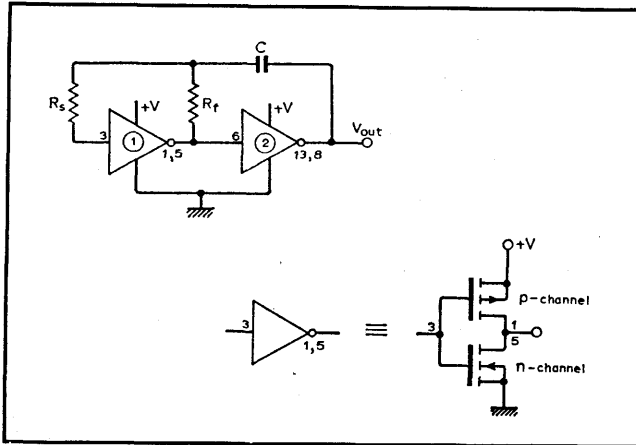


Fig. 9. The four twin-amplifier circuits are also examples of the general bridge circuit.

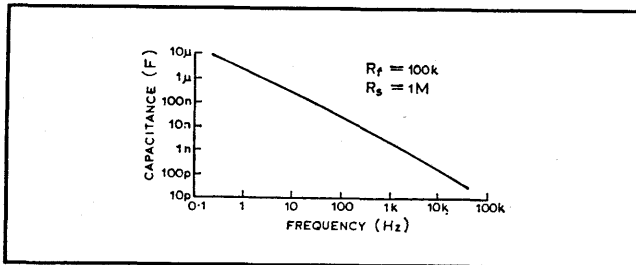
Complementary m.o.s. astable circuit



Typical performance

IC: CD 4007AE
 Supply: +10V
 R_f : 100k Ω ; R_s : 1M Ω
 C : 10nF; f : 424Hz
 Load resistance: ∞
 Supply current: 280 μ A

Square wave available at V_{out}
 Output excursion: 0.03 to 9.9V
 Mark-to-space ratio: 0.93
 Rise time: 200ns



Circuit description

This integrated circuit package comprises three n-channel and three p-channel enhancement-type m.o.s. transistors which may be arranged to form three separate inverters. The above circuit uses two inverters, the first inverter being biased to its amplifying region by resistor R_s , and in this region the loop gain is sufficient to initiate multivibrator action. When the output of inverter 2 goes high, the input is low and the input of inverter 1 is high. As the capacitor charges up via resistor R_f , the voltage across R_f , and hence the voltage applied to the gate of the first inverter, falls. When this voltage at the junction of C and R_f passes through the threshold value of the first inverter, its output becomes high, switching the output of inverter 2 to a low state. Capacitor C will now charge in the opposite direction via resistor R_f and when the voltage at the junction of C , R_f and R_s rises towards and crosses over the threshold level, the output of inverter 1 again

goes low, the output of inverter 2 is switched to the high state and the cycle repeats.

The waveform achieved is fairly symmetrical because the threshold point is close to half the supply voltage value. However, this means that the mark-to-space ratio is not unity, but this may be arranged by circuit modification. Resistor R_s also improves the frequency stability of the circuit with respect to supply voltage changes, and should be at least twice the value of R_f .

Component changes

- With supply of +10V, R_f of 100k Ω , and C of 2.2nF, mark-to-space ratio varies from 0.76 to 0.92:1 for R_s from 0 to 1M Ω .
- Components as listed in typical performance data but with finite load resistance R_L . Output pulse level falls, typically by 10% when $R_L = 2.2k\Omega$.
- Minimum value R_f for acceptable waveform: 6.8k Ω . Waveform improved by using third inverter as buffer. With R_s of zero, R_f : 6.8k Ω , C : 39pF, f is 610kHz (supply 10V). With R_s of zero, R_f : 10k Ω , C : 10pF, f is 650kHz (supply 10V). If supply is increased to 15V, f is 900kHz.

Circuit modifications

- Output waveform duty cycle may be controlled by replacing R_f with the arrangement shown left. The adjustment of this diode shunt causes the frequency of the circuit to vary, and another variable resistance can be added to compensate the change. If a 50% duty cycle is not obtained, reverse the diode D_1 .
- A voltage-controlled oscillator is obtained when R_f is replaced by the arrangement shown centre. With V_G in the range 0 to +10V using an n-channel device, frequency is variable from approximately 20 to 30kHz for a supply of +10V and R_f : 10k Ω , R_s : 100k Ω and C : 2.2nF.
- A simple way of synchronizing the circuit with an external source is shown right. Components R_x , C_x change the natural multivibrator frequency. With components shown, free-run frequency is 2220Hz increasing to 3985Hz when synchronizing components are connected but with zero source signal. Locking frequency range approximately 22/1 but can depend on level of synchronizing pulse. Suitable pulse level 0.5 to 1.0V.

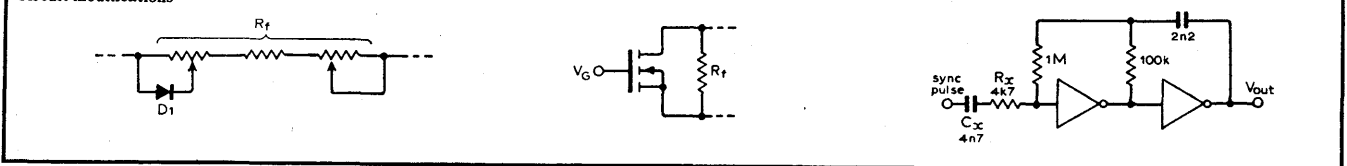
Further reading

RCA COS/MOS Digital Integrated Circuits, SSD-203A, 1973, pp. 353-9.
 Low-speed astable uses c.m.o.s., *Electronic Components*, 6 April, 1973, p. 294.
 Clock oscillator for telemetry systems uses c.m.o.s. chip to minimize power drain, *Electronic Design*, vol. 20, 1972, p. 84.

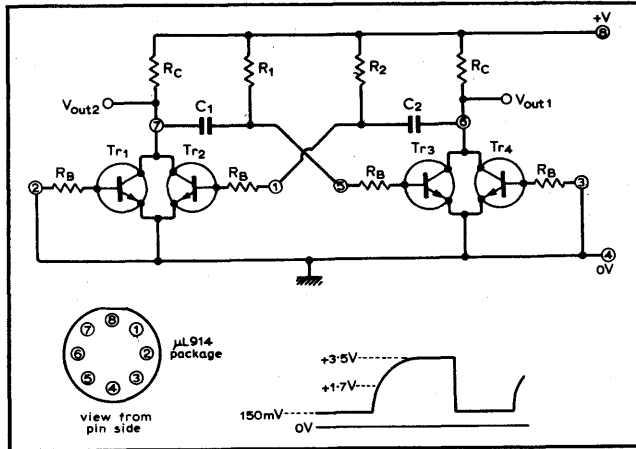
Cross references

Series 8, card 3.
 Series 3, card 11

Circuit modifications



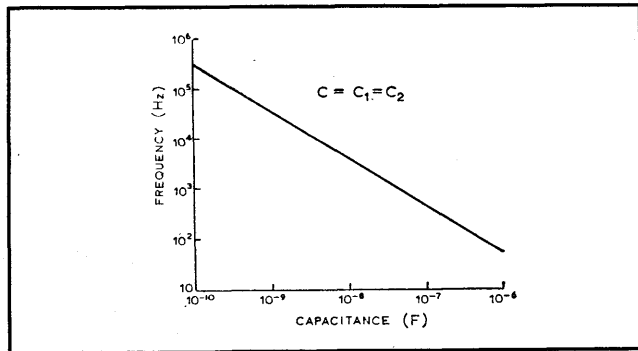
R.t.l. astable circuit



Typical performance

μL914 package contains
 —four 2N708-type
 — R_B of 450Ω; R_C of 640Ω

External components:
 R_1, R_2 : 10kΩ ± 5%
 C_1, C_2 : 100nF ± 10%
 Supply: 3.6V, 6.5mA
 P.r.f. 699Hz (see graph)
 Mark to space ratio: 1.06
 V_{out1} waveform as shown



Circuit description

The μL914 contains two identical resistor-transistor logic (r.t.l.) gates. In the above arrangement one input to each gate is not used, pins 2 and 3 being grounded to effectively remove Tr_1 and Tr_4 from the circuit. Transistors Tr_2 and Tr_3 are interconnected to form a cross-coupled astable which may be considered to be a two-stage amplifier with its output fed back to its input and having very high loop gain. The circuit is inherently self-starting; any dissimilarity however small between the two halves of the circuit causes one transistor to be off and the other saturated.

Consider Tr_2 on and Tr_3 off. In this state the circuit levels are: Tr_2 collector: $V_{CE(sat)}$, Tr_2 base: $V_{BE(on)}$, Tr_3 collector: +V and Tr_3 base: approx. -V due to the negative-going transition at Tr_2 collector. When switched from off to on the charge on C_1 cannot change instantaneously. C_1 's charge will then change with a time constant $C_1 R_1$, as its right-hand plate

attempts to change to +V from -V. However, when this potential slightly exceeds 0V, Tr_3 's base-emitter junction becomes forward-biased and it rapidly turns on, its collector voltage falling to $V_{CE(sat)}$. The negative step passes to Tr_2 base through C_2 switching Tr_2 off. The circuit is now in its other quasi-stable state. This action repeats continuously, producing antiphase square waves at Tr_2 and Tr_3 collectors. The off-times of Tr_3 and Tr_2 are given by $t_1 = 0.6931 C_1 R_1$ and $t_2 = 0.6931 C_2 R_2$ sec. The p.r.f. of the square waves is thus: $f = 1/T$, where $T = t_1 + t_2$. The mark-to-space ratio is adjustable by altering the ratio C_1/C_2 and/or R_1/R_2 .

Component changes

Useful range of supply +1 to +6V (exceeds rating, not guaranteed).
 Frequency stability: +2% for 1V increase in supply, -3.5% for 1V decrease.
 Useful range of C_1 and C_2 : 220pF to 66μF (p.r.f. ≈ 1.4Hz).
 Mark-to-space ratio: 6:8:1 (C_1 :100nF, C_2 :22μF) to 1:8:5 (C_1 :100nF, C_2 :220pF), $V_{out(max)}$: 1.8V.
 Useful range of R_1 and R_2 : 2.2kΩ $V_{out(max)}$: 2.8V to 33kΩ (V_{out} distorted in "0V" region).
 Complementary square wave is available at V_{out2} .
 At either output $V_{out(max)}$ falls by 10% when loaded with 4.7kΩ.

Circuit modifications

● As p.r.f. and mark-to-space ratio depend on the $C_1 R_1$ and $C_2 R_2$ time constants, a variable-frequency square wave is obtained by switching in different, but equal, values of capacitance and varying the p.r.f. continuously with R_1 and R_2 in the form of ganged potentiometers. See circuit left, where R_1, R_2 are 2.2kΩ and R_3, R_4 are 22kΩ. If only one resistor is variable, the mark-to-space ratio is variable but so also is the p.r.f.

● A modification allows the mark-to-space ratio to be made greater or less than unity by adjusting the position of the slider of R_5 (middle circuit) without changing the p.r.f. since $f = 1/T$ and $T = t_1 + t_2$. Hence, with $C_1 = C_2 = C$ and $R_1 = R_2 = R$ then $T \propto C(R + R_x) + C[R + (R_5 - R_x)] \propto C(2R + R_5)$ which is independent of R_x .

● Circuit on right shows modification to use only one capacitor. Useful ranges of C_1 and R_6 are 100pF to 100μF and 470 to 10k respectively. The circuit may be externally synchronized by positive pulses at Tr_1 or Tr_4 base and locks over a frequency range of at least 2:1. Minimum trigger pulse amplitude about 500mV, minimum trigger pulse width about 200ns.

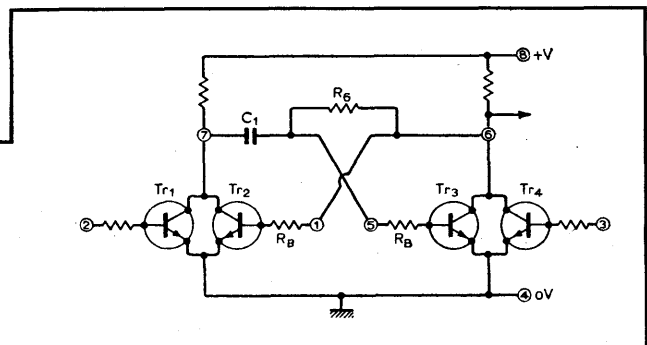
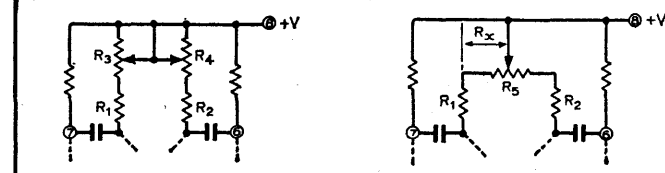
Further reading

Fenwick, P. M., Pulse generator using r.t.l. integrated circuits, *Radio and Electronic Engineer*, 1969, pp. 374-6.

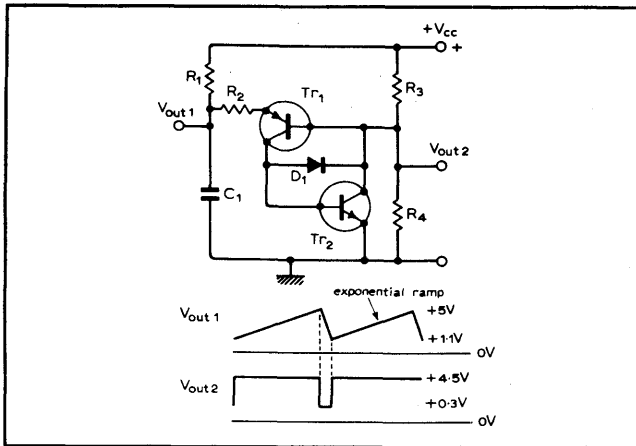
Cross references

Series 8, cards 8, 12.

Circuit modifications



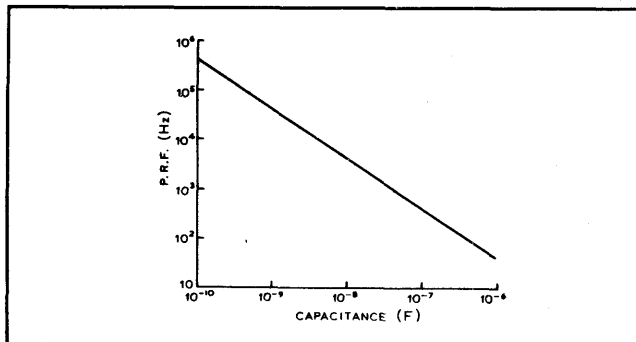
Complementary astable circuit



Typical performance

Supply: +9V, 4.5mA
 Tr₁: BC126; Tr₂: BC125
 Diode: HP5082-2800
 R₁: 27kΩ; R₂: 47Ω

R₃, R₄: 1kΩ, C₁: 10nF
 P.r.f. 6.1kHz
 Mark-to-space ratio:
 49:1
 Rise time of V_{out2}: 1.2μs



Circuit description

When the supply is connected, Tr₁ base and Tr₂ collector are at a potential determined by the ratio R₄/R₃, which could be in the form of a potentiometer to set the upper level of V_{out1} and V_{out2}. The p.d. across C₁ is zero, so the base-emitter junction of Tr₁ is reverse-biased and both transistors are cut off. Capacitor C₁ begins to charge exponentially with time constant C₁R₁ causing the p.d. across it to rise towards +V_{cc}. When the capacitor voltage slightly exceeds the base potential of Tr₁ the base-emitter junction begins to be forward-biased, significant conduction occurring when the capacitor voltage is approximately 0.5V more positive than Tr₁ base.

Positive feedback, due to the interconnection of the bases and collectors of the complementary pair of transistors, ensures that this transition to the on-state is very rapid. Thus

C₁ discharges through Tr₁ and Tr₂ with R₂ providing a discharge current-limiting action. Diode D₁ prevents the transistors saturating and ensures that the circuit can re-cycle.

The capacitor does not completely discharge, but as the current in the transistors falls the loop gain around Tr₁ and Tr₂ reduces to a value that cannot maintain conduction, which ceases when Tr₁'s emitter voltage falls to about 1V. Both Tr₁ and Tr₂ rapidly switch off allowing C₁ to recharge through R₁ and V_{out2} returns to its initial value determined by R₄/R₃. During the discharge of the capacitor, a narrow negative-going pulse is obtained at the junction of R₃ and R₄ due to the conduction of Tr₂.

Component changes

Useful range of supply: +2 to +18V.
 Useful range of C₁: 100pF to 1,000μF.
 Minimum load resistance at V_{out2} ≈ 220Ω.
 Frequency stability: +0.3%/V increase in supply.
 Tr₁: ME0413, 2N3906, BCY71
 Tr₂: ME4103, 2N3904, BC107.

Circuit modifications

- Replacing R₃ and R₄ by a potentiometer across the supply changes the value of the capacitor voltage required to trigger the transistors into conduction and hence controls the period and amplitude of the output waveforms for given values of C₁ and R₁.

- Narrow positive-going pulses in antiphase with those at V_{out2} can be obtained by including a small resistor in series with Tr₂ emitter to the 0-V rail.

- The 'exponential' waveform, V_{out1}, can be made into a more linear sweep by replacing R₁ with a constant-current source. This sweep output can be extracted without significant loading by using an emitter follower. Linearity of the sweep output may be improved by splitting R₁ into R_{1a} and R_{1b} and bootstrapping their junction with C₂, as shown left, where C₂ should be of the order of 100μF.

If D₁ is required to be a silicon diode, additional silicon diodes D₂ and D₃ should be added as shown centre where all diodes could be of the 1N914-type.

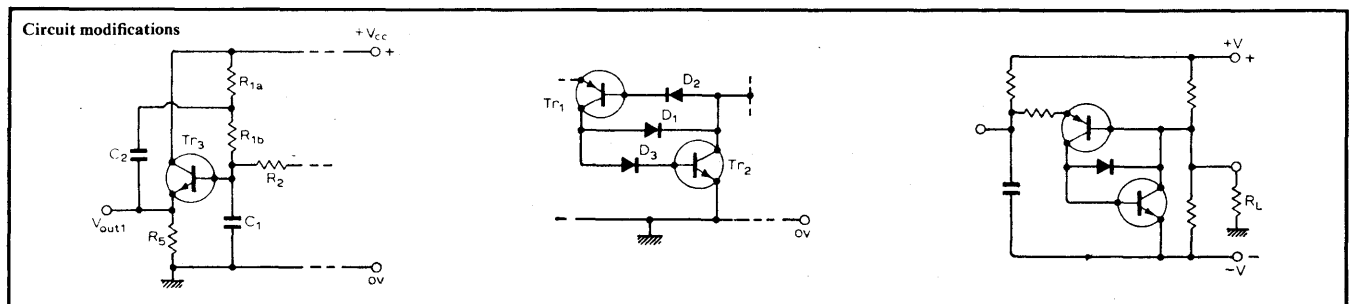
- A dual-supply version of the circuit, which is otherwise identical with the single-supply form, is shown right where both outputs are taken w.r.t. ground. Both outputs can then be made to switch between more widely-varying levels and by adjusting the ratio R₄/R₃ to set Tr₁ base to zero volt in the off-state, negative pulses may be obtained at V_{out2}.

Further reading

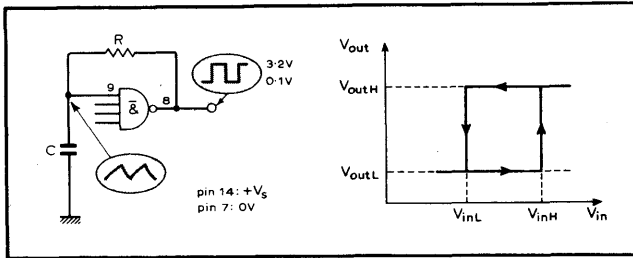
Hemingway, T. K., *Electronic Designer's Handbook*, section 15, Business Publications, 1967.
 Thomas, H. E., *Handbook of Electronic Circuit Design Analysis*, Reston Publishing Co. Inc., 1972, pp. 41-7.
 Coers, G., Astable multivibrator needs only one capacitor, *Electronics*, vol. 46, 18 Jan. 1973, p. 171.

Cross references

Series 2, cards 5, 12. Series 3, card 6.
 Series 6, card 8. Series 8, card 1.



T.t.l. Schmitt astable circuit



Typical performance

IC: $\frac{1}{2}$ SN7413

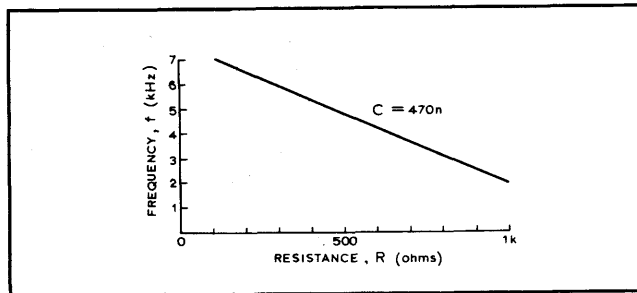
Supply: 5V

R: $330 \pm 5\%$ C: $220\text{nF} \pm 5\%$

f: 10.8kHz

Mark-to-space ratio:

0.5:1



Circuit description

This circuit is internally constructed to behave as a Schmitt trigger, i.e. having two distinct output states, switching between them according to the voltage at the input, with a constant hysteresis between the input levels for switching. The relationship between the input and output states is shown above. The output remains low for $V_{inH} > V_{in} > V_{inL}$. When the input level exceeds V_{inH} , the circuit enters its active region, the action is regenerative, and the output becomes V_{outH} and would remain at this level until the input voltage was reduced to less than V_{inL} .

In the circuit, a convenient starting point of the output high and the input low may be assumed. The capacitor will tend to charge up towards the output voltage, but when the capacitor

voltage reaches the transition level for the i.c., the output falls to near zero voltage. The capacitor then discharges through R until its potential reaches that at which the reverse of the output states occurs, where the output again goes high.

Component changes

Useful range of R: 220 to 1000Ω . Astable will not function for $R \geq 1.5k\Omega$.

Useful range of C: 2.2nF to $22\mu\text{F}$.

Useful range of supply: 4.5 to 5.5V. Operation outside this rated range is possible, but performance not guaranteed. Typically with R: 680Ω and C: 470nF , frequency range is 1.9 to 2.4kHz for supply ranging from 3.5 to 7V.

For supply of 4.5 to 5.5V, frequency stability is approximately $\pm 3\%/V$.

Circuit will supply loads from infinity down to $1k\Omega$, with a reduction of frequency of less than 2.5%.

Circuit modifications

- All four inputs of the nand gate may be paralleled or the unused ones may be taken to $+V_s$.
- One or more of the nand-gate inputs may be grounded to gate the trigger off. This holds the output permanently in the high or logic 1 state (circuit left).
- Two such circuits operating at different frequencies may be locked by capacitive coupling between junctions of C, R elements. The coupling capacitor might be typically $C_1/10$, where C_1 is the smaller of the two multivibrator capacitors (circuit middle).
- A further interconnection is shown right. When V_{out} is high C_2 charges at a faster rate when the first oscillator's output is high than when it is low. Similarly, the discharge rate of C_2 depends on whether one or both outputs are low, thus giving alternate signal outputs of low mark-to-space ratio, followed by high mark-to-space ratio.

Further reading

Texas Instruments: 1971 Seminar Slide Book.

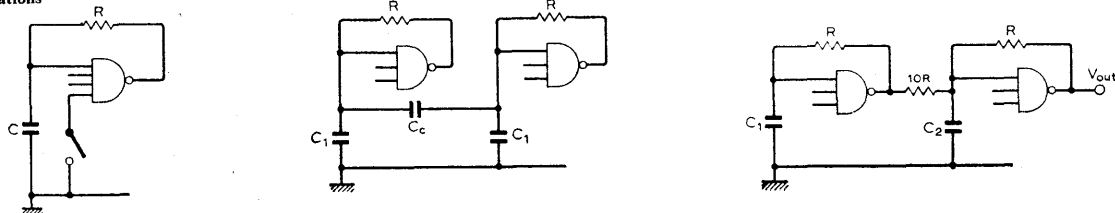
Texas Instruments: SN7413 Data Sheet.

Cross references

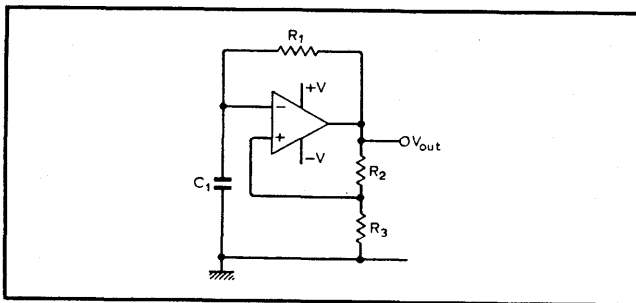
Series 2, cards 3, 8.

Series 8, card 10.

Circuit modifications



Operational amplifier astable circuit



Typical performance

IC: 301

Supply: $\pm 15V$

C_1 : $4.7nF \pm 5\%$

R_1 : $4.7k\Omega \pm 5\%$

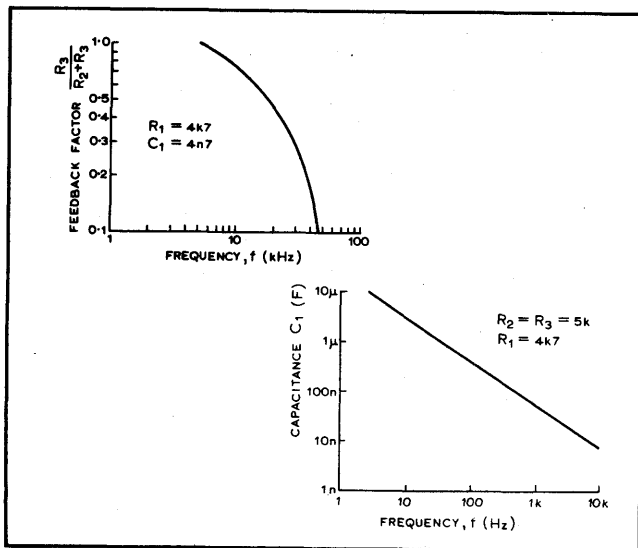
R_2, R_3 : $5k\Omega \pm 5\%$

Output square wave:

28V pk-pk

Slew rate: $8V/\mu s$

Variation of frequency with feedback factor and capacitance C_1 shown on graphs.



Circuit description

The circuit shown uses an operational amplifier where the output switches between the positive and negative saturation levels of the amplifier, giving a square wave output. The period of the waveform depends on the time constant $C_1 R_1$ and the feedback factor, determined by the ratio of $R_3/(R_2 + R_3)$. Assume the output has switched to the positive saturation level; the voltage at the non-inverting input is $+V_{sat} R_3/(R_2 + R_3)$ and the voltage at the inverting input is negative with respect to this value. However capacitor C_1 now begins to charge towards $+V_{sat}$, but when the capacitance voltage

is almost equal to the feedback voltage, the amplifier comes out of saturation, and the regenerative action due to the positive feedback drives the amplifier quickly into negative saturation before the capacitance voltage can alter. C_1 will now charge towards $-V_{sat}$, but again a rapid transition to the positive saturation state will occur when the voltage across C_1 reaches $-V_{sat} R_3/(R_2 + R_3)$, and the cycle repeats. The duty cycle of this astable circuit is almost independent of the pulse repetition frequency, because the threshold levels are fairly well specified to each op-amp.

Component changes

Useful range of R_1 : 6.8k to 2.2k Ω .

Useful range of C_1 : 10 μF to 4.7nF for $R_1 = 4.7k\Omega$.

Frequency stability: For C_1 : 22nF, R_2, R_3 : 5k Ω , R_1 : 4.7k Ω , supply of $\pm 15V$ and $f = 4470Hz$, decreasing supply to $\pm 10V$ reduces frequency by $< 1\%$.

Operation possible down to $\pm 3V$; frequency down by 8%. Any other operational amplifier may be used, e.g. 741, but frequency range restricted because at higher frequencies waveform becomes trapezoidal. Typically, for $R_1 = 2.2k\Omega$, $R_3/(R_2 + R_3) = 0.7$ and C_1 from 10 μF to 220nF, frequency in the range 24Hz to 1.8kHz. Slew rate 0.6V/ μs .

A comparator such as the 72710 will give an output pulse excursion of -0.5 to $+2.8V$ for supplies of $+12V$ and $-6V$. For R_2, R_3 : 5k Ω , useful range of R_1 is 1.5k to 6.8k Ω and C_1 47pF to 22 μF giving frequencies in the range 630kHz to 3Hz.

Circuit modifications

- Interchange C_1, R_1 and the input connections as shown left. For similar component values as in main diagram, frequency is reduced to approximately one third. Note that the derivative of square-wave output is obtained at the non-inverting input.

- Output levels may be clamped for driving t.t.l. loads by connecting a zener diode/resistance network across the output. Clipping at much lower current levels is possible with some amplifiers (e.g. 301), where access is available to the drive point of the output stage. An adjustable arrangement is shown in the middle circuit.

- An unequal mark-to-space ratio may be obtained by using the circuit shown right. The two resistors R_{1a} and R_{1b} are selected by the switching action of diodes D_1 and D_2 , D_1 conducting when the output is negative, and D_2 when the output is positive.

Further reading

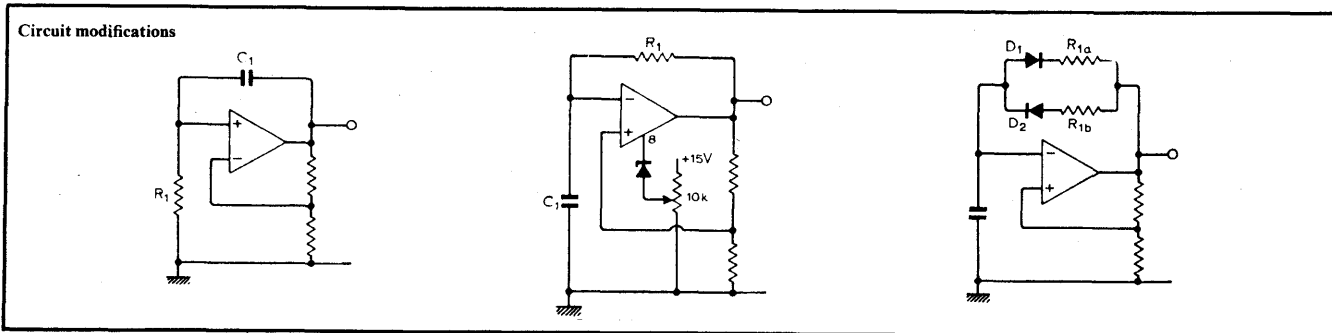
Clayton, G. B., Operational Amplifiers, *Wireless World*, vol. 75, 1969.

Shah, M. J., Feedback pot extends multivibrator duty cycle, *Electronics*, September, 1971, p. 62.

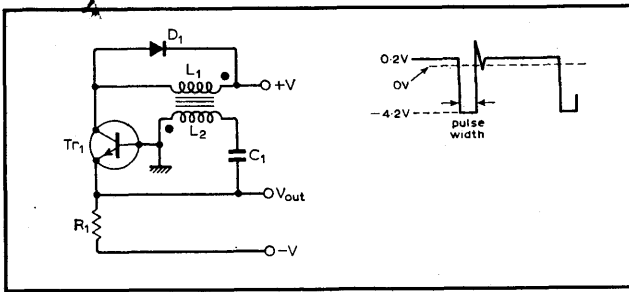
National Semiconductor application note AN4-1.

Cross reference

Series 3, card 5 Series 8, cards 10, 12.



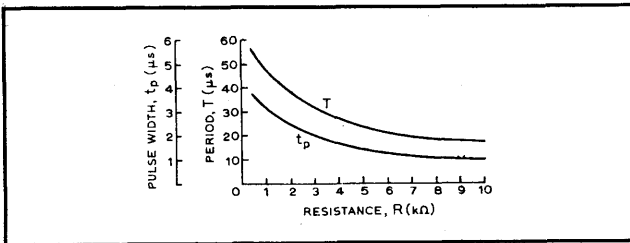
Astable blocking oscillator



Typical performance

Supply: +10V, 860 μ A,
-3V 1.1mA
Tr₁: BC125, D: SD2
R₁: 6.8k Ω ; C₁: 4.7 μ F
L₁: 30 turns of 36 s.w.g.
en. Cu

L₂: 15 turns of 36 s.w.g.
en. Cu; both on FX2049
ferrite core.
P.r.f.: 46.6kHz
Pulse width: 1.15 μ s



Circuit description

Many multivibrators have their timing determined by the interval for which an energy storage element holds an active device in the off state. The output pulse is then available at a high output resistance point, at a low power level, and its rise and fall times are significantly influenced by stray capacitance. The blocking oscillator is an example of circuits which overcome these problems by timing the output pulse within the low output resistance, high-current, saturation region by use of a transformer to provide positive feedback with a loop gain greater than unity. Successful design depends on correct choice of the transformer, which should have small stray capacitances and a magnetizing inductance much larger than its leakage inductance. These requirements can be met by making the transformer physically small, interleaving the windings and using a high-permeability core.

At switch-on the base-emitter junction is forward-biased and the collector current rapidly rises to almost equal the emitter current which depends on R₁ and -V. The transformer ensures that a much larger emitter current flows to saturate Tr₁ and C₁ charges in a direction that reverse-biases the base-

emitter junction causing Tr₁ to cut-off. A very narrow pulse is generated and the circuit will not regenerate until C₁ has discharged through R₁. When Tr₁ cuts off D₁ protects the base-collector junction from the large induced e.m.f. in L₁ and restricts V_{CB} to +V. Capacitor C₁ should be large enough to ensure that the magnetizing inductance of L₁ controls the pulse width and C₁ controls the off-time. The pulse width depends on C₁ rather than L₁ if C₁ is too small.

Component changes

Useful range of +V: +4 to +14V; -V_{min}: -1V.
Useful range of R₁: 470 Ω to 10k Ω .

C₁(min): 470nF.

Minimum load resistance at V_{out}: 2.2k Ω .

Frequency stability: -0.96%/V increase in +V, -0.77%/V increase in -V.

Circuit modifications

- It is often convenient to obtain the output pulse from a third winding L₃ to provide d.c. isolation, a suitable transformer turns ratio for L₁, L₂ and L₃ being n:1:1.

- A diode D₂ can be connected as shown left to prevent saturation of the transistor. As the collector current increases during switch-on, the collector voltage falls until it reaches +V_A causing D₂ to conduct clamping the collector at approximately +V_A. The current shunted from the collector by D₂ decreases as that in the magnetizing inductance of L₁ increases, the on period of Tr₁ ending when the diode current falls to zero.

- Middle left circuit shows a single-supply version of the circuit with R₁ and C₁ in the emitter. The R₁C₁ time constant determines the time for which Tr₁ is off and hence the mark-to-space ratio can be varied by means of R₁. Alternatively, the p.r.f. may be adjusted by means of R₂ which controls the base potential and hence the timing of the off/on transition.

- The R₁C₁ timing components may be connected to the base of Tr₁ as shown in the middle right circuit with a potentiometer R₂ fixing the emitter voltage and hence the time taken for Tr₁ to switch on as C₁ charges through R₁.

- An R-C circuit capable of producing very narrow pulses and very small mark-to-space ratio is shown right. Pulse widths of around 250ns with a mark-to-space ratio of at least 1/100,000 are obtainable with -V of -6V, -V_A of -0.5V, Tr₁: BSX29; Tr₂: BSY17; D₁: EA828; R₁: 100k Ω , R₂, R₄: 50 Ω , R₃: 2M Ω and C₁: 50nF.

Further reading

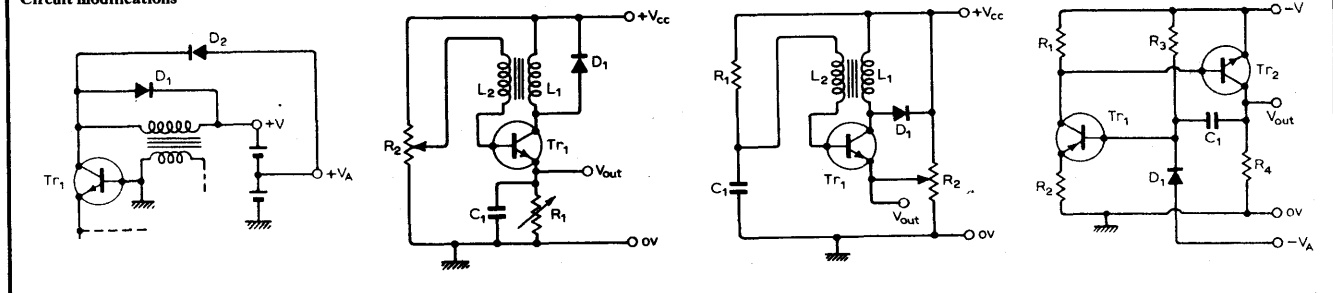
Linwill, J. G. & Mattson, R. H., Junction transistor blocking oscillators, *Proc.I.R.E.*, 1955, pp. 1632-9.

Strauss, L., Wave Generation and Shaping, Chap. 12, McGraw-Hill, 1960.

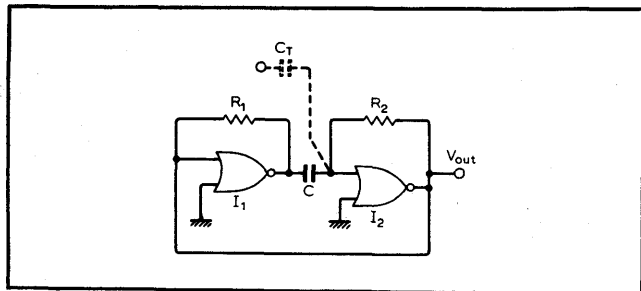
Fontaine, G., Transistors in Pulse Circuits, Chap. 10, Philips, 1971.

Tesic, S., Multivibrator with very small mark-to-space ratio, *Electronic Engineering*, 1967, pp. 671-3.

Circuit modifications



T.t.l. dual inverter astable circuit



Typical performance

Supply: +5V

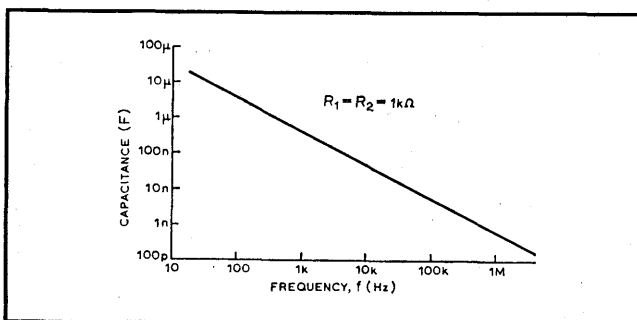
IC: $\frac{1}{2}$ 7402

$R_1, R_2: 1k\Omega \pm 5\%$

$C: 100pF \pm 5\%$

Frequency: 2.78MHz

Stability $> \pm 1\%$ for supply in the range 4.75 to 5.25V for a span of 3kHz to 3MHz



Circuit description

Capacitor C_1 alternately charges and discharges through R_2 because loop gain of the system ensures that the output of the second inverter switches between logic 0 and logic 1 states. When the potential difference with respect to ground at the input of I_2 crosses the critical level. Resistor R_1 is necessary to bias the first inverter I_1 , and thus prevent the possible stable state of inverter I_1 output at almost 0V in the logic zero state, and the output of I_2 in the logic 1 state. Charge and discharge cycles have different durations because the input switching level is not symmetrical with respect to the output 0 and 1 states, and also there is an additional charging path for the input of the second inverter at 0 state. Note that when using nor gates as inverters, the unused input should be tied to logic 0 voltage level.

Component changes

Useful range of C : 100pF to 22μF.

Useful range of R_1 : 220Ω to 1kΩ.

Useful range of R_2 : 150Ω to 1kΩ.

Alternative IC: SN7404 hex inverter.

Circuit operates within the supply range 4.5 to 6V, but not guaranteed outside t.t.l. voltage limits.

If an attempt is made to achieve high frequencies, the range of resistance values is critical. Typical values $R_1: 1k\Omega, R_2: 330\Omega, C: 120pF, f: 6.7MHz$. With nor or nand gates, a spare input is available for external synchronization. Frequency will lock over the range of 4:1 with input pulse widths down to 100ns (positive-going pulse for nor, and negative-going for nand). Capacitive coupling of the trigger source may be used with the inverters of SN7404. Typically $C_T = C/100$. Three separate, harmonically-locked astables can then be produced.

Circuit modifications

- Remove R_2 and connect the capacitor in the feedback loop (circuit left). The third inverter sharpens up the waveform. With supply of +5V, $C: 22nF, R_1: 1k\Omega$ max. to 100Ω min., frequency is in the range 22.5 to 165kHz; mark-to-space ratio approximately 0.6/1.

- Middle circuit uses the SN7404 again, where frequency of oscillation is determined by the propagation delays through the gates. The external capacitance changes the delay associated with two gates and thus alters the frequency. Useful range of C : 1 to 10nF. Frequency 4 to 0.5MHz. Waveform essentially square, but deterioration evident above about 3MHz. Frequency stability fairly poor. Approximately $\pm 10\%/V$.

- Tuning resistors in the middle network comprise the integrated circuit resistors. With perhaps resistance variations of $\pm 20\%$ from device to device, and a like tolerance over the temperature range -55 to $+25^\circ C$, the possibility of frequency and pulse width variations exists. This effect can be minimized for a given output by connecting precision external components as shown right. Output frequency may then remain within $\pm 5\%$ for device or temperature changes. Typically R should be $1k\Omega \pm 1\%$.

Further reading

Malmstaat and Enke, Digital Electronics for Scientists, Benjamin, 1969.

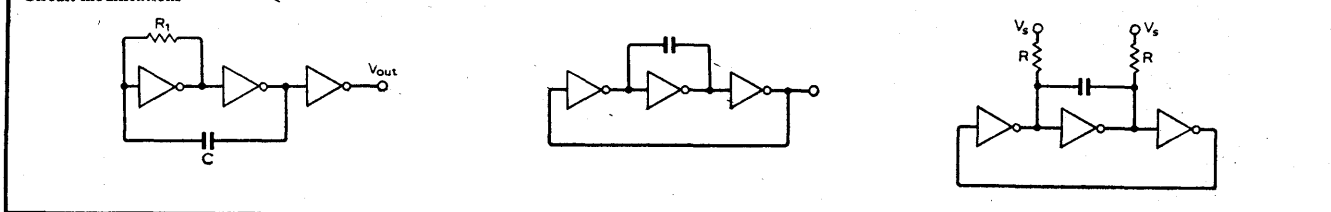
Wide range multivibrator costs just 25c to build, *Electronics*, 1971, p. 59.

MDTL Multivibrator Circuits, Motorola application note AN-409.

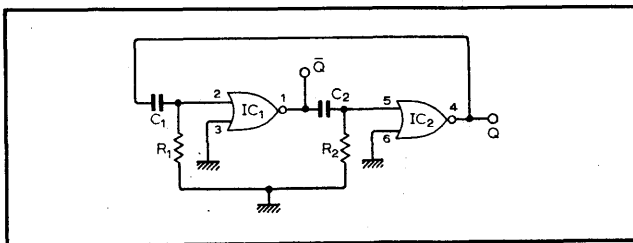
Cross reference

Series 8, card 8.

Circuit modifications



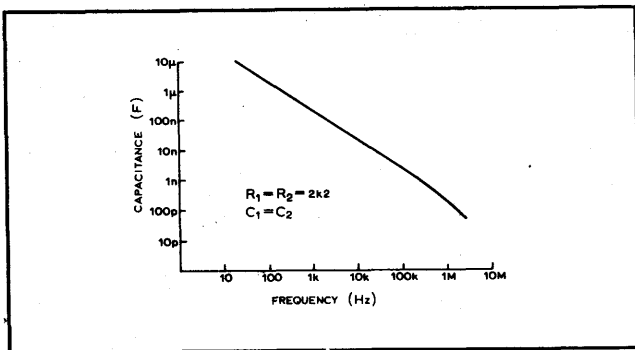
Coupled logic gates astable circuit



Typical data

Supply: +5V
 ICs: $\frac{1}{2}$ SN7402N
 $R_1, R_2: 2.2k\Omega \pm 5\%$
 $C_1, C_2: 0.1\mu F \pm 5\%$
 Frequency: 2015Hz

Mark-to-space ratio:
 1.27:1
 Pulse excursion: 0.3 to 3.2V
 Connect unused inputs to ground for inverter operation.



Circuit description

The values of the current sinking resistors R_1 and R_2 are critical in this type of circuit, which uses nor logic gates in a cross-coupled mode. It is possible for both gate inputs to sink logic 0 level input currents simultaneously, and this produces a stable state in which both outputs are at logic 1. To avoid this choose values of R_1 and R_2 so that the gate input levels are near the logic threshold level; as the capacitors go through their charging cycles, one gate will be above and one below the threshold level. Assume the Q output has changed from the 0 to 1 logic level of approximately +3V due to the input having reached the threshold value. This output transition is coupled through the capacitor C_2 to make the input of the first gate high, and hence the output Q is low or logic 0. As C_2 charges up towards the positive supply via R_2 , the voltage across R_2 and hence the input level at the first gate decreases. At the same time C_1 charges via the base resistor of the input transistor of the gate. The output will change state at a time dependent on whichever gate input first crosses the threshold level. Output \bar{Q} will then be high (approximately +3.0V) and

Q low. Capacitor C_1 will now tend to charge in the opposite direction towards d.c. supply via R_1 , and C_2 also charges in the opposite direction via the input transistor of the first gate until the outputs change state, then the cycle repeats. Frequency of oscillation is determined by C_1, C_2, R_1 and R_2 . If $C_1 = C_2 = C, R_1 = R_2 = R$, the frequency is approximately $1/2\pi CR$ Hz, where C is in farads and R in ohms. Provided resistors are carefully chosen to ensure self-starting, a wide range of frequencies are available by altering C_1 and C_2 . This type of circuit using standard gates or inverters does not provide stable frequencies as the threshold voltages depend on temperature and supply voltage.

Circuit changes

Use MC7402F, though note pin numbers different. Useful range of R_1, R_2 is restricted to ensure that circuit starts: 2.2 to 3.3k Ω . Useful range of C_1, C_2 : 27pF to 10 μ F with the above resistor values. With $C_1 = C_2 = 0.1\mu F$, and a supply of +5V, a variation of supply voltage of $\pm 5\%$ produces a percentage frequency change of +5% or -9% respectively. Nand gates may be used in place of nor gates. In this case, unused pins should be connected to the positive supply line. Waveform of basic astable circuit is improved when the output is applied through an additional gate. Mark-to-space ratio adjustable by having different values of C_1 and C_2 .

Circuit modifications

- Range of resistance values for R_1 and R_2 ensuring self-starting increased slightly, if R_2 is taken to output (left). Range then 1.5 to 3.3k Ω .
- Emitter-coupled logic gates in the middle configuration can provide a high frequency signal, but the component values tend to be critical. IC₁, IC₂, IC₃: $\frac{1}{2}$ /MC1011 quad-nor gates using 1.5k Ω pull-down resistors; supply: -5.2V. Unused input pins connected to -V_s. Repetition frequency 25MHz. Waveform improved by third gate.
- Arrangement shown right would use a single quad two-input nand gate, useful C range being 10pF to 1 μ F. Self-starting problem may be overcome with an extra gate, but a 3-input nand gate then required (cf. Mullard).

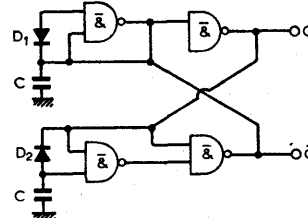
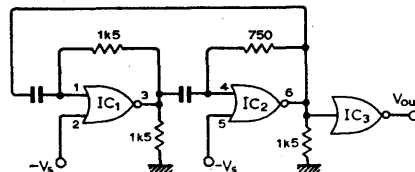
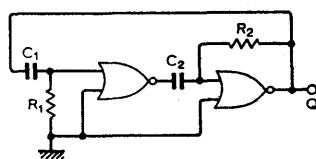
Further reading

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 2MHz Square Wave Generator uses two TTL gates, p. 110, 400 ideas for design, vol. 2, Hayden.
 Simple clock generator has guaranteed start-up. *Electronic Design*, vol. 13, 1971, p. 86.

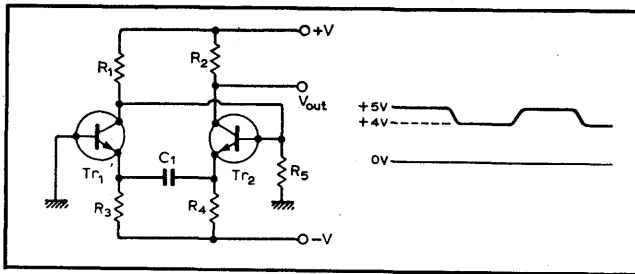
Cross reference

Series 8, cards 7 & 2.

Circuit modifications



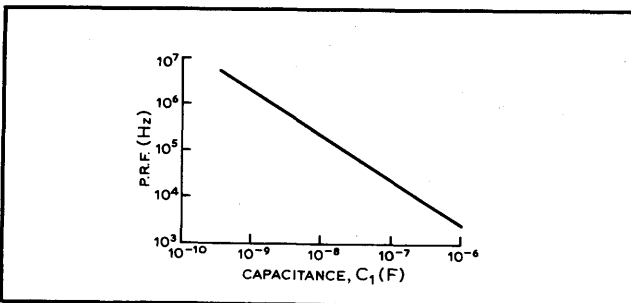
Emitter-coupled astable circuit



Typical performance

Supplies: +5V, 6.4mA;
-5V, 2.5mA
Tr₁, Tr₂: 2N706
R₁, R₂, R₅: 470Ω
R₃: 3.3kΩ; R₄: 4.7kΩ

C₁: 1.5nF
P.r.f.: 1MHz
Mark-to-space ratio:
1.04:1
Rise and fall time:
≈ 30ns



Circuit description

Compared with a conventional saturating, cross-coupled astable circuit, this emitter-coupled circuit uses a single timing capacitor, is capable of producing an improved output waveform, can operate at higher frequencies and can be designed to provide a much better frequency stability. The higher switching speeds are obtained because neither transistor is allowed to saturate and the output waveform does not switch between wide limits.

Consider the circuit to be in the state of Tr₂ off and Tr₁ on. The emitter current of Tr₁ divides into a component in R₃ and a charging current in C₁ and R₄. At the instant that Tr₁ switches on this charging current produces negligible p.d. across C₁ and a p.d. across R₄ sufficiently large to ensure that Tr₂ is off. As C₁ charges, its p.d. increases and that across R₄ falls until the base-emitter junction of Tr₂ becomes forward-biased. Transistor Tr₂ begins to conduct and the emitter current of Tr₁ falls causing the collector potential of Tr₁ and the base potential of Tr₂ to rise. This action causes Tr₂ to conduct more heavily and Tr₁ to switch off. This sequence is then repeated with Tr₂ emitter providing the current to charge C₁ through R₃ until the switching action restores the circuit to its original state of Tr₂ off and Tr₁ on.

Tr₁ will be on for a time:

$$t_1 = C_1 R_1 R_5 (R_3 + R_4) / R_3 (R_1 + R_5) + R_1 R_5$$

and Tr₂ will be on for a time:

$$t_2 = \frac{C_1 (R_3 + R_4)}{\left[\frac{+V}{|-V| - V_{BE(on)}} \right] \left[\frac{R_4}{R_1} - \frac{R_5}{R_1 + R_5} \right] + \frac{R_4 (R_1 + R_5)}{R_1 R_5} - \frac{R_4}{R_3} - 2}$$

Component changes

Useful range of +V: +2 to +14V.

Useful range of -V: -2 to -10V.

Useful range of R₁, R₂ and R₅: 220Ω to 4.7kΩ.

R_{3(min)} 1kΩ (m-s ratio 3.4/1), R_{3(max)} 27kΩ (m-s ratio 1/10).

R_{4(min)} 1kΩ (m-s ratio 1/4), R_{4(max)} 33kΩ (m-s ratio 8/1).

Useful range of C₁: 180pF to 1000μF.

Frequency stability: -1.2%/V increase in +V, -6%/V increase in -V.

Tr₁ and Tr₂: ME4103, 2N708, HE301, BSY95A.

Circuit modifications

- If R₃ and R₄ are replaced by a potentiometer connected across C₁ with its sliding contact taken to the -V rail, the mark-to-space ratio of the output waveform may be varied without changing its frequency.

- The on-time of Tr₁ is independent of the supply voltages, and the on-time of Tr₂ depends on the ratio $\frac{+V}{|-V| - V_{BE(on)}}$. Hence, high frequency stability is obtained if the ratio of the supply voltages is constant. This condition is assured if only a single supply is used as shown left which can provide a frequency stability of 1% for a ±50% change in +V_{cc}.

- R₃ and R₄ in the original circuit may be replaced by constant-current tails. Middle circuit shows a pair of parallel current mirrors making the emitter currents of Tr₁ and Tr₂ equal and controlled by R₈, allowing the frequency to be varied without affecting the mark-to-space ratio.

- The constant-current sources may be voltage controlled, by R₉ in the circuit shown right and the currents in Tr₁ and Tr₂ controlled independently by R_{3'} and R_{4'} respectively.

A larger amplitude output, at slower speed, may be obtained by connecting the original output point to the base of a p-n-p transistor with its emitter connected to the +V rail and its collector returned the -V rail through say a 1-kΩ resistor. If this resistor is connected instead to the 0-V line a t.t.l.-compatible output is obtainable using +V = +5V.

Further reading

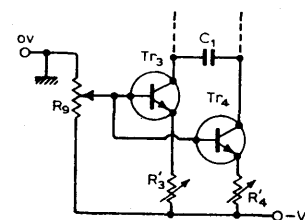
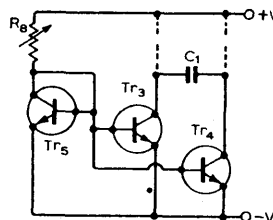
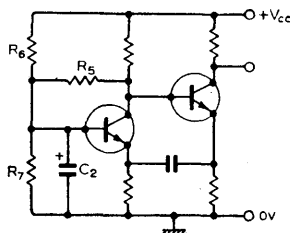
Beneteau, P. J. and Evangelisti, A., An Improved Emitter-Coupled Multivibrator—SGS-Fairchild application report APP-59, 1963.

Electronic Circuit Design Handbook, TAB Books, 1971, pp. 86/7.

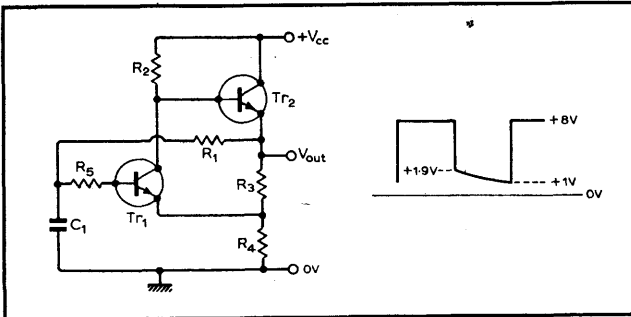
Cross reference

Series 3, card 2.

Circuit modifications



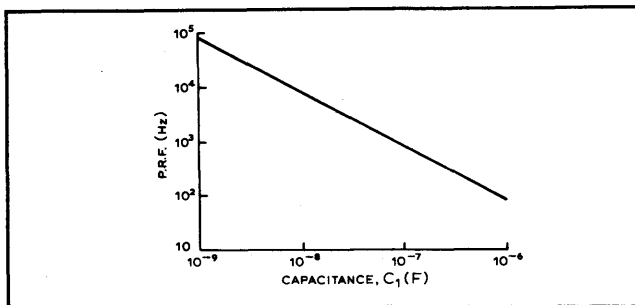
Discrete-component Schmitt astable



Typical performance

Supply: +9V, 5.6mA
 Tr₁, Tr₂: 2N706
 R₁: 10kΩ; R₂: 4.7kΩ
 R₃, R₄: 470Ω

R₅: 2.2kΩ; C₁: 10nF
 P.r.f.: 9.83kHz
 Mark-to-space ratio:
 1.13/1
 Rise time: 400ms
 Fall time: 300ns



Circuit description

This circuit is a Schmitt trigger with overall feedback provided via R₁ and C₁. Consider C₁ to be uncharged, then when the supply is connected Tr₂ emitter is initially at 0V but Tr₂ immediately conducts due to the base drive through R₂. Thus Tr₂ emitter rapidly switches to a level close to +V_{CC} and, with R₃ = R₄, Tr₁ emitter rises to half this value. However, Tr₁ remains cut off due to the lack of base drive. Capacitor C₁ begins to charge "exponentially" through R₁ aiming to reach the emitter potential of Tr₂, but when the capacitor voltage exceeds that at Tr₁ emitter the capacitor begins to discharge mainly through R₁, R₃ and R₄ and partially through R₅, Tr₁ base-emitter junction and R₄, thus driving Tr₁ on and into saturation.

The collector potential of Tr₁ falls to a low value as also does Tr₂ emitter, their being insufficient p.d. available to keep Tr₂ in conduction so that it switches off. C₁ continues to discharge until Tr₁ comes out of saturation, when its collector potential

rises sharply. This rise is transferred to Tr₂ emitter, which begins to conduct, and hence to the emitter of Tr₁. This positive feedback rapidly cuts off Tr₁ leaving Tr₂ in full conduction until, as C₁ charges again through R₁, the higher potential needed at Tr₁ base to restart the cycle is attained.

Component changes

Useful range of +V_{CC}: +4 to +15V.
 Useful range of C₁: 100pF to 1000μF.
 Frequency stability: -0.46%/V increase in +V_{CC}.
 R_{1(min)} 470Ω (m-s ratio 1/5), R_{1(max)} 1MΩ (m-s ratio 48/1).
 R_{2(min)} 680Ω (p.r.f.: 77kHz), R_{2(max)} 100kΩ (V_{out} reduced to +5.2V).
 R_{3(min)} 47Ω (m-s ratio 5/1), R_{3(max)} 2.2kΩ (m-s ratio 1/3, f ≈ 31kHz).
 R_{4(min)} 47Ω (m-s ratio 1/2.8, f ≈ 33kHz), R_{4(max)} 2.2kΩ (V_{out} reduced to 5V pk-pk).
 Useful range of R₅ ≈ 220Ω to 47kΩ.
 Tr₁ and Tr₂: ME4103, BC107, BC109, 2N3904.
 Observe V_{BE(max)} as well as V_{CE(max)} rating.

Circuit modifications

- If R₁ is not too large the p.r.f. only is affected by adjusting the R₁C₁ time constant; this will not be the case when R₁ reaches a value that is comparable with the relatively high resistance discharge path through R₅, Tr₁ base-emitter junction and R₄. Both the p.r.f. and mark-to-space ratio can be made variable by varying the R₃/R₄ potential divider ratio. Resistors R₃ and R₄ can conveniently be made into a continuously-variable potentiometer or R₄ can be made a voltage-variable resistor, e.g. by use of a f.e.t.
- When a small, but controlled, mark-to-space ratio is required R₁ may be replaced by the resistor-diode combination shown left. Both diodes could be of the 1N914 type. The circuit may be synchronized from an external oscillator by resistive coupling to the emitter of Tr₁, or by capacitive coupling to its base.
- Addition of Tr₃, as shown middle, allows the timing of the output square wave to be more nearly controlled by the R₁C₁ time constant. Tr₁ and Tr₃ form a long-tailed pair so that the junction of C₁ and R₁ are effectively connected to one input of a differential operational amplifier.
- Circuit right shows a similar form of modification which has the merit of allowing V_{out} to swing almost between the levels of the supply rail potentials.

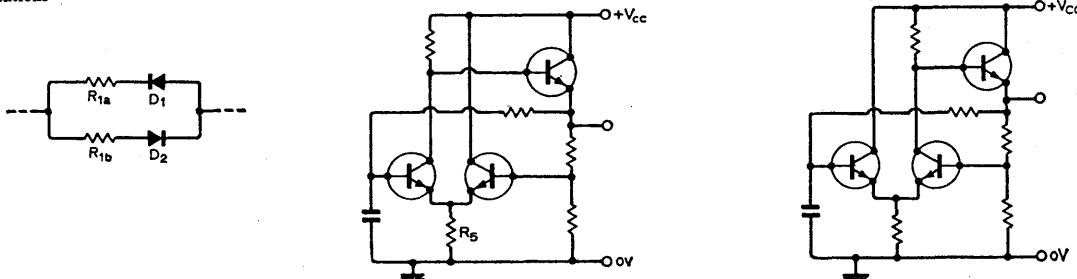
Further reading

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 Sylvan, T. P., The Unijunction Transistor, Characteristics and Applications, General Electric Co., N.Y., 1965, pp. 52/3.

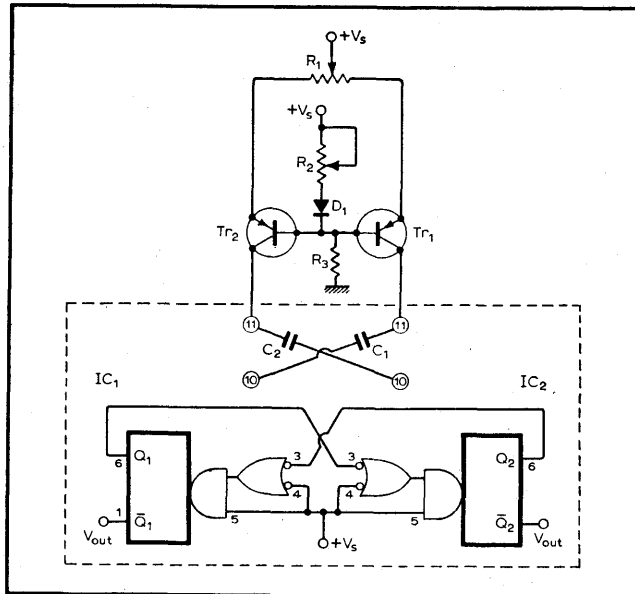
Cross references

Series 2, cards 2, 7.
 Series 8, cards 4, 5 & 12.

Circuit modifications



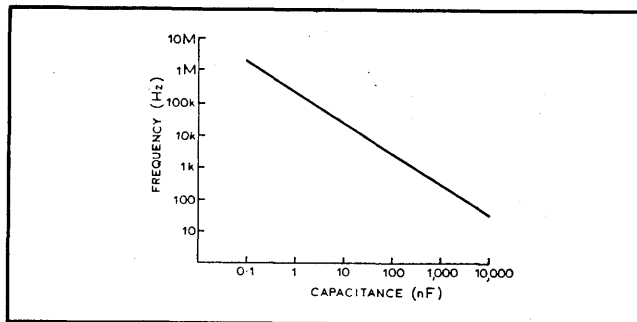
Dual-monostable astable circuit



Typical performance

Supply: +5V, 46.5mA
 ICs: SN74121N
 (monostable)
 Tr₁, Tr₂: ME0413;
 D₁: PS101
 C₁, C₂: 10nF
 R₁: 2kΩ

R₂, R₃: 1kΩ
 Frequency: 28.6kHz
 Pulse excursion: 0.2 to 3.6V
 Rise time: 100ns
 Fall time: 50ns



Circuit description

Astable circuits can be constructed out of cross-coupled monostable circuits provided that the output of one monostable can be used to initiate the timing cycle of the other. The circuit shows the interconnection between two t.t.l. monostable i.cs. The timing capacitors for the two parts of the

period are C₁ and C₂, and these might be equal or different, depending on the need for unity or other mark-to-space ratios. In place of the resistive part of the timing circuit, Tr₁ and Tr₂ provide constant currents, so that the capacitors charge linearly. This allows a ramp waveform to be extracted at pin 11 on either i.c. It has the further advantage that varying the common potential at the bases of Tr₁ and Tr₂ with R₂ allows both charging currents to be varied simultaneously, i.e. a change in frequency without change in mark-to-space ratio. By varying the tapping point on R₁, the balance between the currents in Tr₁, Tr₂ collectors are changed and the mark-to-space ratio is varied with a relatively small change in total period, i.e. in frequency. Diode D₁ gives temperature compensation for the base-emitter potential changes of Tr₁ and Tr₂. Long periods may be attained by lowering R₂ so that the charging currents in the transistors are small, but the period then becomes more temperature and supply sensitive. Independent anti-phase voltages are obtainable at the Q outputs of the two i.cs without any loading effects on the interconnection circuitry.

Component changes

With C₁, C₂: 10μF and minimum setting of R₁, frequency is 1.6Hz. With C₁, C₂: 10nF, and maximum setting of R₁, mark-to-space ratio is variable from 0.03 to 0.98. Frequency stability within ±3% for a supply change of ±0.5 on 5V. Resistive loading may be reduced to 2.2kΩ to maintain pulse height within 90% of maximum level. Absolute minimum load 150Ω where pulse level is then down to 1.9V. For fixed capacitance, frequency range roughly 10/1 by varying R₁.

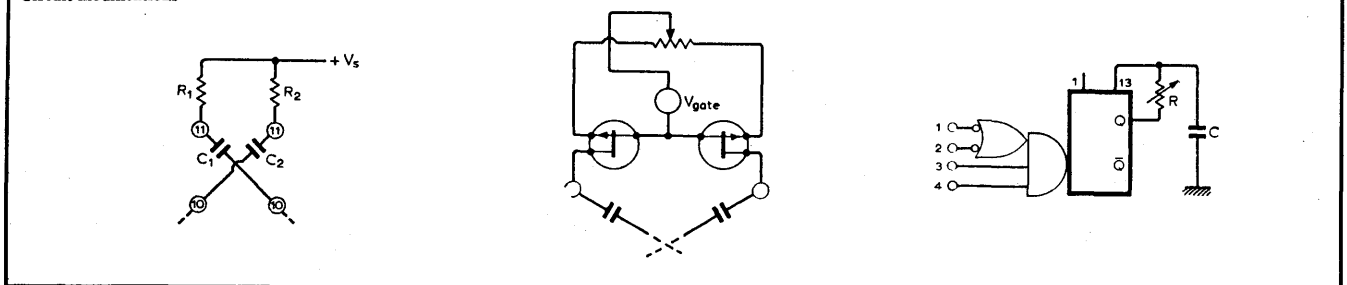
Circuit modifications

- Replacement of Tr₁, Tr₂ by resistors connected from pin 11 on each i.c. to supply line, i.e. replacing the current source by the more normal resistive source, gives non-linear charging of C₁ and C₂, but is a lower cost arrangement (left).
- Any other current source may replace Tr₁ and Tr₂, e.g. p-channel field-effect transistors, with either variable voltage on the gate, or variable resistance in the source (middle circuit).
- Use a retriggerable monostable SN74L122 in the configuration shown right. This monostable has a similar behaviour to a Schmitt trigger when the timing capacitor is driven from the Q output. Pins 1 and 2 may be taken to supply line and 3 and 4 to ground.

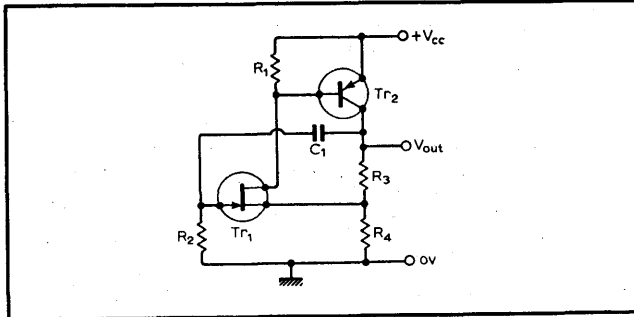
Further reading

Photo-f.e.t.s make multivibrator respond to incident light, in 400 Ideas for Design, vol. 2, Hayden, 1971, p.107. Smith, D. T., Multivibrators with seven-decade range in period, *Wireless World*, Feb. 1972, pp. 85/6.

Circuit modifications



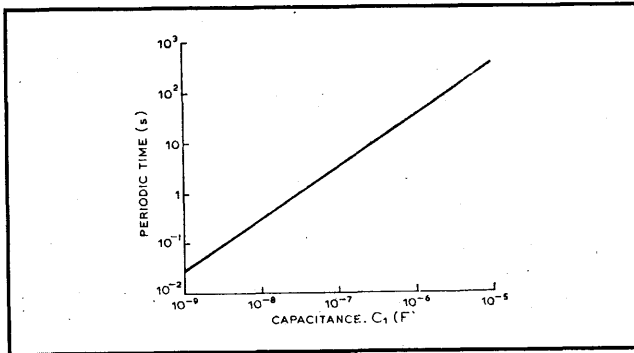
Astable circuit with f.e.t.



Typical performance

Supply: +9V, 760μA
 Tr₁: 2N5457;
 Tr₂: BC126
 R₁: 100kΩ; R₂: 10MΩ

R₃: 3.3kΩ; R₄: 6.8kΩ
 C₁: 1nF
 P.r.f. 31.2Hz
 Mark-to-space ratio:
 1:17:1



Circuit description

When the supply is connected both active devices conduct with currents determined by the negative feedback due to R₃ and R₄. The collector potential of Tr₂ jumps sharply to a level approaching +V_{CC} and the source of Tr₁ jumps towards V_{CC} R₄/(R₃ + R₄). With C₁ initially uncharged, the full positive step at Tr₂ collector is passed to the gate of Tr₁ so that the gate-source junction becomes forward biased by about 500mV and the charging current flows through it to ground via R₄. The initial charging current in C₁ is thus larger than it would have been if C₁ charged simply through R₂. The p.d. across R₂ falls rapidly as C₁ charges through R₄ causing the f.e.t. junction to be reverse biased and the charging time constant of C₁ to charge to the much larger value of C₁R₂. Capacitor C₁ continues to charge until the gate potential of Tr₁ falls below its source potential by an amount that approaches the pinch-off value causing Tr₂ to switch off due to the reduction of base current. The output switches back to virtually 0V as C₁ discharges.

through R₂, R₃ and R₄, where R₂ ≫ R₃ + R₄, until the gate source p.d. allows sufficient drain current in Tr₁ to switch Tr₂ on and the circuit re-cycles. Due to the low reverse-bias gate current of the f.e.t., long time intervals can be obtained between the changes of state using reasonable component values, provided that C₁ is a low-leakage type. The accuracy of these long time intervals however depends on ill-defined value of the pinch-off voltage of the f.e.t.

Component changes

Useful range of supply: +6 to +30V.
 Useful range of C₁: 10pF to 100μF, low-leakage type.
 R_{1(min)}: 4.7kΩ.
 Useful range of R₂: 1 to 200MΩ.
 Useful range of R₃: 2.2 to 33kΩ.
 Useful range of R₄: 330Ω to 10kΩ.
 Minimum load resistance: 220Ω.
 Frequency stability: +1%/V increase in V_{CC}.

Circuit modifications

- Another circuit that can produce output pulses separated by a long time interval is shown left. When the supply is connected, the low-leakage capacitor C₁ charges through R₁ until the gate potential of the n-channel j.f.e.t. reaches the threshold voltage of the programmable unijunction transistor Tr₂ minus V_{GS(off)} of Tr₁. On reaching this gate voltage of +V_{CC}(R₆/R₅ + R₆) - V_{GS(off)}, the gate-source junction of Tr₁ becomes forward biased and C₁ discharges through it to ground via Tr₂ and R₃ providing an output pulse across R₃. Transistor Tr₁ then cuts off and the charge-discharge cycle of C₁ is repeated.

- Caution is necessary in replacing bipolar junction transistors, in circuits that are known to work by field-effect transistors. In the middle circuit Tr₁ of a Schmitt astable has been changed to an n-channel j.f.e.t. but for a given pinch-off voltage and R₁-value there may be insufficient drain-source p.d. to allow the switching action to take place, except by critically adjusting the ratio R₃/R₄. This situation is improved by inserting a zener diode between the drain of Tr₁ and the base of Tr₂.

- Circuit right shows a cross-coupled astable where n-p-n transistors have been replaced with n-channel j.f.e.t.s. This circuit will not switch unless R₃ and R₄ are returned to ground instead of +V_{CC} and should be returned to a slightly negative rail to ensure self-starting.

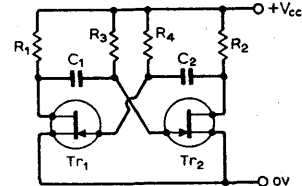
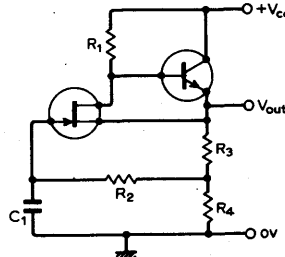
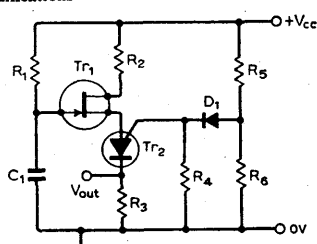
Further reading

FET and UJT provide timing over a wide temperature range, in 400 Ideas for Design, Hayden, 1971, pp. 192/3.
 Watson, J., Introduction to Field Effect Transistors, Siliconix, 1970.

Cross references

Series 8, cards 2, 5 & 10.

Circuit modifications

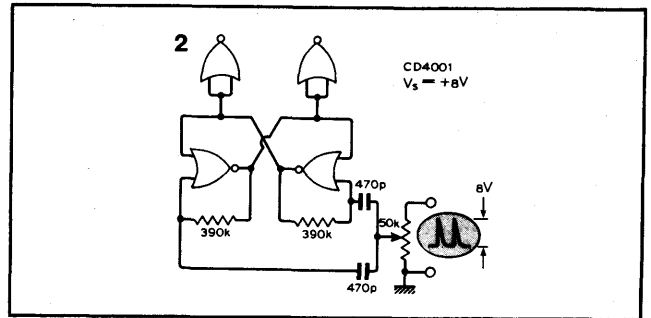
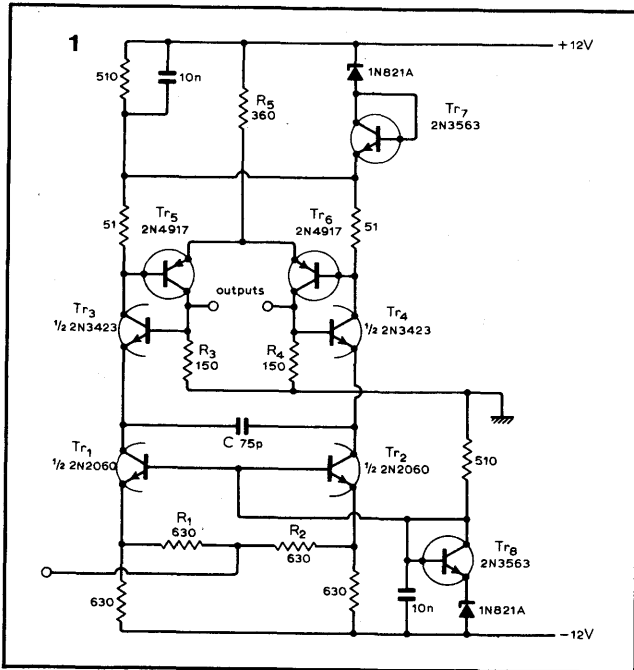


Astables

1. This circuit is a high performance complementary astable. The elements determining the switching functions are the complementary pairs $Tr_{3,5}$ and $Tr_{4,6}$. These act as complementary Schmitt triggers, channeling the current in R_5 via the long-tailed pair formed by $Tr_{5,6}$ into one or other of the load resistors R_3, R_4 . Transistors $Tr_{5,6}$ form part of two distinct circuit functions viz the Schmitt and long-tailed pair switches. The l-t.p. action ensures excellent switching performance since the total current remains constant while the voltage changes required at the bases of $Tr_{5,6}$ to achieve this switching remain small. Timing capacitor C_3 is charged alternately by Tr_1, Tr_2 the other transistor of the pair having

its collector current absorbed by whichever of pair Tr_3, Tr_4 is conducting at that time. Transistors $Tr_{7,8}$ provide temperature compensation for the biasing networks that stabilize the operating currents.

Because of the totally complementary nature of the circuit the outputs are anti-phase square waves of 0 to +3V, with operating frequencies claimed to exceed 50MHz. Frequency control input gives linear control over an unspecified range. Aldridge, S. F. Square-wave generator stresses frequency stability, *Electronics*, vol. 46, May 10, 1973, p.98.



2. This circuit is not itself an astable circuit but is a simple frequency divider that can be added to other astables. Division ratio can be adjusted to any even number from two upwards with increasing dependence on parameters such as supply voltage pulse-height etc. If these two are proportional, as would be the case for an astable operating from the same supply as the divider, then the supply dependence is reduced and ratios up to 30 are claimed to be possible by adjusting the input potentiometer. Operation also depends on the 5-pF gate input which integrates the pulses received from the potentiometer.

Newton, D. C-MOS gate package forms adjustable divider, *Electronics*, March 1, 1973, p.104.

Wireless World Circard

Series 9: Optoelectronics

The relative merit of infra-red/visible light sensors as far as spectral response is concerned, is easily assessed in three illustrations of the article introducing the optoelectronics cards. From Fig. 9 in particular it's clear why cadmium sulphide (CdS) photoconductive cells are used in photographic work. Card 10 gives the various performance and application details, together with those pertaining to silicon and selenium photovoltaic cells. Card 11 shows how to use them in comparing intensities and for integrating intensity over a period of time. Another card, number 7, gives ways of interfacing phototransistors with various logic circuits, pointing out that silicon photodiodes give sufficient current to directly operate c.m.o.s. circuits.

Three cards show how to use light-emitting diodes in various ways, one of them, number 5, giving basic design guidelines. Digital circuits can be driven using one of the six circuits on card 2, covering t.t.l., d.t.l., r.t.l. and c.m.o.s. circuits, while op-amps can be interfaced according to card 6. L.e.ds by themselves can act as voltage level indicators for the range 1.2 to 2.2 volts and higher voltages can be simply indicated by adding a series Zener diode. In many voltage-limiting circuits, l.e.ds can simply replace the limiting diodes, giving simple overload indication.

Some useful data on optical isolators are included on five cards, especially card 8, which points out some characteristics not always quoted by manufacturers.

- Null, level and overload l.e.d. indicators 1
- Driving l.e.ds: digital circuits 2
- Switching with an opto-isolator 3
- Integrated-circuit optoelectronic switch 4
- Characteristics and applications of l.e.ds 5
- Op-amp/comparator driving of l.e.ds 6
- Phototransistor logic circuit drivers 7
- Optically-coupled isolator: static characteristics 8
- Optically-coupled isolator: pulse characteristics 9
- Photoconductive and photovoltaic cells 10
- Light intensity measurement and detection 11
- Choppers and rectifiers 12

Optoelectronics: devices and applications

Three aspects of the link between light and electronic circuitry are considered in this article: detection and measurement of light, generation of light from an electrically operated source, and use of light as an intermediary in some electronic process. Of the wide array of optoelectronic devices available a small number cover a wide range of requirements. More specialized components, such as semiconductor lasers, photomultipliers, must be left to a later series.

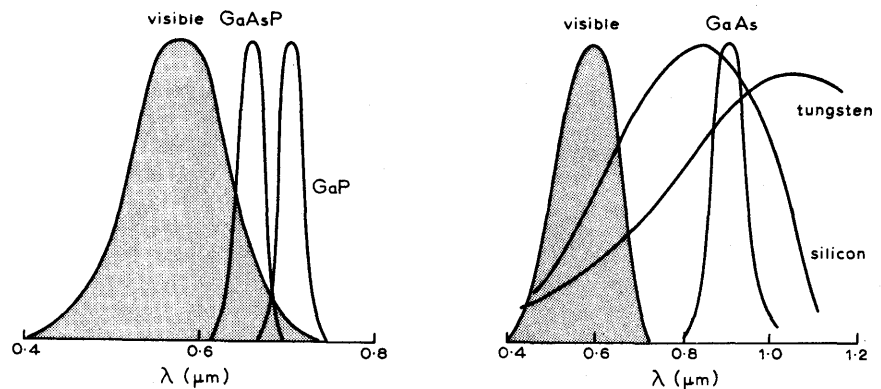
The electronics industry is seeing a production and pricing pattern in one family of optoelectronic devices reminiscent of the t.t.l. war at its fiercest. Light-emitting diodes have changed from an r & d novelty to a consumer component with great rapidity and are already falling to a price level comparable with the transistors used to drive them. They will replace filament and neon lamps for a wide variety of applications and it is on their characteristics that this series concentrates as far as light generation is concerned.

Light-emitting diodes are p-n junctions governed by the same rules as silicon diodes. The choice of materials is decided by the spectrum of light required while the efficiency can depend critically on doping levels. The most common materials are gallium arsenide and gallium phosphide (and other compounds) (Figs 1 & 2) with a current requirement of 5 to 50mA for normal brightness. The resulting p.d. is in the range 1 to 2V (Fig. 3) making the diodes compatible with both linear and digital circuits. Relative intensity of the lamp is an almost linear function of current over a wide range of currents (Fig. 4) falling at higher temperatures.

Response of the human eye is greatest in the yellow/green region of the spectrum, and the gallium arsenide-phosphide devices (Fig. 1) — which emits peak in the red — appear less bright than otherwise comparable-efficiency yellow diodes. Nonetheless, economies of scale dictate the use of the lower-cost red l.e.ds for most applications.

Emission in other parts of the electromagnetic spectrum is possible with suitable materials including the highly specialized semiconductor lasers with a very narrow spectral response. Pulsed operation with low duty-cycle high-current pulses gives the highest efficiency with such devices, and they find application in optical communication systems.

Solid-state light-sensitive devices are based either on p-n junctions (photodiode, photovoltaic cell, phototransistors) or poly-



Figs. 1 & 2. Materials used in light-emitting diodes are chosen to give the desired colour. Although GaAsP diodes are of comparable efficiency to GaP diodes, they appear brighter because eye response is greater in that region (Fig. 1, left).

crystalline materials (CdS photoconductive cells). Light falling on a p-n junction if of short enough wavelength generates hole-electron pairs. Current flows if the p-n junction is short-circuited (Fig. 5) or connected to a low-resistance load; current is proportional to the intensity of light falling on the junction, and the terminal p.d. for a silicon diode may be up to about 0.5V (Fig. 6).

A specialized form of photodiode, the photovoltaic cell, is used as a power source. Selenium, or more commonly now silicon, is the material used and the cells are arranged to have a high surface area to maximize their light catchment. Although having non-linear characteristics, the simple maximum-power-transfer

theorem gives a first order answer for the optimum load (Fig. 7) i.e. that load which maximizes the output power is given by $R_{opt} = V_{oc}/I_{sc}$. The theorem applies to linear systems but gives reasonable results for some non-linearity.

Further increase in load resistance or illumination makes for little extra p.d. as it becomes comparable with the p-n junction barrier potential. In a phototransistor the collector-base junction is illuminated and the transistor amplifies the resulting current. Currents of 10mA and more are possible but the speed of response is slower than for the diode alone when reverse biased. This latter mode is the fastest with a separate d.c. supply providing a high field to sweep

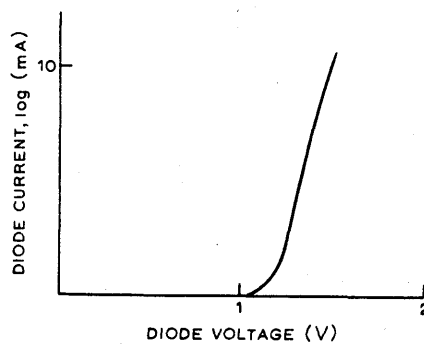


Fig. 3. Light-emitting diodes have a p.d. of 1 to 2V for normal brightness levels, making them compatible with both linear and digital circuits.

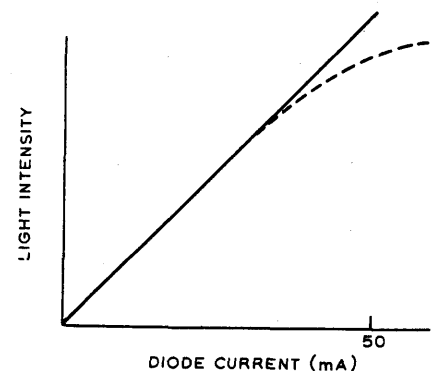
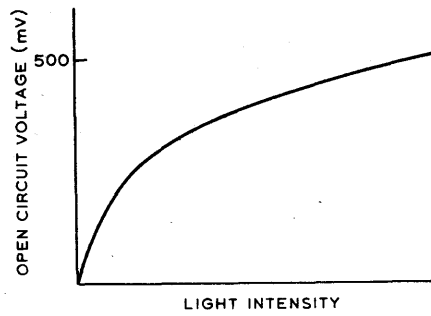
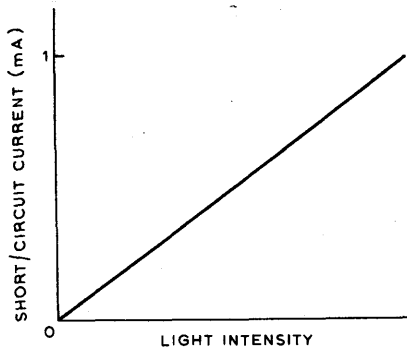


Fig. 4. Lamp intensity is a linear function of current, except at higher temperatures (broken line).



Figs. 5 & 6. In light-sensitive devices based on a p-n junction, current is a linear function of light intensity falling on the junction (Fig. 5, left). Open-circuit voltage can be as high as 0.5V with silicon photodiodes (Fig. 6, right).

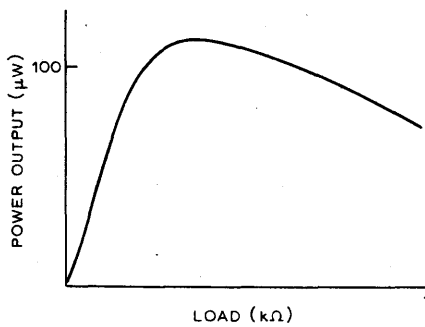


Fig. 7. For selenium or silicon photo voltaic diodes maximum power is delivered for a load given by open-circuit voltage divided by short-circuit current.

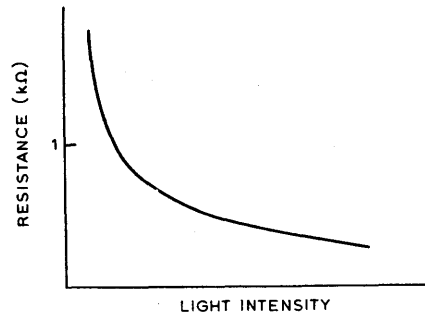


Fig. 8. Photoconductive cells using cadmium sulphide normally have a non-linear light intensity solidus current characteristic although the voltage-current curve is linear.

the carriers out of the junction. The limiting effect on speed of response is often the RC time constant of the external circuit particularly as the currents are low, forcing the use of large resistors where large voltage swings are needed. Any voltage swing up to the supply voltage may be obtained provided that the supply voltage does not exceed the diode reverse-breakdown voltage.

Any other semiconductor device or circuit using a p-n junction may in principle be made into an optoelectronic device with suitable packaging. Thus photothyristors can be obtained which are fired by an optical pulse, though they are presently restricted to relatively low-current applications e.g. for firing a higher power thyristor. In this case the photothyristor needs to have high voltage breakdown characteristics but can be a low-power device supplying pulses of a few tens of mA to the power device.

Photo-Darlington and other transistor combinations are equally feasible where higher output currents are desired in linear circuits, but in many cases it is equally simple to combine a photodiode or phototransistor with other conventional transistors. M.o.s. devices are very useful in optoelectronic circuits, with the threshold voltage varied by the light

intensity. By adapting the m.o.s. processes, complex circuits have been recently produced by i.c. manufacturers, to provide outputs capable of driving small relays and switching at preselected light intensity.

Cadmium sulphide photoconductive cells have a very high resistance in the absence of light, with the resistance

falling rapidly with increasing light intensity (Fig 8). At all light intensities the device has a linear V/I characteristic, and for some devices the variation of conductance with light is itself almost linear. As the spectral response of CdS cells closely matches that of the human eye (Fig 9), they are used in photographic equipment. These cells have to be used with some external power supply, a.c. or d.c., as they do not generate any power. The advantage is that the power controlled in the load can be large enough to operate a wide range of lamps and relays.

A powerful tool for designers made possible by recent development is the photon coupler. This is typically an 8-pin dual-in-line package containing a light-emitting-diode tightly coupled, optically, to a silicon phototransistor. There is no electrical connection and breakdown voltages between the two sections of several kilovolts have been achieved. Similarly the capacitance coupling is small, and these devices are ideal for transmitting signals between circuits where direct electrical connection is impossible or inadvisable. Examples of such situations include firing of thyristors from grounded source and coupling between digital systems where ground noise is a problem.

Optoelectronics has passed the healthy infant stage and is developing into a mature branch of engineering where the choice is wide enough to encourage all designers to extend their skills.

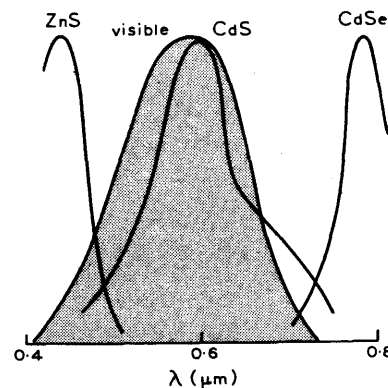
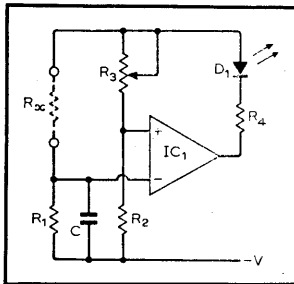


Fig. 9. Because the spectral response of cadmium sulphide cells matches eye response they are widely used in photographic applications.

Null, level and overload i.e.d. indicators

**Typical performance**

Supply: -9V
 IC: 301
 Diode: TIL209
 R_1, R_2 : $10\text{k}\Omega$
 R_3 : $10\text{k}\Omega$ variable
 C : $1\mu\text{F}$
 LED lights for $R_x < R_3$
 Zero offset typically 1 to 2mV —may be trimmed out.

Circuit description

Light-emitting diodes are convenient indicators of out-of-balance conditions in bridges provided the unbalance signal is sufficiently amplified. Using an operational amplifier for which the common-mode input range can include one side of the supply, a simple Wheatstone Bridge can be used for rapidly determining resistance.

In the circuit shown, with $R_1 = R_2$ bridge balance is achieved when R_3 is adjusted to equal R_x . The lower limit of $R_x \rightarrow 0$ (though with zero offset problems in op-amp) while the upper limit is set by R_3 and the minimum input current of amplifier.

As shown, the i.e.d. is off for $R_x > R_3$ which makes bridge convenient for portable use with unknown contacts normally open. Quiescent current is standby current of op-amp plus current in R_2 and supply voltage is not critical. Bridge uses whole of available supply voltage to maximize sensitivity and limit is set by zero offset in amplifier.

Capacitor C minimizes hum pickup effects with hand-held probe. Method can also be applied with rectified a.c. supply if C is omitted and leads/unknown can be well-screened.

Component changes

R_1 : 100Ω to $1\text{M}\Omega$. At lower values current drain is excessive. At high end amplifier input current unbalances bridge.

R_2 : As above. Accuracy of matching of R_1, R_2 or a suitable ratio determines overall accuracy.

R_3 : Range chosen to suit R_x . A linear pot is easier to calibrate.

C: Not critical; used to minimize effects of hum pickup.

IC_1 : Any op-amp if bridge balance is achieved at potentials near midpoint of supply. For R_3, R_x small, amplifier common-mode range must extend near to most positive supply potential. 301, 307.

D_1, R_4 : Not critical. R_4 not required if op-amp has internal current limiting. D_1 : HP 5082-4440, Monsanto MV5094, etc.

Circuit modifications

- While op-amp may be used as null indicator for ground-referred voltages this may require \pm supplies. By using a Darlington pair and a GaAs i.e.d., the transistors will conduct for input voltages close to zero (i.e.d. p.d. $\approx 1.4\text{V}$ at 5mA). For negative inputs transistors non-conducting, all current flows in i.e.d.; for positive inputs all currents diverted into transistors. If input $> +500\text{mV}$ input base-collector becomes forward biased.

- The i.e.d. is a simple fixed voltage level indicator (≈ 1.2 to 2.2V depending on type) provided source can supply sufficient current to illuminate, yet insufficient to damage device. Voltage level can be increased by series zener diode(s).

- The i.e.d.s can replace diodes in voltage limiting circuits while indicating overload. Examples include voltmeter protection with a series limiting resistor to protect i.e.d. For a high resistance voltmeter of say 1V f.s.d. a resistor of $1\text{k}\Omega$ could limit current to i.e.d. safely from inputs up to 50V with small meter error under normal conditions. Limiting of amplifier inputs would also allow indication of polarity of overload or presence of a.c. (both i.e.d.s illuminated).

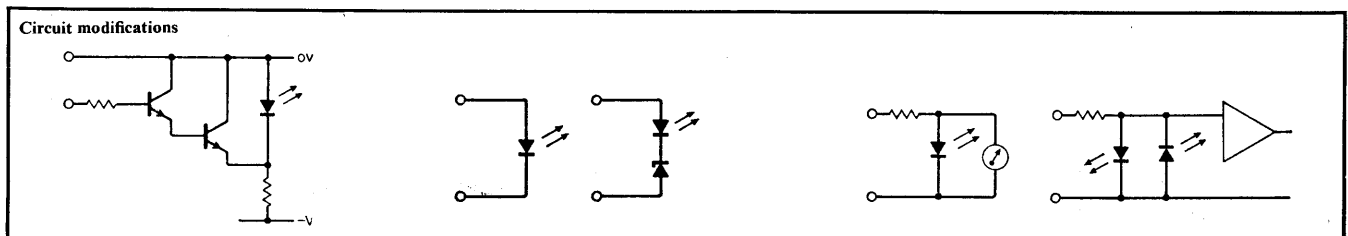
Further reading

Loughmane, M. H., Light-emitting diode pair forms null indicator, *Electronics*, 2 Aug. 1971, p.58.

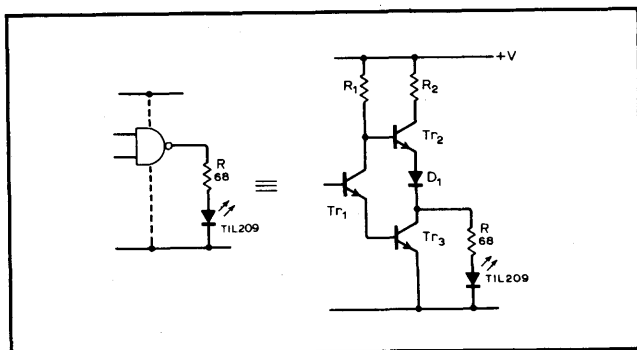
Steen, T. J., Three-i.e.d. circuit indicates a.c. or d.c. and polarity, *Electronic Design*, 25 May 1972, p.68.

Cross references

Series 9, cards 5 & 11.



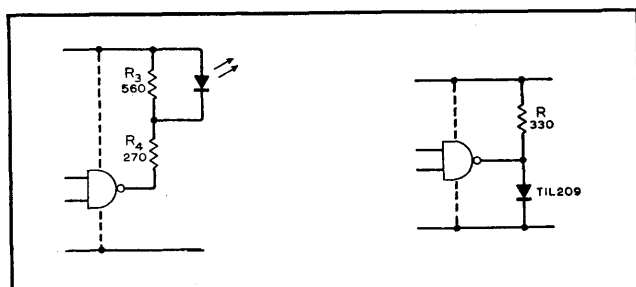
Driving I.e.ds: digital circuits

**Circuit - 1 t.t.l. logic 1 indication**

The totem-pole output stage has both pull-up and pull-down active devices. When Tr_1 is driven into conduction by input logic signals, Tr_3 conducts also. The low saturation voltage of Tr_1 , together with junction p.d.s of Tr_2 and D_1 prevent Tr_2 from being forward biased. The I.e.d. is effectively short-circuited, speeding fall-time of light pulse. With Tr_1 off, Tr_3 receives no bias but Tr_2 is biased via R_1 to positive supply. If higher current allowable in I.e.d., R may be dispensed with.

Circuit - 2 t.t.l. logic 0 indication

In previous circuit internal resistors R_1 , R_2 plus junction p.d.s limit current to I.e.d., simplifying drive conditions. For logical 0 output, Tr_3 becomes near short circuit and R_3 , R_4 limit I.e.d. current while providing passive pull-up to switch I.e.d. off rapidly. This configuration is more wasteful of current (below, left).

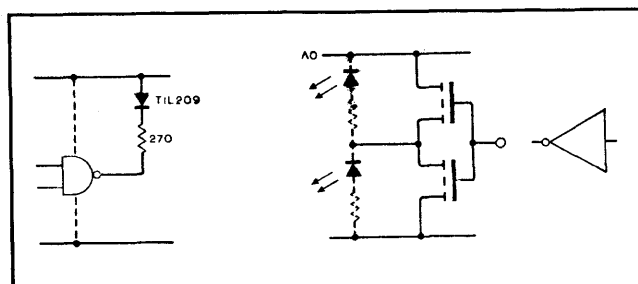
**Circuit - 3 d.t.l. logic 1 indication**

Simpler output circuit of d.t.l. demands load resistor R supplying current to I.e.d. when output transistor is off. R chosen to define current in I.e.d. When conducting, corresponding to logical 0, the current is diverted from the I.e.d. into transistor. More current is drawn with I.e.d. off than when it is on (above, right).

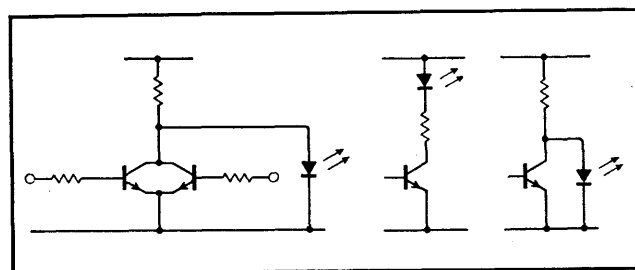
Circuit - 4 d.t.l. logic 0 indication

The load has still to be taken between output and supply positive. As with t.t.l. resistive limiting is required. This

configuration has advantage over previous that power consumption is much reduced when I.e.d. is off corresponding to logical 1 at output (below, left).

**Circuit - 5 driving with m.o.s.**

Complementary m.o.s. inverters have one device at a time conducting. As these devices have comparable but complementary characteristics the load may equally be placed between output and the + or 0 lines. Selecting suitable output stages, the inherent current limiting of the m.o.s. transistors may be sufficient to allow direct drive without the resistor. Care must be taken at the much higher voltages that may be used with c.m.o.s. (above, right).

**Circuit - 6 r.t.l. and open collector t.t.l.**

RTL output stage has transistor with collector load. With transistor off the current flows in the I.e.d., operated at the nominal supply voltage for r.t.l. the resulting I.e.d. current is low. Open-collector t.t.l. stages may be treated similarly to d.t.l. outputs.

Summary

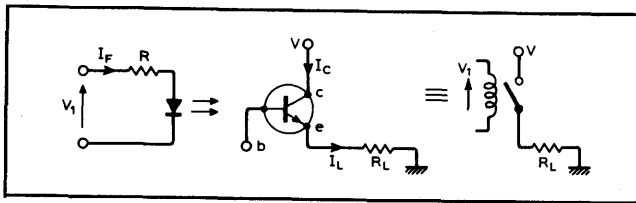
In all cases a minimum current of about 5mA is likely to be needed for good visibility. Typically around 10mA is used and the I.e.d. p.d. is 1.5 to 2.2V. This p.d. exhibits little change for a given device and subtracted from the supply voltage indicates the required series resistance $R = (V_s - V_{I.e.d.}) / I_{I.e.d.}$.

Alternative I.e.ds: HP 5082-4440
HP 5082-4444
Monsanto MV5094
RS Components LED2

Cross references

Series 9, cards 5, 6 & 11.

Switching with an opto-isolator



Typical performance

Opto-isolator: TIL112

For $I_F = 40\text{mA}$ and $C_{CEon} < 300\text{mV}$;

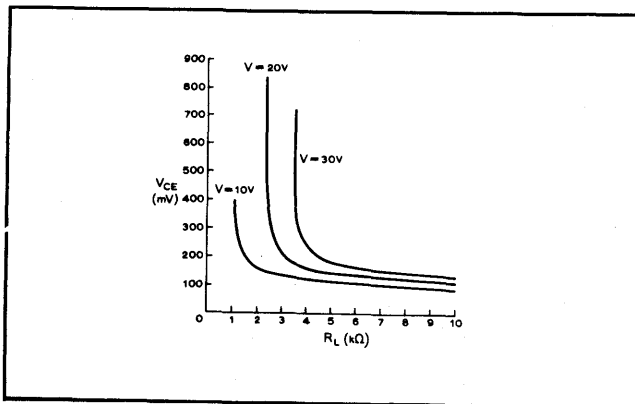
$R_L > 1.25\text{k}\Omega$ for $V = 10\text{V}$

$R_L > 2.5\text{k}\Omega$ for $V = 20\text{V}$

$R_L > 4\text{k}\Omega$ for $V = 30\text{V}$
(see graph)

$r_{off} = V_{CE}/I_C$ with $I_F = 0$
 $> 5\text{M}\Omega$

$r_{on} = V_{CE}/I_C$ with $I_F \neq 0$
(see graph at bottom)



Circuit description

Opto-isolators are used where electrical isolation between input and output is essential. They find use in medical electronics, applications where noise problems are met, and as a replacement for relays and pulse transformers. Circuit shows an opto-isolator in a single-pole, single-throw switch configuration. When used as a switch the transistor must operate in the saturated region of its characteristics (card 8, this series) when I_F is applied. This is ensured if V/R_L is less than the current at the knee of the characteristic for the particular I_F used. Graph shows that the greater the value of R_L , the smaller is the value of the voltage drop over the transistor. R_L , however, can not be increased indefinitely—

if it becomes comparable to r_{off} then voltage will appear over R_L when I_F is zero. The characteristics show that the volt drop over the transistor is reduced the larger the value of I_F . For example, a 50% reduction in V_{CE} was observed when I_F was increased from 15 to 60mA, V and R_L being fixed at 20V and 10kΩ respectively. In addition, for a given V , increasing I_F allows lower values of R_L to be used. For this device V_1 must exceed 1.2V for the photodiode to conduct. Information on possible switching rates is on card 9, this series, and may be increased, see ref. 1. Alternative components: OP1032, OP1062, MCT26, ISO-LIT12.

Further applications

A single-pole, double-throw switch can be constructed as shown above. When V_3 is less than 0.6V, Tr_1 is off and current passes through D_3 and D_1 (a general-purpose diode). Phototransistor PT_1 then conducts and V_1 is applied to the load, R_L . When Tr_1 conducts, however, the voltage across Tr_1 and D_2 falls below that necessary to make D_1 and D_3 conduct because the voltage across Tr_1 when conducting is less than that necessary to make D_1 conduct. PT_1 is then switched off, PT_2 is switched on and V_2 is applied to R_L . As shown above, V_1 and V_2 must both be positive. If either is negative then reversing the appropriate collector and emitter connections is necessary.

With R_1 of 10kΩ, R_2 100Ω and R_L 10kΩ, minimum V_3 is 4V and maximum V_4 is 4V. Range for V_1 and V_2 : up to $\pm 30\text{V}$. Use of an opto-isolator with a photo-darlington output stage increases current to the load.

Opto-isolators may be used in multiplexing circuits as shown above (ref. 2).

G_1 and G_2 are gating signals which allow V_1 and V_2 to pass to the op-amp in sequence. Resistors R_1 and R_2 may be different if V_1 and V_2 have to be weighted as well as multiplexed.

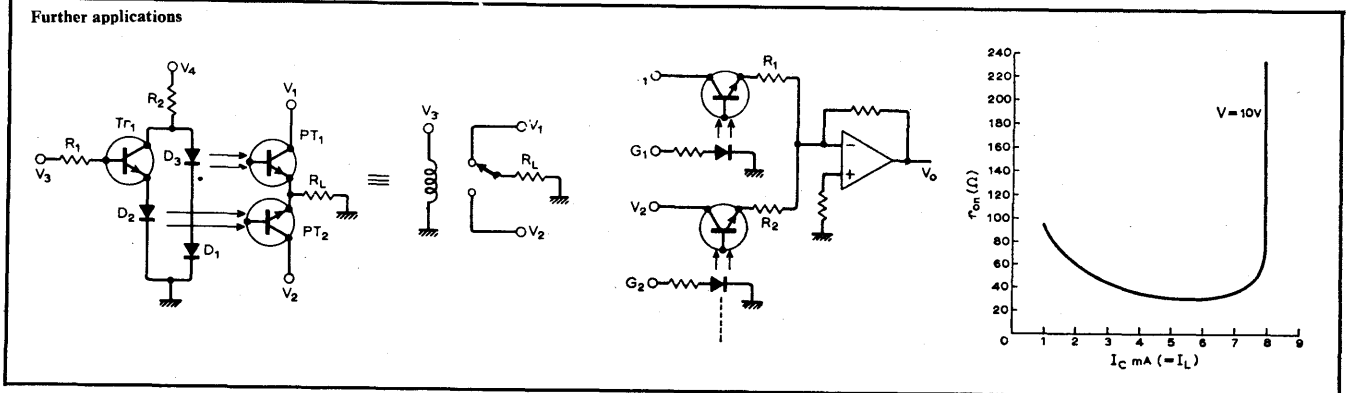
References

1. Gligler, D. F., Increase phototransistor bandwidth, *Electronic Design*, vol. 21, April 1973.
2. Das, S., Multiplexing analogue signals with optically coupled isolators. *Int. J. Electronics*, vol. 34, April 1973.
3. Bottini, M., *EDN/EEE*, 15 April, 1972.

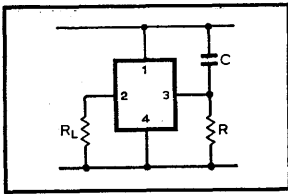
Cross references

Series 9, cards 4, 7, 8, 9 & 12.

Further applications



Integrated-circuit optoelectronic switch



Typical performance

High sensitivity:

R: 680k Ω , R_L: 1k Ω

C: 220nF

Supply: 15V

Ambient light excluded

Output switches for 150 μ A in i.e.d. (TIL209)

at distance of 2.5cm with optimum alignment. With i.e.d. and IPL15 in contact (negligible leakage) switching initiated by $10\mu\text{A}$ in i.e.d.

Low sensitivity:

R: 100k Ω

C: 3.3nF

Switching at room ambient levels.

Load voltage on-

condition: 12V

Load voltage off-

condition: $\approx 0\text{V}$

waveform is fed to an internal circuit with a threshold set by the light intensity, while a small amount of hysteresis prevents the output switching back and fore on small variations in that intensity. The output voltage swing is equal to the supply voltage on open circuit, but can supply enough current to operate a need relay or drive a transistor operating a higher power relay.

By setting a large RC time constant the sensitivity can be increased to the level at which the light from a match can be detected at a distance of several feet. Output is compatible with t.t.l. circuits as well as c.m.o.s. Spectral response of the circuit is such that it responds to the output of filament lamps as well as the low-cost light-emitting diodes. Speed of response is less than that of the frequency of oscillation, but at moderate sensitivities, switching rates of up to 1kHz are possible, making the circuit suitable for counting objects in most industrial systems.

Component changes

R 47k Ω to 1.5M Ω

C 2nF to 2 μ F

R_L 330 Ω to ∞

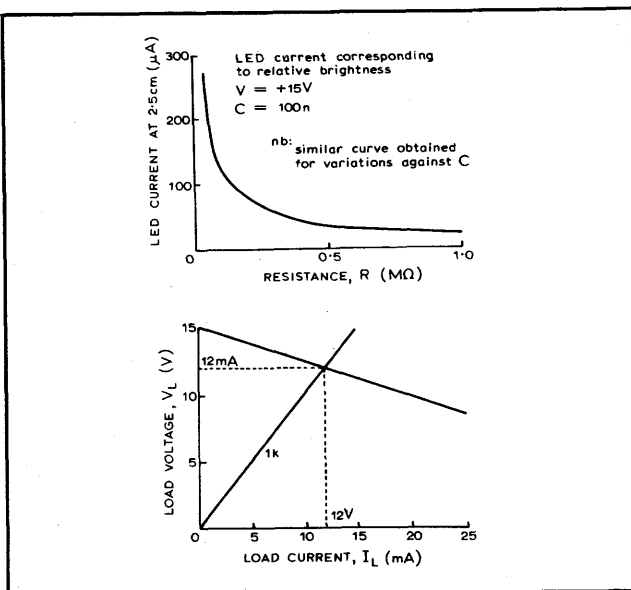
I_L up to 20mA

Supply 12 to 18V (some units operate down to 9V)

Time constant determines sensitivity of the circuit to light intensity.

Circuit modifications

- Output current is sufficient to drive a reed relay directly or digital circuits such as t.t.l. Addition of an emitter follower allows higher-current relays to be driven (left-most circuit).
- Internal functioning of the i.c. requires the generation of a ramp waveform whose frequency determines the sensitivity of light detection. This waveform is available at pin 3 and the ramp may be linearized by an external current generator.
- If the output is used to drive a light-emitting diode or filament lamp via a transistor then the optical equivalent of a Schmitt trigger results. The i.e.d. is switched to a fixed brightness at some pre-selected level of light input to the i.c.



Circuit description

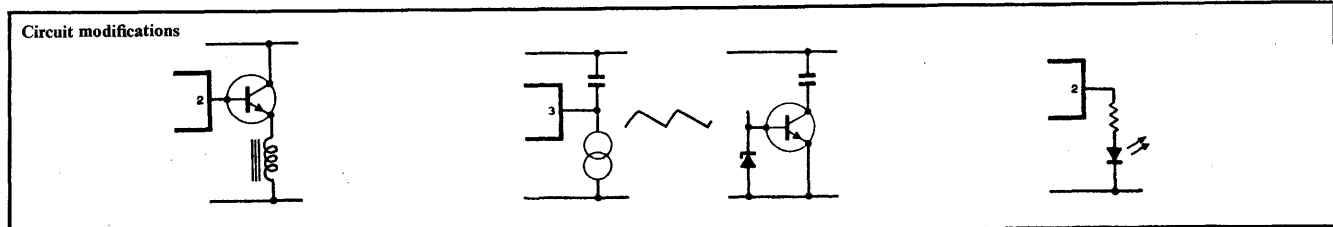
Circuit is that of a photo-sensitive trigger in which the voltage at the output (pin 2) changes by a large amount when the light intensity on the device exceeds a threshold level. This level may be varied over a very wide range by choice of R and C. These control the frequency of oscillation of an internal oscillator, in which the current in R controls the fall-time and an internal switch the rise-time. The resulting

Further reading

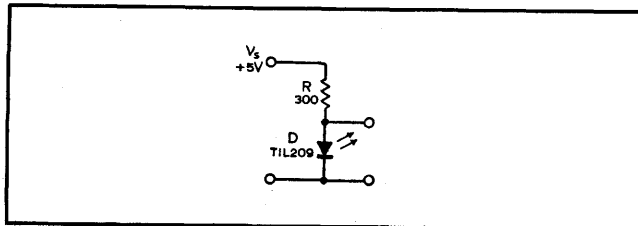
Noble, P. J. W., Cost advantages of integrated optoelectronic sensors, *Electronic Components*, 28 July 1972, pp.724-6.

Cross references

Series 9, cards 3 & 8.



Characteristics and applications of I.e.ds

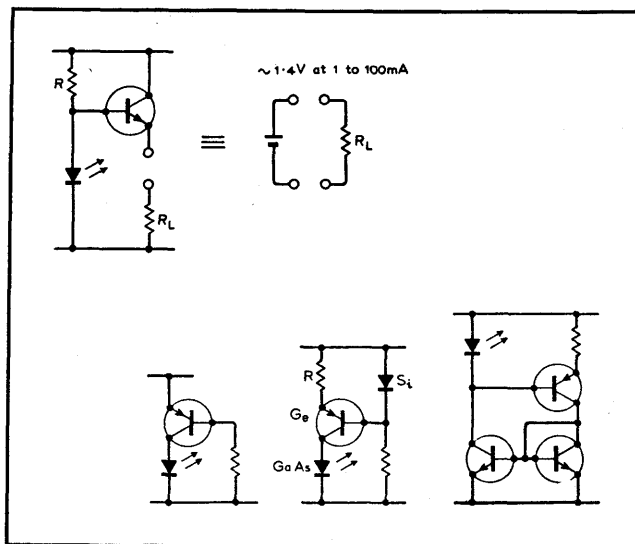


Circuit description

The device is a semiconductor p-n junction; gallium arsenide or gallium arsenide-phosphide being the most commonly used semiconductors for I.e.ds radiating at the red end of the spectrum. Other diodes are available with, for example, green emissions but low-cost units are presently restricted to red. Current levels are typically 5 to 25mA through the forward-biased junction and the terminal p.d. is typically 1.5 to 2.2V. These levels are compatible with the outputs of operational amplifiers, comparators, t.t.l. circuits and some m.o.s. circuits. The $v-i$ characteristic is identical in kind to that of the silicon p-n junction with equal increments in p.d. for a given multiple, in current i.e. $V \propto \log_e I$. At currents above a few milliamperes, the bulk resistance of the material dominates and the p.d. rises linearly with current. The p.d. falls with rising temperature having a comparable coefficient to silicon p-n junctions ($\approx -2\text{mV/K}$) and the diode may be used as a simple voltage reference at the same time as providing illumination.

The supply voltage should be large enough to allow the series resistor to be the dominant element in fixing the diode current. The diodes may also be operated in a pulsed mode with higher peak currents and low duty cycles. This mode is

common for diodes forming part of an alpha-numeric display where a single decoding part of an alpha-numeric display drives a number of diode arrays in succession at a rate fast enough to avoid flicker effects, i.e. allowing the eye to respond to mean levels of brightness.

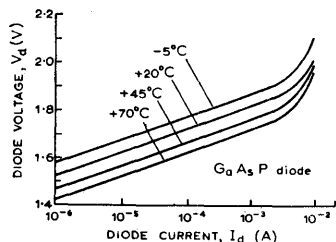
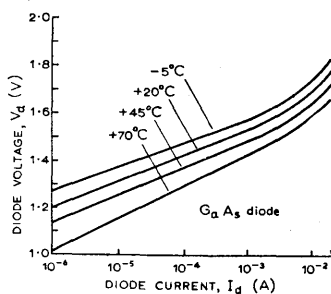


Circuit modifications

- Because of their comparable drift with temperature, the I.e.d. and a silicon transistor may be used as a simple voltage regulator in which the transistor acts as an emitter follower. For a GaAsP I.e.d., the output voltage is around 1.4V if the I.e.d. current is $\approx 10\text{mA}$ and the transistor operated well below its maximum current rating. With a suitable transistor (BFY50, BFR41, 2N3053) the circuit makes a good replacement for a single dry cell while operating from an unregulated supply of say 3 to 5V at currents of up to 100mA or so. At higher currents substitute 3055 transistor to minimize V_{be} drop.

- To operate a I.e.d. from the lowest possible supply voltage, a transistor may be used to drive the I.e.d. While the base current of the transistor and hence its collector current falls with falling supply, the I.e.d. current may be sustained at a reasonable level or supplied to within a few tens of millivolts of the I.e.d. terminal p.d. Any low-voltage constant-current circuit may be substituted including those using Si/Ge transistor/diode combination or those based on current-mirrors (see Circards series 6).

Circuit modifications



Further reading

Yen, T. T., LED doubles as sensor, *Electronics*, 4 Dec. 1972, p.113.

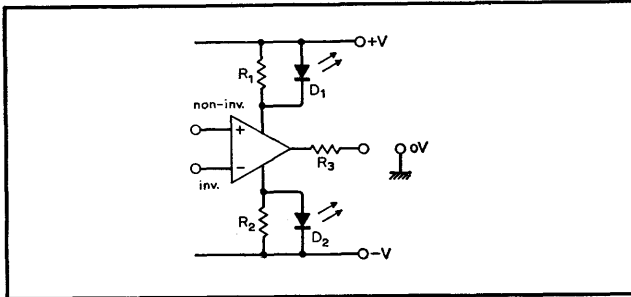
Texas Instruments, Measuring I.e.d. output, *Electronic Equipment News*, March 1971, pp. 24-8.

LEDs have advantage as constant-brightness sources, in 400 Ideas for Design, Hayden, vol. 2 1971, pp.277/8.

Cross references

Series 9, cards 1, 2 & 11.

Op-amp/comparator driving of l.e.ds



Circuit description

If two light-emitting diodes are connected in the supply lines of an op-amp having low quiescent current then resistors may be placed in parallel with the l.e.ds to ensure that the p.d. across them in the quiescent state is insufficient to produce illumination. Provided the inputs are maintained at a common-mode potential (c.m.) within the guaranteed operating range for the particular amplifier, then a small difference-mode signal at the inputs controls the illumination of the l.e.ds. The amplifiers used have internal current limiting that prevents the l.e.d. current exceeding a safe value. For low current l.e.ds, or where the amplifier has no current limiting the output may have a series limiting resistor added. Which lamps are lit will also depend on whether the output is taken to the zero, $+V_s$ or $-V_s$ lines. Leaving the output free of direct loading by the l.e.ds allows all the normal feedback circuits to be used to produce Schmitt, astable and monostable characteristics.

Where the status of the input need only be indicated by a single on/off l.e.d. the other may be removed and replaced by a shortcircuit. If negative feedback is applied then for small differences at the inputs neither l.e.d. will be significantly illuminated.

Lamp control by op-amp difference mode

Supplies: $\pm 15V$, CM: $\pm 12V$

DM	R_3	D_1	D_2	Notes
+	0	on	off	L.e.d. current limited by op-amp s/c current limit Current may be reduced by increasing R_3 D_1, D_2 may glow for $R_1, R_2 > 150\Omega$
-	0	off	on	
+	$-V_s$	on	off	
-	$-V_s$	off	off	
+	$+V_s$	off	off	
-	$+V_s$	off	on	

Difference mode voltage i.e. V /non inv. $-V$ inv.	Potential to which free end of R_3 is connected	D_1 reduces positive voltage applied to op-amp	D_2 reduces negative voltage applied to op-amp	IC: 741 $R_1: 220\Omega$ $R_2: 220\Omega$ $R_3: 220\Omega$ $D_1, D_2: TIL209$
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For CM voltages within 1.5V from most negative potential on l.c. output latches into positive state with same result as for DM+ regardless of relative voltages applied to inverting and non-inverting inputs. Some op-amps (LM307, etc.) will follow above table for CM voltages up to and including most positive voltage on op-amp.

Component changes

- The op-amp may be any general-purpose type such as 741, 748, 301, 307. Used as a comparator there is no feedback and presence or absence of compensation capacitors affects only speed of response.
- In circuit over and circuit above, high-speed comparators such as 710, 711, 311, etc., may be used where rapid pulse response is required. If no current limiting is available internally, protect amplifier/l.e.d. by series resistor (100 to 500 Ω). Some comparators only have single polarity outputs allowing on/off applications only.
- Diodes D_1, D_2 may be gallium arsenide, gallium arsenide-phosphide l.e.ds with p.d.s of 1.5 to 3V and currents of 530mA. TIL209, HP 5082-4484, Monsanto MV5094, RS LED2.
- Resistors R_1, R_2 chosen to prevent illumination by quiescent currents. Typically 150 Ω to 1k Ω .

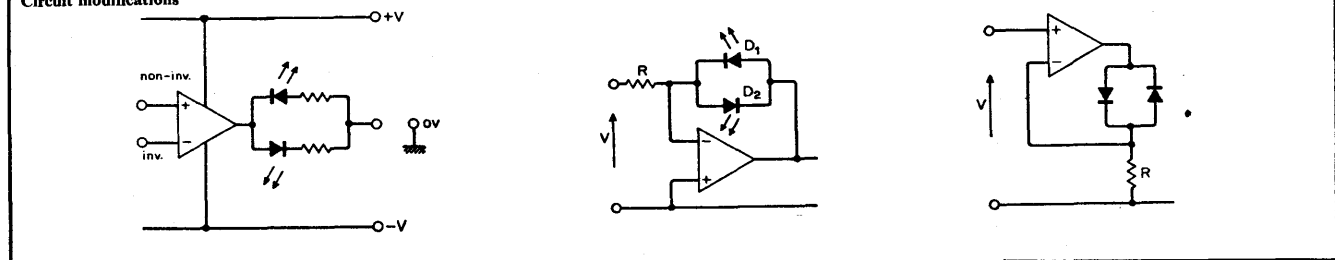
Circuit modifications

- The l.e.d. may be connected directly to the output of the op-amp and again by taking the l.e.d(s) to the zero-volts line changeover of lamp illumination takes place for small changes in the differential input. This can be applied to a red/green pair of lamps for example. Again limiting resistors may be used. Single-ended supplies with one l.e.d. to either supply line gives simple on/off indication. Parallel series resistors prevent illumination in the nominally off state (left).
- As well as monitoring op-amp output states when used as comparator, the l.e.ds may be placed in the feedback path. The input resistance of the circuit is R and the current in the l.e.d. is V/R with D_1 lit for V negative and D_2 for V positive. Emission from the conducting l.e.d. is a nearly linear function of the input voltage (above, left).
- Feedback may be series-applied at the input giving a similar output in the l.e.ds but with a very high input impedance (above, right).

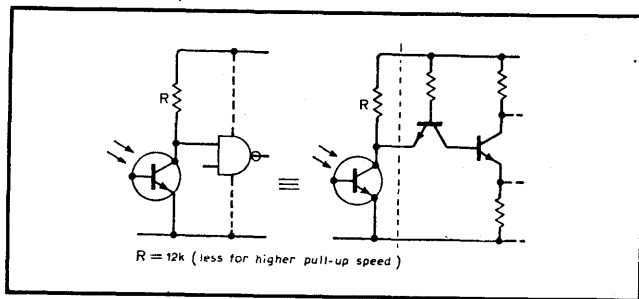
Cross references

Series 9, cards 2 & 11.

Circuit modifications

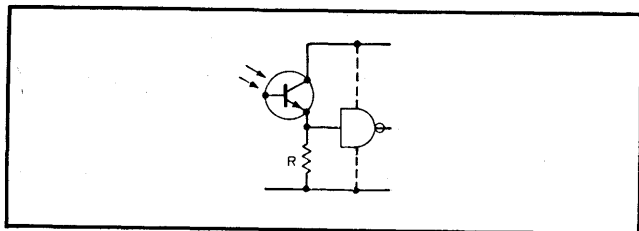


Phototransistor logic circuit drivers



Circuit description

This circuit is applicable in principle to most forms of digital circuit including t.t.l., d.t.l., r.t.l. and c.m.o.s. logic. The value of R may differ considerably, and that shown is suitable for t.t.l. and d.t.l. With c.m.o.s. the value can be increased for higher sensitivity with rise and fall times that become limited by the correspondingly increased input circuit time constants. Circuits of the r.t.l. type require lower values of R (say 2.7 to $4.7k\Omega$) to ensure that the input is effectively logical '1' with the phototransistor non-conducting. In each case the phototransistor must provide a current flow sufficient to bring the input down to logical '0' when illuminated. This demands a higher light intensity for r.t.l. with the values given.

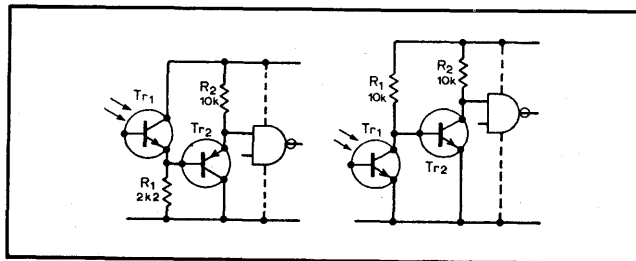


Circuit description

For r.t.l. and c.m.o.s. gates the value of R may be the same as that used for the previous circuits, i.e. 2.7 to $4.7k\Omega$ for r.t.l. and from $10k\Omega$ upwards for c.m.o.s. To achieve a time logical '0' at the input with t.t.l., d.t.l. either the resistance to ground must be low, say $< 300\Omega$, or an additional negative bias time be available. The choice of R is a compromise between worsening noise level margins when $R \rightarrow 300\Omega$ and requiring excessive current from the phototransistor as $R \rightarrow 0$. With a negative rail the quiescent voltage at the gate may be zero, while the on-current required is minimized.

Circuit description

Conventional transistor amplifiers may be interposed between the phototransistor and the logic circuit to overcome the problems associated with t.t.l. driving. The first method shown leaves a minimum logical '0' input of one V_{be} , worsening the noise margin. The second circuit interchanges the positions of Tr_1 and R_1 while adding a logic inversion by using Tr_2 in common emitter mode. This retains overall function required (phototransistor illuminated gives logical '1' at input) and retains good noise margin as Tr_2 can saturate almost to ground level.



Circuit description

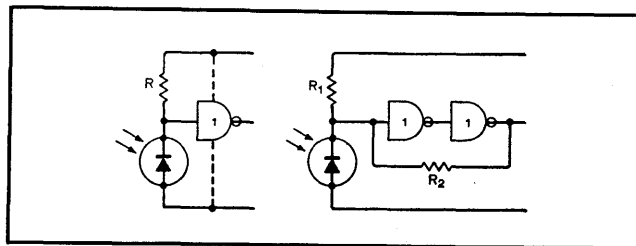
The input impedance of c.m.o.s. circuits is so high that photodiodes can readily produce enough current to give a high logic-level swing by choosing R large enough; e.g. $R = 1M\Omega$ requires a photocurrent of $< 10\mu A$ in diode for V_s of $10V$. Diode speed can be high and R should be minimized to avoid excessive rise times. Fall time is dominated by diode current in conduction. Complementary symmetry allows the interchange of diode and R and the use of inverting or non-inverting gates as well as nand/nor elements. To avoid output jitter, positive feedback is applied overall (two inverting or one non-inverting buffer). Hysteresis is controlled by the ratio $R_1:R_2$. The output transition is also speeded up and allows operation of succeeding counters from slowly-changing light intensities.

Further reading

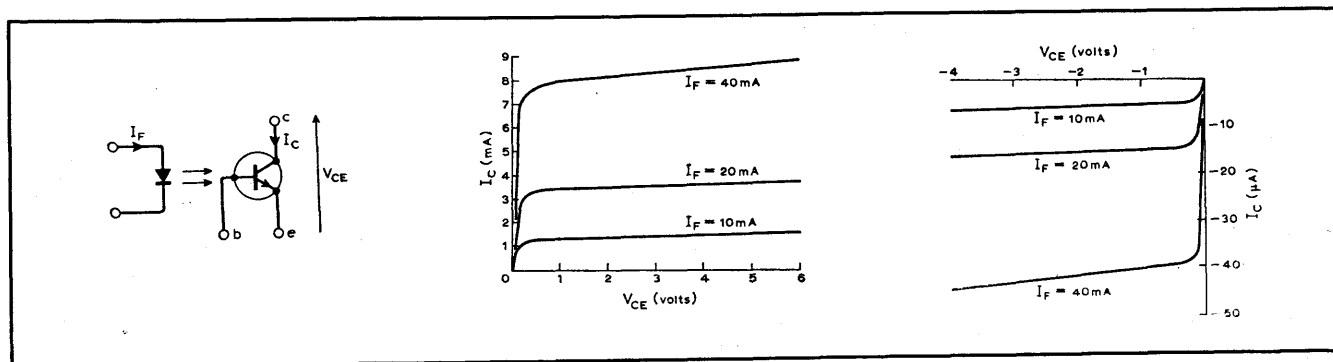
Korn, S. R., Photo Couplers, General Electric application note 200.62.

Cross references

Series 9, cards 3, 8, 9 & 11.



Optically-coupled isolator: static characteristics



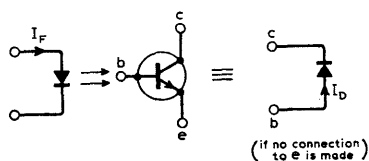
Component description using TIL112

Optically-coupled isolators are generally used for their isolation properties, replacing switches, pulse transformers, etc. Their output side may be simply a photodiode which is fairly fast, a phototransistor which is slower but has a higher output current (because of the higher output current they may be faster than a photodiode if feeding a highly capacitive load) or a photo-darlington circuit. The component under test is a silicon n-p-n phototransistor activated by a gallium

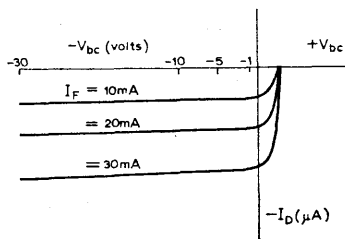
arsenide i.e.d. The characteristics given are not claimed to be typical in a quantitative sense though they will be qualitatively. The graphs point out some aspects of the characteristics not quoted by the manufacturer. The photodiode requires approximately 1.2V in the forward direction and the characteristics are identical in form to that for a normal transistor, except that the drive signal is I_F rather than I_b . As access to the base is available, the transistor can be controlled by base current drive or by optical means (diode current drive).

Alternative components ISO-LIT12, MCT26, OP1032, OP1062.

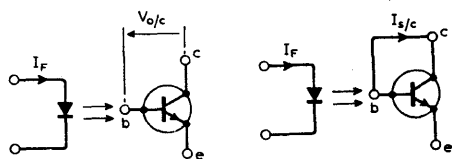
Photodiode operation



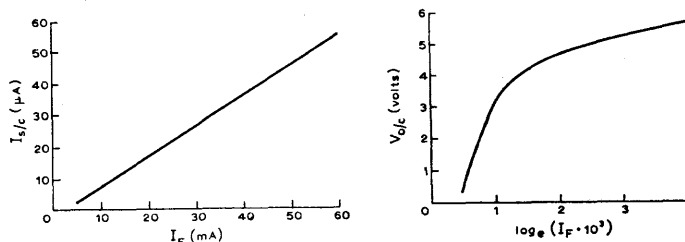
Photodiode characteristics



Photovoltaic operation



Photovoltaic characteristics



$V_{O/c}$ measured with instrument having $1M\Omega$ input resistance
 $I_{S/c}$ measured with instrument having terminal volt drop of less than 12mV.

Graph right shows that, over a wide range of I_F (corresponding to $I_F > 10mA$), $V_{O/c} \propto \log I_F$
 Graph left shows $I_{S/c} \propto I_F$.

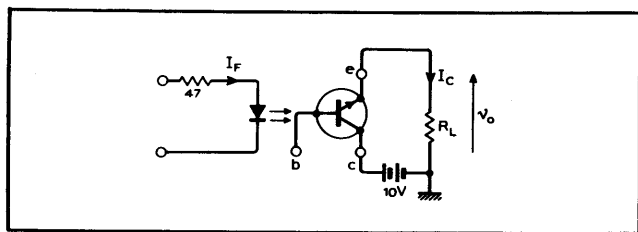
Further reading

Electronic Design, 1971, vol. 19, no. 11, pp. 44-55, and no. 12, pp. 46-52.

Cross references

Series 9, cards 3, 4, 7 & 10.

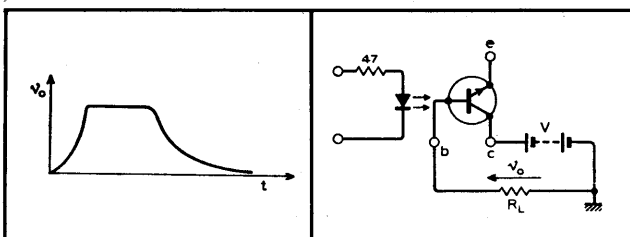
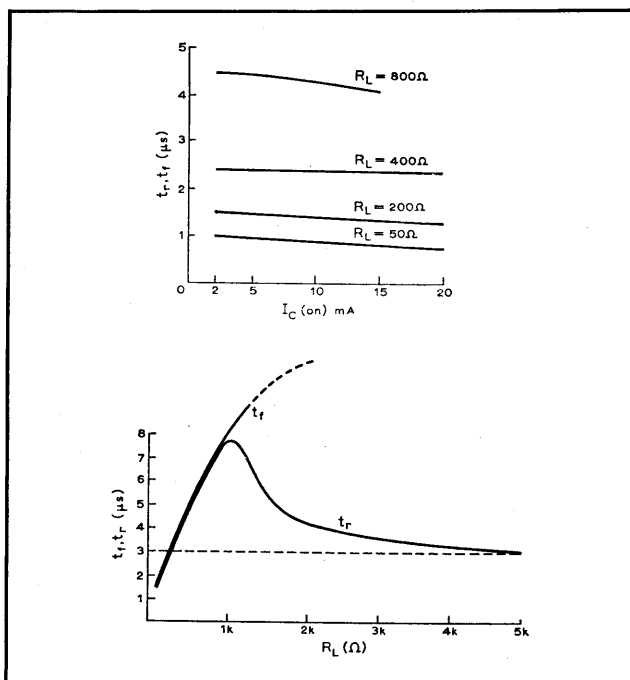
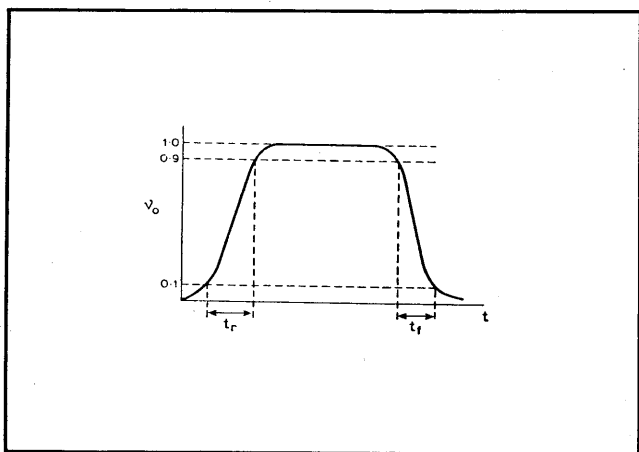
Optically-coupled isolator: pulse characteristics



Measurement data

Phototransistor TIL112.
 V_o monitored on c.r.o.
 with $1M\Omega$, $28pF$ input
 impedance.

I_F generated from pulse
 source having $10ns$ rise
 and fall times when feed-
 ing 50Ω load.



Experiment description

Graph top right was obtained by varying I_F to obtain particular values of I_C , and by deduction V_{CE} , and then measuring t_r and t_f . Values of t_f were identical, within experimental error, to the values for t_r . Clearly t_r and t_f are proportional, though not linearly, to R_L and are only slightly affected by I_C . The transistor operating point in all of these graphs is outside the saturated region of the static characteristics (card 8, this series).

Graph at bottom right was obtained by keeping I_F constant at a value such that I_C when supplying a short circuit was $10mA$. In this case R_L was varied sufficiently to obtain readings when transistor was saturated. Graph above shows a typical pulse shape for low R_L and graph right for high R_L , giving saturated operation. Operation in the saturated region, as is done when using the transistor as a switch (card 3, this series), alters the pulse shape and causes the graphs of t_r and t_f to diverge.

Alternative components

ISO LIT12, MCT26, OP1032, OP1062.

Photodiode operation, shown in above, provides smaller rise times ($< 1\mu s$) but generally smaller current capability since the load current is the transistor base current which is light-controlled. The maximum value of V is $30V$ and R_L can be increased until V_o is approximately equal to V . When feeding a capacitive load or a large-value resistor, corresponding to large output voltage swing, the rise time will increase.

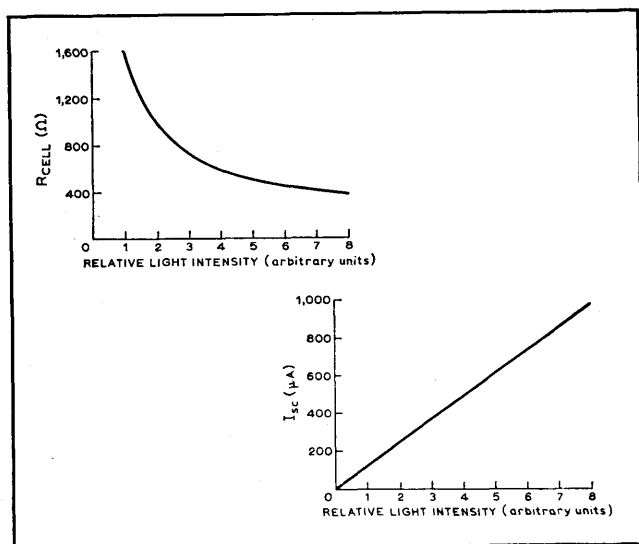
Further reading

Geiger, D. F., Increase phototransistor bandwidth, *Electronic Design*, no. 8, vol 21, April 1973.

Cross references

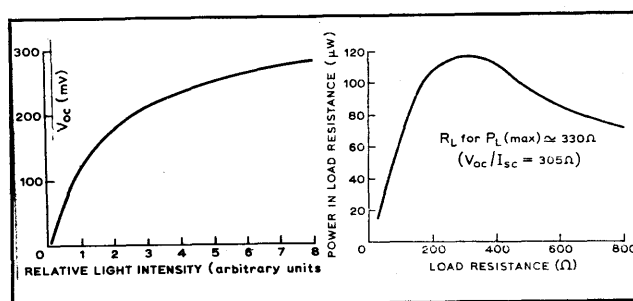
Series 9, cards 3, 7 & 12.

Photoconductive and photovoltaic cells



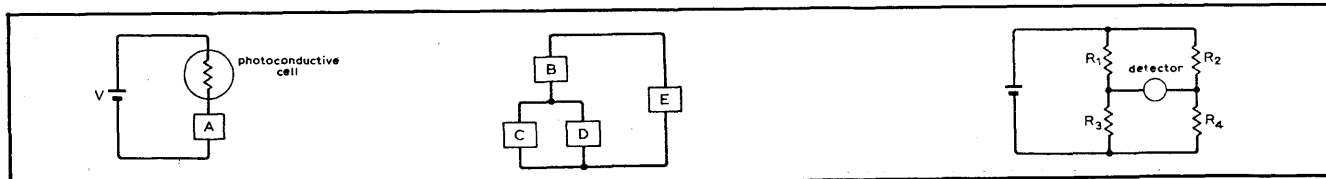
Typical characteristics

A photoconductive cell of the cadmium sulphide type behaves as an ohmic impedance which depends on incident illumination. Graph on left shows the variation of cell resistance as a function of light intensity to be $R_{cell} \propto 1/(\text{light intensity})^n$ where n is typically in the range 0.5 to 1.0. When connected in a circuit fed from an external source the resulting current can be controlled or modulated by the light intensity falling on the photocell. Such cells do not respond instantaneously to changes in light intensity due to their inherent capacitance and may therefore be considered to have a time constant which depends on the existing and previous illumination. Cell rise



time normally exceeds its fall time and both responses are slower at low light levels. Response time to pulsed light-changes falls with increase in light intensity and decrease in ambient temperature.

Although silicon and selenium photovoltaic cells are fabricated by different processes, their electrical characteristics are very similar. Both types have a p-n semiconductor junction, the photo-sensitive layer being p-type in the silicon cell and n-type with selenium. If a photon of the appropriate wavelength strikes a valence electron, the latter may receive sufficient energy to convert it into a conduction electron. Because each photon releases a single photoelectron the short-circuit current is proportional to light intensity, see second graph and to cell area. Open-circuit voltage of a photo-voltaic cell is a logarithmic function of light intensity—third graph. Hence the terminal voltage is proportional to the log of the photo-generated current. The cell is therefore a non-linear device which can be represented by a current generator in parallel with a real diode possessing bulk resistance leakage resistance and capacitance. These cells are used for energy conversion but the load resistance for maximum power transfer is not given by $R_{L,opt} = V_x/I_{sc}$ as predicted by linear circuit theory. However in practice, maximum load power is obtained with a resistance close to that predicted above.



Applications

Many photoconductive cell applications employ the simple series circuit shown left. Often the p.d. across the component A can be usefully employed for a control function that is dependent on the light intensity falling on the photocell in which case A would normally be a resistor. If A is the coil of a relay its current may be held below the operate value until the photocell resistance falls with increase in illumination. A contact of the relay could be used to open the lamp circuit thus providing a repetitive flashing signal. If A is a microammeter it can be calibrated to serve as a light intensity or exposure meter. If a lamp which illuminates the photocell when it is on, the circuit can be made into a "lock-on" type by illuminating the photocell from another source which is then removed. If A is a resistor and a suitably-engraved disc is rotated to modulate the light falling on the photocell, the waveform available across the resistor can be made sinusoidal, complex or noise-like.

A second common arrangement is shown, middle, where E is normally a power source. With B as a resistor, C a relay coil and D a photoconductive cell, the latter shunts current from the relay until the light source is removed when the relay operates due to the increase in photocell resistance. If B is

the lamp illuminating the photocell then lamp failure causes the relay and its associated equipment to be isolated from the supply. If E is the amplifier input circuit, B a coupling capacitor, C a photovoltaic cell and D a resistor, the p.d. across the latter will follow the signal variations due to illumination of the photocell through, for example, a film sound track.

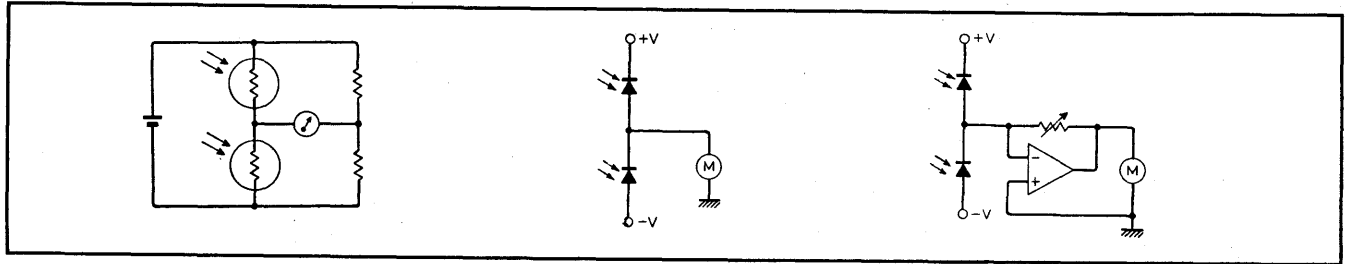
The bridge shown right is also widely used and can be driven from a photovoltaic cell or bank of cells for resistance measurement. If R_3 is replaced by a photovoltaic cell used as a photoconductive element that "backs-off" the supply voltage the circuit can be used as a sensitive exposure meter. If R_1 and R_2 are both photoconductive cells driven from the outputs of the left and right channels of a stereo amplifier via lamps the detector becomes a balance indicator.

Photovoltaic cells can be used to charge batteries, operate low-power electronic equipment, charge capacitors to operate a photoflash and actuate relays directly from a light source to operate higher power equipment.

Cross references

Series 9, cards 8 & 11.

Light intensity measurement and detection



Light intensity measurement and detection

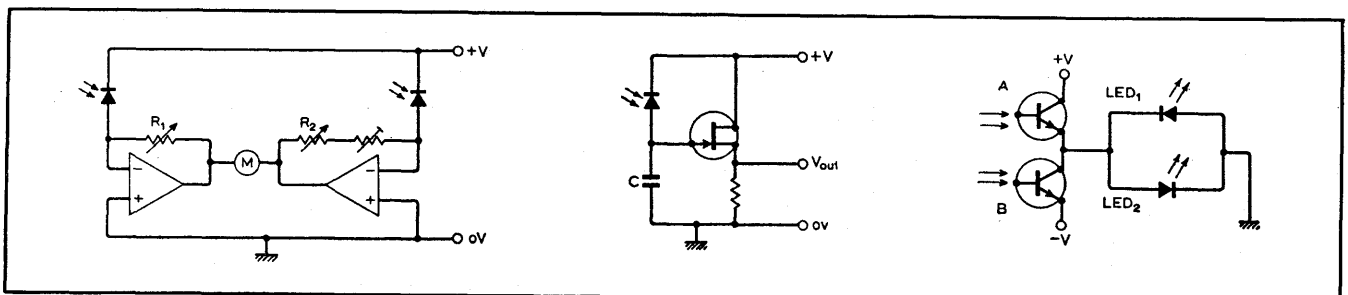
Circuit shows a normal Wheatstone bridge circuit where two resistors have been replaced by photoconductive cells of the cadmium sulphide type. If the two cells have accurately matched characteristics the bridge can be used to measure the relative light intensity from two different sources. When illuminated simultaneously from the source the unbalance current gives a measure of the degree of mismatch between the cells.

Photodiodes could be used in place of the photoconductive cells above in which case the unbalance current will be virtually independent of supply voltage and of the p.d. across the meter as the diodes are effectively constant-current sources; i.e. the circuit shown middle is applicable where the supply voltages are uncritical and meters requiring up to about 1 volt terminal p.d. could be used.

In either of these circuits the unbalance current could be fed

to the input of a feedback operational amplifier where the output voltage could be set to, say, 0-10 volt for a given difference in light intensities, circuit right being one such arrangement. The system is bipolar, i.e. either light intensity could be greater than the other with the meter reading proportional to their difference.

To monitor the ratio of two light intensities the circuit shown over can be used. Resistors in the feedback networks of the operational amplifiers can be scaled to accommodate the initial unbalance between the photodiode sensitivities and to set the desired ratio of light intensities. The observed unbalance voltage will be the ratio R_1/R_2 multiplied by the ratio of the sensitivities of the photodiodes. When the diodes feed into an operational amplifier it is equally feasible to use positive feedback to obtain a switching action at a given level of light intensity, i.e. to make the circuit a form of Schmitt trigger.



Another type of light intensity measurement is involved when determining the total light from a source over a period of time, i.e. an integrating procedure. As the photodiode is essentially a current source, it is not necessary to use a feedback integrator. As shown middle, the current from the photodiode may simply be fed to a capacitor so that the p.d. built up across the capacitor is to a first-order approximation measure of the integral of the light intensity over the exposure time. For example, this could be used in a camera to control the time for which the shutter is open; closing the shutter when a certain total amount of light has reached the indicator and hence the film. To avoid loading the capacitor, and disturbing the integrating function, it should be followed by a field-effect transistor, or similar high input impedance amplifier. This problem would not be so serious if the photodiode were used with an operational amplifier in a conventional integrator. By replacing the capacitor C with a resistor, the circuit could be applied to light intensity control.

The simple circuit shown on the right can be used when it is desired to detect the presence or absence of given light sources

rather than to determine their intensity. The pair of light-emitting diodes connected in parallel back-to-back are fed from two phototransistors that are separately energized by the light beams A and B. This is a simple logic circuit in which LED1 will emit for a negative output which occurs when phototransistor B is on and phototransistor A is off. LED2 emits for the opposite state of A on and B off. If neither phototransistor is illuminated both i.e.s are extinguished. Both i.e.ds could emit light if the phototransistors are not well matched and/or the light sources are of unequal intensity.

Further reading

National Semiconductor Linear Applications Manual AN8-8, AN31-3, AN31-18, 1972.

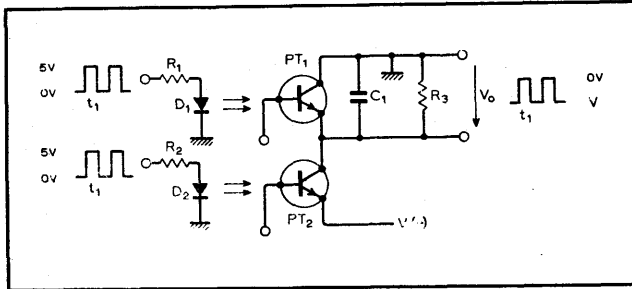
Graeme, J., Op-amp boosts phototransistor speed, *Electronic Design*, 2 March, 1972. p.62.

Vaisnys, A., LED-phototransistor coupler isolates analog signals, *Electronic Design*, 13 April, 1972. p.90.

Cross references

Series 9, cards 1, 2, 5, 6, 7 & 10.

Choppers and rectifiers



Typical performance

Opto-isolators TIL112
 R_1, R_2 : 100 Ω ; R_3 : 10k Ω
 C_1 : 10pF
 Drive frequency: 1kHz
 Drive pulse height 5V
 V : 10mV from 10- Ω source
 V_0 =10mV pk-pk,
 reading 5mV on r.m.s.-
 calibrated mean-rectified
 instrument.

With $V = 0$ volt d.c.
 measurement of V_0 is
 0.05mV and a.c.
 measurement is 0.13mV.
 With l.e.d. drive varying
 from 3 to 6V, meter
 change < 1%. With drive
 at 5V and C_1 varying
 from 10pF to 0.01 μ F,
 meter change < 2%.
 With drive at 5V and R_3
 varying from 2 to 33k Ω ,
 meter change < 4%.

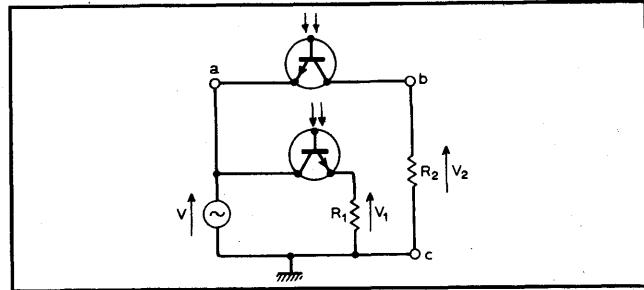
Circuit description

Circuit 1 shows a chopper circuit using a pair of opto-isolators with phototransistor output stages. Diodes D_1 and D_2 are light-emitting diodes driving the phototransistors PT_1 and PT_2 respectively. When one of the phototransistors is illuminated, its collector to emitter path becomes of low resistance (card 3) though with an offset voltage term inherent to the bipolar nature of the device. If the l.e.ds are driven in antiphase then only one phototransistor is on at a time and either the signal voltage is applied to the resistor R_3 through PT_2 with PT_1 non-conducting or PT_2 blocks the signal while PT_1 absorbs any current which leaks through PT_2 . This is a conventional series-shunt chopper but energized by optical rather than direct electrical pulses, thereby avoiding the breakthrough of spikes into the output circuit. Use of the shunt transistor PT_1 is not essential if the d.c. signal V is large but it does have the advantage that any stray or load capacitance represented by C_1 can be rapidly discharged during the off part of the cycle, i.e. the fall time will be comparable to the rise time (card 9). This in turn is important where the signal to be chopped is a rapidly varying one in which case the chopping frequency has to be as much above the signal frequency as possible.

The drive circuits for the l.e.ds can be conventional (card 6, series 8, card 2) but as they require to be antiphase they could be from the collectors of a standard two transistor astable or from the Q and \bar{Q} outputs of any t.t.l. counter circuits. If the chopped waveform is then fed to an a.c. amplifier the output can be rectified. Or better still rectified by a second pair of photo transistors synchronized with the drive pulses. The method is applicable to photo-f.e.t.s where the absence of a d.c. offset term allows chopping of much smaller signals. A third possibility is to use cadmium sulphide photoconductive cells in place of the phototransistors although the chopping frequency cannot then be as high. For positive values of V the collector-emitter connections of both phototransistors would have to be reversed.

Further reading

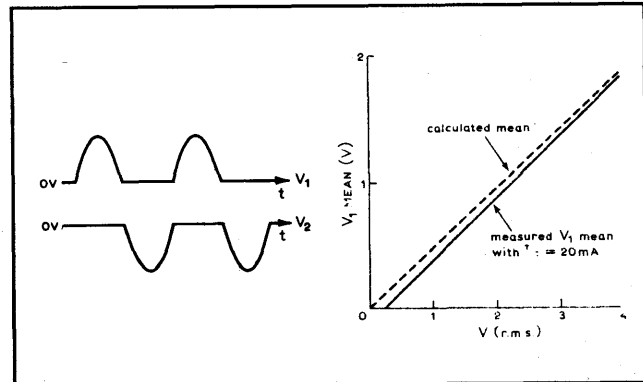
Pulse generator for diode emitters, W.W. 1972, p. 18.



Performance data

R_1, R_2 : 1.5k Ω
 V : 0 to 4V r.m.s.
 Signal frequency: 10kHz

Opto-isolators TIL112
 I_F (not shown) 20mA for
 both opto-isolators.
 Graph obtained with
 these values.



Circuit description

Circuit 2 shows a pair of opto-isolators being used as simple diodes to provide a half-wave rectified signal which could be fed to a mean-reading instrument to effectively measure the value of the signal voltage, V . For this, branch abc is not necessary—it simply shows that both positive and negative half cycles can easily be obtained. The advantage of this circuit is that the volt drop across the transistor when conducting (card 8) in the forward direction is much less than that across a normal semiconductor diode (0.6V). Hence the phototransistor can be used as a rectifier for low voltages. When operating in the forward direction the phototransistor is operating in its saturated region and this together with the maximum reverse voltage (4V in this case) determines the value of R_1, R_2 .

Circuit operation is not very sensitive to the value of I_F . Although the forward characteristic of the phototransistor as a diode is better than that of a diode its reverse characteristic is poorer. This can be greatly improved if I_F is made zero during that half of the cycle in which the phototransistor is nominally non-conducting. This could be done by driving I_F from a comparator whose input is the signal, V . All the other standard methods of improving diode performance could also be used, e.g. card 1, series 4.

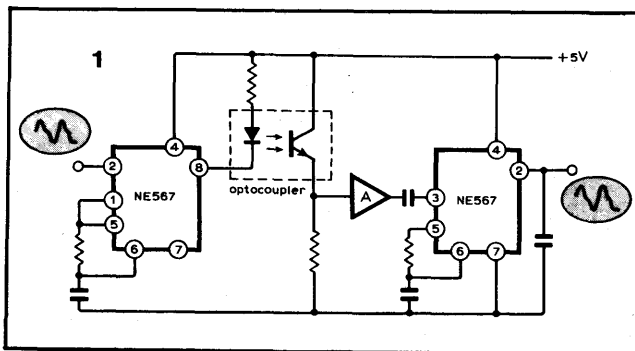
Cross references

Series 9, card 8
 Series 4, card 1.

Optoelectronics

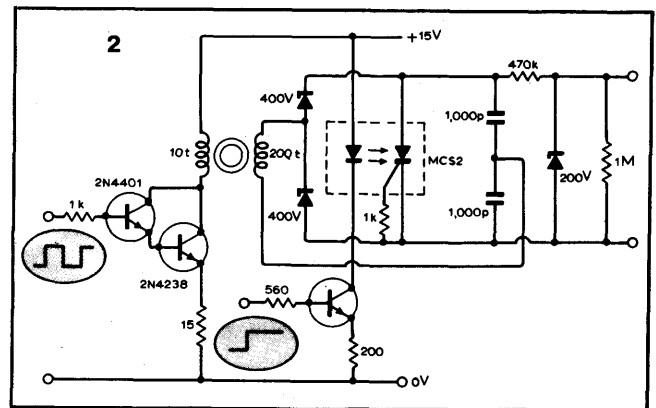
1. Transmission of analogue data via an optical link removes the problem of noise at the interface between two parts of a system due to earth loops, common supply lines etc. Alternatively it permits the two parts to be at quite different potentials. Non-linearities in the l.e.d. and phototransistor or in other transmitter/receivers distort the waveform, and a solution is to use a frequency-modulated carrier. In the example shown two phase-locked loop i.c.s are used, one as a frequency modulator, using only the v.c.o. section, while the other demodulates the signal after transmission via the optical link.

Light-coupled isolator uses phase locked loops, *Electronic Engineering*, vol. 46, May 1974, p.19.



2. Complete isolation between drive circuitry and transducer is particularly important in medical electronics – in some cases primarily for safety reasons, in others where ground loops create difficulties. A pulse circuit is shown in which an electrode is fed by a section of a circuit having no power supply. A toroid is fed with short duration pulses, around $30\mu\text{s}$, derived from a 555 timer (not shown). These are stepped up, rectified, filtered, regulated and, if required, attenuated. An inhibit pulse to the 555 can simultaneously operate the l.e.d. bringing the thyristor into conduction rapidly discharging the capacitors.

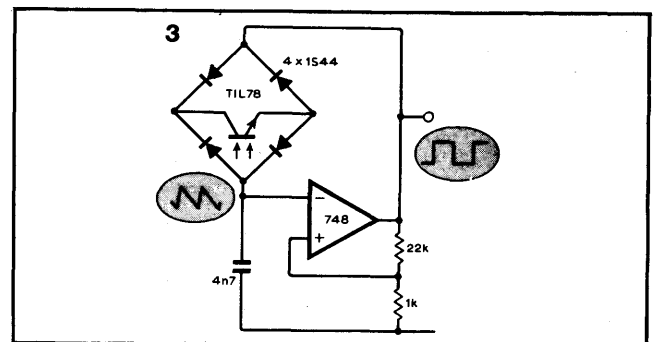
Sonsino, J. Toroid and photo-SCR prevent ground loops in high isolation biological pulser, *Electronics Design*, vol. 21, June 21, 1973, p.128.



3. The current in a correctly biased phototransistor is an almost linear function of the light intensity. If that current also flows in a capacitor then a linear ramp results. The circuit shown is a basic op-amp astable in which the bridge rectifier allows bidirectional current flow in the capacitor while restricting the phototransistor to unidirectional current.

With the given component values the reference article indicates a change in oscillator frequency from 50Hz to 50kHz for 0.02 to 20mW cm^{-2} of radiant energy from a tungsten lamp. Because of the almost constant current nature of the photo transistor aided by the small ramp size (small hysteresis) the capacitor waveform should be a rear-perfect triangular wave.

Miller, M. Introduction to Optoelectronics, ch. 17 in Semiconductor Circuit Design Vol. II, Texas Instruments, Ed. Norris, B.



Wireless World Circard

Series 10: Micropower circuits

Most of the micropower circuits described rely on the fact that the current gain of silicon planar transistors, discrete or integrated, holds up down to nanoampere levels (see card 12). Transistors operated in this region also show an increased input resistance and a low saturation voltage. Some, such as the amplifiers described on cards 1, 7 and 9 and the astable of card 2, are designed to operate at low voltages, around 1V, while others are designed for low-current operation, such as the crystal oscillator of card 6. One trick used in circuits involving amplifiers is to operate complementary class-B stages using the common-emitter rather than the more usual common-collector configuration, see for example, cards 2, 7, 8 and 9. Another trick is to design circuits to minimize "stand-by" consumption. This is often seen in Schmitt and monostable circuits, but an unusual application is in the optical link of card 4.

Interesting circuits include a polarity-reversing converter, useful where a second battery might otherwise be needed for biasing. Another "power" supply circuit is the low-voltage regulator of card 3 which stabilizes a supply of $2V \pm 20\%$ to $400mV \pm 0.5\%$ and regulates output to 0.5% from load variations of 20% . A modification given allows supply voltages down to 1V. The lower V_{be} of germanium transistors makes this circuit possible.

- Low-voltage a.c. amplifier 1
- Low-voltage astable circuit 2
- Low-voltage regulators 3
- Optical link with low standby power 4
- Signal-powered circuits 5
- Micropower crystal oscillator 6
- Micropower d.c. amplifier 7
- RC oscillator for low voltages 8
- Class B low-voltage amplifier 9
- Low-voltage d.c. converter 10
- Low-current use of diodes 11
- Micropower active devices 12

Micropower circuits

Think small! In tune with the broadening of the frequency spectrum, so has the range of powers grown at which electronic circuits may be coerced into functioning. Within modern integrated circuits it is common to find individual transistors operating at microampere currents, with device p.d.s of a volt or so. Discrete transistors can retain useful gain at currents several orders of magnitude less at room temperatures (even with silicon transistors leakage current imposes constraints on the usable current/temperature combinations). Since the leakage currents may be markedly reduced by controlling the doping levels and depth of penetration, by attention to surface impurities and by reduction of device area, it is difficult to define this lower limit. Collector currents of 1nA at 20°C and 1μA at 100°C are possible, though it would be foolhardy in the extreme to suggest that these could not be improved on — undoubtedly before publication some brilliant new process will appear capable of improving on these figures by a factor of ten or more!

The position is quite different in respect of minimum operating voltages. Fig. 1 shows the variation of I_c and I_b against V_{be} for a low-leakage planar silicon transistor operated at constant V_{ce} . No matter how low the current is reduced the fundamental relationship $I_c \propto \exp V_{be}$ ensures that the value of V_{be} changes at a much slower rate. For example, a ten-fold reduction in I_c corresponds to a reduction of approximately 60mV in V_{be} at room temperatures. Since the minimum value of I_c is likely to be fixed by load requirements etc. no amount of juggling can reduce the minimum supply voltage below that of the corresponding V_{be} , and in most circuits the supply voltage will have to be significantly higher. Excepting special cases such as certain complementary oscillators where only one of a complementary pair needs to conduct at a time, the minimum supply voltage will be greater than 1V and may have to be greater than 1.5V for op-amp type circuits.

In the above discussion no mention has been made of constraints imposed by transistor V_{ce} . At low currents the value of $V_{ce(sat)}$ is very much less than the V_{be} values above, though it must be observed that $V_{ce(sat)}$ often rises with temperature while V_{be} always falls (typically the temperature coefficient of V_{be} is -2mVK^{-1}). At high current densities the bulk resistance of the semiconductor comes to dominate the junction characteristics and the minimum p.d.s may well exceed a volt for both base and collector-emitter paths. Self-heating will cause the value of V_{be} to fall somewhat but the effect should be neglig-

ible for circuits coming under the micropower heading.

The use of germanium transistors to minimize voltage requirements is well

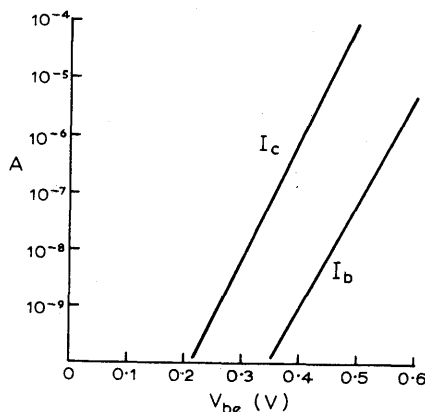


Fig. 1. Variation of I_c and I_b against V_{be} for a low leakage planar silicon transistor operated at constant V_{ce} .

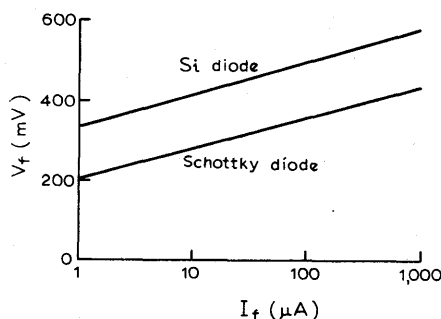


Fig. 2. The Schottky-barrier diode has a p.d. appreciably lower than that of a silicon diode.

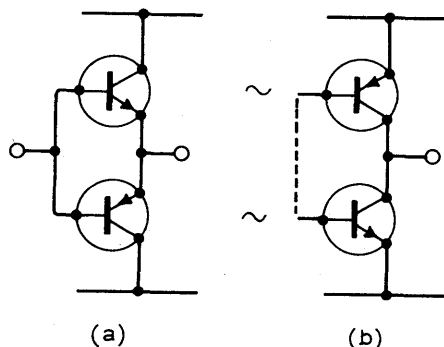


Fig. 3. Replacement of a complementary pair of emitter followers (a) by common emitter stages with emitters taken to opposite sides of the supply (b).

known, but the leakage currents are such that high temperatures are inconsistent with micropower operation. When using diodes, an intermediate region is provided by Schottky-barrier devices. As shown in Fig. 2, these have a p.d. appreciably less than that for a silicon diode, and can be used as bias elements as well as for rectification.

The interpretation of this term "micropower" has been made a generous one in this series because the techniques used in micropower circuits can be usefully applied in many other fields. For a given specification of load resistance and the required voltage or current swing, the minimum theoretical voltage can be determined assuming a voltage supply (if the circuit is to be supplied from a constant current then it is the minimum value of current that is defined). If this voltage is significantly greater than the minimum operating p.d.s of semiconductor devices as discussed above, then standard circuit configurations may well give satisfactory results. At lower load p.d.s new circuits are necessary to allow operation from correspondingly low supply voltages. As a rough guide, it is now possible to produce a.c. amplifiers, oscillators both RC and LC , voltage regulators, and astable and monostable circuits that will operate from supply voltages in the region of 1V, though with obvious limitations on output voltage swing and with reduced stability against supply/temperature variations. Other circuits such as power amplifiers and operational amplifiers may require somewhat higher voltages but nearly all functions can be provided while operating from a single dry cell. In designing these circuits a critical parameter is the minimization of wasted voltage, and this is equally applicable to conventional circuits where the load-swing is to approach the available supply voltage. A good example is the replacement of a complementary pair of emitter followers Fig. 3(a) by common emitter stages with emitters taken to opposite sides of the supply Fig. 3 (b). In the former case, since the bases can only be driven with difficulty to within hundreds of millivolts of the supply, then the output suffering an additional V_{be} loss may be up to 1V less than the supply at both extremes, i.e. the output has a peak-peak value up to 2V less than the supply voltage even at relatively low currents. In the circuit of Fig. 3 (b), though the drive conditions may be more difficult to meet, only the transistor $V_{ce(sat)}$ appears in series with the load, and outputs to within a couple of hundred millivolts of the supply are feasible. It is for the same reason that the

series-pass transistor is in the common-emitter mode in voltage regulators where minimum input-output differential is important.

Two other related parameters that may be important in micropower circuits are maximum efficiency and the minimum quiescent power. As discussed in the article Power Amplifiers (June issue), these two conditions are often associated, as, for example, in class B power amplifiers. For minimum distortion it may be necessary to increase the quiescent power. Where the amplifier normally operates close to maximum output then this contribution to the power consumption is negligible. Conversely if the amplifier operates at full output only for short periods, then the mean power is strongly dependent on the quiescent power. In summary, for continuous operation at maximum outputs, saturation voltages will be the limiting factor, while the quiescent power needs most design ingenuity for large peak/mean ratios in output.

A different problem arises when efficiency at maximum output is really critical. Then the need to saturate the output transistor(s) to minimize lost voltage would bring the corresponding disadvantage that the base current becomes a large fraction of the load current. The combination of high current gain together with low saturation voltage is not an easy one, though at least high breakdown voltages are not required of the device. Special transistors called "super- β " devices are now used as the input stages for high input impedance operational amplifiers. These have a very thin base region, achieve gains in excess of 1,000 but have very low breakdown voltages. They are the extreme examples of another source of the trend towards low-power operation — in this case for the higher input impedance that it brings rather than for the low power itself.

The definition and control of operating current becomes difficult at low currents because of the high-value resistors needed, which are not compatible with monolithic processing in its most economic form. Circuit techniques based on the current-

mirror have mitigated this problem, so that all the currents in an amplifier are controlled by a single low-current source. Recent micropower op-amps leave this to the choice of the user, with a single external resistor programming the operating currents of all the transistors in the i.c. At low supply voltages the p.d. across any such resistor, whether internal or external, becomes temperature-dependent and the design problems multiply.

Temperature problems are even more severe in low-voltage/low-current voltage regulators since conventional voltage reference elements cannot be used — the lowest zener diode has a breakdown voltage of approximately 2.7V. Combinations of dissimilar diodes (e.g. Si and Ge) can be produced that have a voltage difference which is almost temperature compensated, while i.c. designs have exploited the properties of forward-biased silicon p-n junctions to achieve the same effect.

The one area of operation where the inherent limitations of micropower operation have not been overcome is the high-frequency region. As the quiescent current in a transistor is reduced, so the rate at which it can charge its own internal and/or external stray capacitances falls. The gain-bandwidth product is an almost linear function of quiescent current (as shown in Fig. 4) with an upper limit to this parameter short of its maximum operating current for most devices. Thus a device normally thought of as a "100MHz transistor", when operated at collector currents below $1\mu\text{A}$ may have a cut-off frequency of less than 10kHz. Clearly it becomes of critical importance to minimize the stray and load capacitances in such applications. For micropower operations at high frequencies, transistors with the very highest quoted gain-bandwidth products should be selected — even 1GHz devices are not out of place provided they can sustain current gain at these low currents. Low-voltage operation brings increased problems since, for example, the collector-base diode has increasing (non-linear) capacitance as the p.d. approaches zero and eventually becomes slightly forward biased.

A major area of concern is in the digital field, where ever larger numbers of gates and other functions are being concentrated into single monolithic i.cs. These l.s.i. (large scale integration) circuits are limited in complexity by two mechanisms — the number of external connections, and the total dissipation. Complementary m.o.s. with its extremely low standby power is the ideal logic family from this latter standpoint and is likely to dominate the market. The dissipation is significant only where high-speed operation is demanded, since then the charge/discharge of internal capacitances dissipates power. Since the choice of circuits available in this family is growing so rapidly, the user is best advised to refer to the manufacturers' data sheets, while the properties of the basic gates will be discussed in the following article.

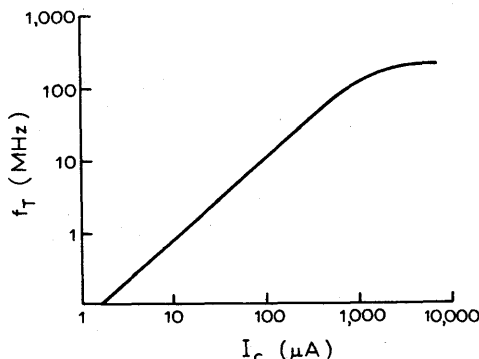
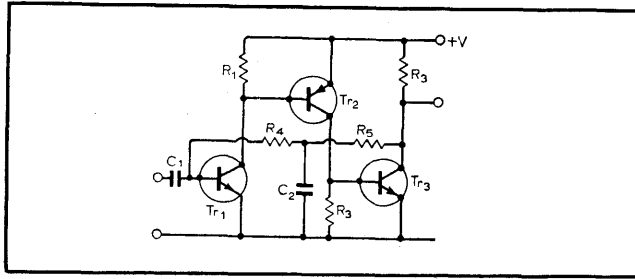


Fig. 4. At high frequencies the gain-bandwidth product of a transistor is an almost linear function of quiescent current.

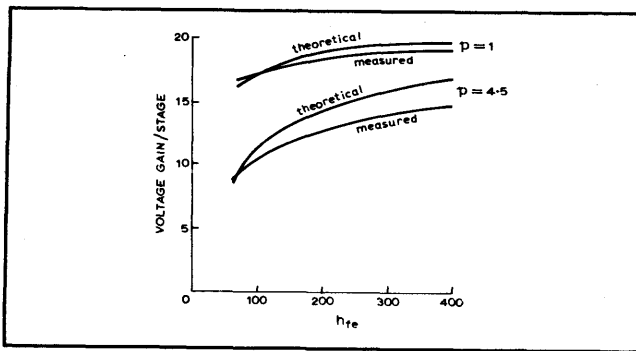
Low-voltage a.c. amplifier



Typical performance

R_1 100k Ω , R_2 22k Ω
 R_3 4.7k Ω , R_4 , R_5 470k Ω
 C_1 , C_2 10 μ F
 Tr_1 , Tr_3 BC125
 Tr_2 BC126
 Supply +1.5V
 Voltage gain 2100
 Output impedance 3k Ω

Input impedance 150k Ω
 (all at 1kHz)
 With $R_1 = pR_2$, $R_2 = pR_3$ etc.
 and with equal h_{re}
 Voltage gain $> \frac{20h_{re}}{h_{re} + 20p}$
 ignoring h_{oe} , h_{re} effects.



Circuit description

At very low supply voltages the base-emitter p.d. of each transistor becomes the limiting factor. For silicon this p.d. is around 0.6V and design is difficult at supply voltages below 1V. This circuit operates from single cells (mercury, Ni-Cd or dry cells) down to 1.0V though the gain increases at higher voltages. In a directly-coupled n-p-n amplifier the p.d. across each collector load is supply dependent as the p.d. is the difference between supply voltage and the V_{be} of the following transistor. This results in varying collector current for each transistor, and hence a large variation in gain is possible.

Here, the current in Tr_1 is determined by the p.d. across R_1 which is in turn fixed by the V_{be} requirements of Tr_2 . The V_{be} of a silicon transistor has a typical variation of <10% for normal ambient temperature, supply and tolerance effects and so Tr_1 current is well-defined. The same argument applies to Tr_2 current in terms of Tr_3 V_{be} . In each case a further advantage of this circuit is that the voltage gain of each stage

can be made to depend largely on the well-defined exponential V_{be} characteristics, rather than the widely varying h_{re} . As h_{re} increases so does the input impedance of a common emitter stage operating at constant emitter current. Where the transistor is fed from a relatively low impedance there is a corresponding fall in base current helping to offset the increase in current gain. A compromise is reached between the two extremes of (a) equal resistors throughout, wasting power in the earlier stages with relatively low input impedance and current gain but high voltage gain and (b) current levels increasing progressively from input to output, maximizing the input impedance and current gain but with lower voltage gain that is more dependent on h_{re} .

Component changes

Tr_1 , Tr_3 can be n-p-n small-signal transistor, particularly those with high current gain at low currents e.g. 2N2484, BC109, etc.

Tr_2 : p-n-p complement to above.

R_1 , R_2 , R_3 : see circuit description for ratios. Resistor R_3 determines output current ($\approx V_+/2R_3$), and R_1 indirectly the input current ($\approx V_{be}/h_{re}R_1$). R_1 : 4.7 to 470k Ω , R_2 : 4.7 to 100k Ω ; R_3 : 1 to 47k Ω .

R_4 , R_5 should be $< 10R_1$ if Tr_1 base current is not to develop too large a p.d. across R_4 , R_5 pushing output voltage towards +V.

C_1 , C_2 determine lower frequency cut-off. Capacitor C_2 decouples negative feedback, gives rise to inductive term in input impedance at low frequencies. C_1 , C_2 : 0.1 to 500 μ F.

Supply: circuit operates down to 1.0V with reasonable gain, to 0.85V with reduced gain and output swing. Upper limit set by transistor V_{ce} ratings—increase R_3 to limit Tr_3 current.

Circuit modifications

- Negative feedback can be used to control gain, define input impedance, lower output impedance. Phase shift over three stages allows oscillation unless $R_8 \gg R_7$ when n.f.b. may be too small to be of value. Feedback can be increased if dominant CR lag (R_6 , C_3) cuts gain before phase shift reaches 180°. Resistor $\approx 0.1R_1$ to $0.01R_1$. C_3 R_1 time constant sets open-loop 3-dB point—may need to bring upper 3-dB point down to <100Hz on open-loop for stability with heavy feedback. (left)

- Output swing can be increased by replacing R_3 by constant-current load provided by current mirror. (middle)

- Direct-coupling possible to sources with near-zero direct voltage if R_9 , R_{10} chosen in conjunction with R_7 (typically $R_9 + R_{10} \approx R_7$ for supply of 1.0V). Decoupling required if supply impedance is significant. (right)

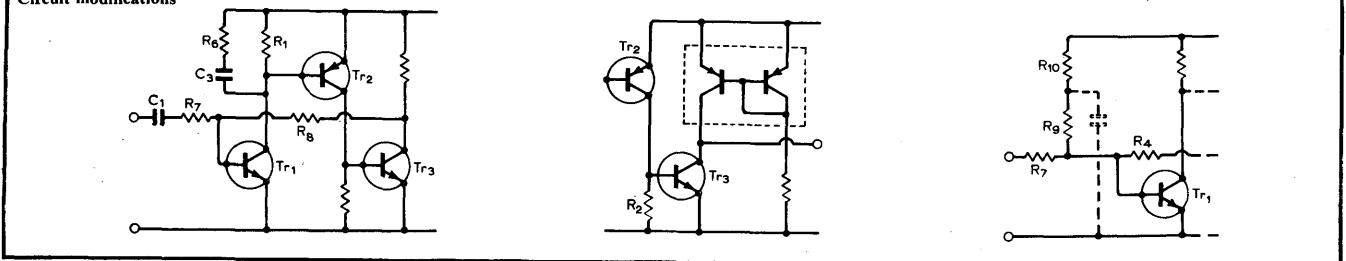
Further reading

Meindl, J. D., & Hudson, P. H., Low-power linear circuits, *IEEE J. Solid-State Circuits*, vol. 1, 1966, pp. 100–11.

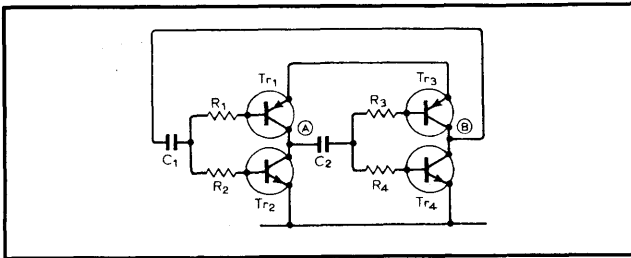
Meindl, J. D., *Micropower Circuits*, Wiley, 1969, pp. 45–50 & 67–81.

Grebene, A. B., *Analog Integrated Circuit Design*, Van Nostrand, 1972, pp. 255–60.

Circuit modifications



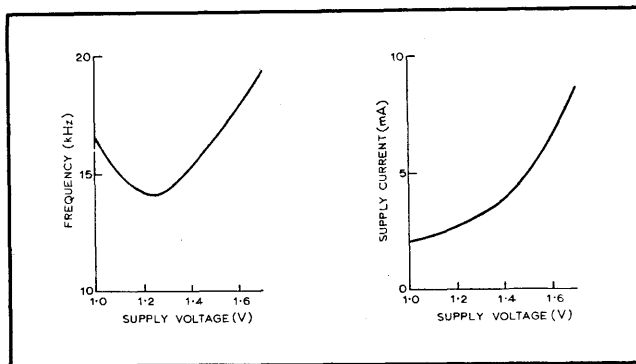
Low-voltage astable circuit



Typical performance

Tr₁, Tr₃ BC126
 Tr₂, Tr₄ BC125
 R₁ to R₄ 1kΩ

C₁, C₂ 10nF
 Supply 1.2V
 f 14.1kHz
 R_L (A-B) 470Ω



Circuit description

The conventional two-transistor astable circuit is difficult to design for very low voltages. The collector load resistance places a severe limit on the current that can be delivered to an external load. In addition, the rise-time of the output waveform, being controlled by the passive components, is much greater than the fall-time, while the rise-time improvement methods that have been devised are not readily applicable as they introduce additional diode drops. The circuit shown is a complementary astable that can operate to as low as 1V using silicon transistors, has equal rise and fall times and gives an output voltage swing almost equal to the supply voltage. It is thus ideally suited for dc to dc convertors and has high efficiency, because of the low standby power and the high utilization of the very low supply voltage. If the voltage exceeds 1.5V the advantages are lost and the

current increases sharply. Frequency and mark-space ratios are most easily controlled by C₁, C₂ while the resistance values can be raised if the load resistance is high. Using germanium transistors the minimum and maximum supply voltages are more than halved. Leakage currents prevent the use of germanium transistors at very low current, but good efficiency is possible at higher currents.

Component changes

Tr₁, Tr₃ : Any p-n-p silicon transistor for given supply voltage range. For higher currents use BFR81.
 Tr₂, Tr₄ : n-p-n, otherwise as above (BFR41). For lower supply voltages substitute germanium complementary pairs 2N1302/03 or 2N1304/05.
 R₁ to R₄ : 100Ω to 100kΩ. Use lower values with high current transistors and low load resistance. High resistance minimizes off-load current but suitable for high load resistances.
 C₁, C₂ : 100pF to 100μF depending on frequency required. At high frequencies charge storage effects increase no-load current.

Circuit modifications

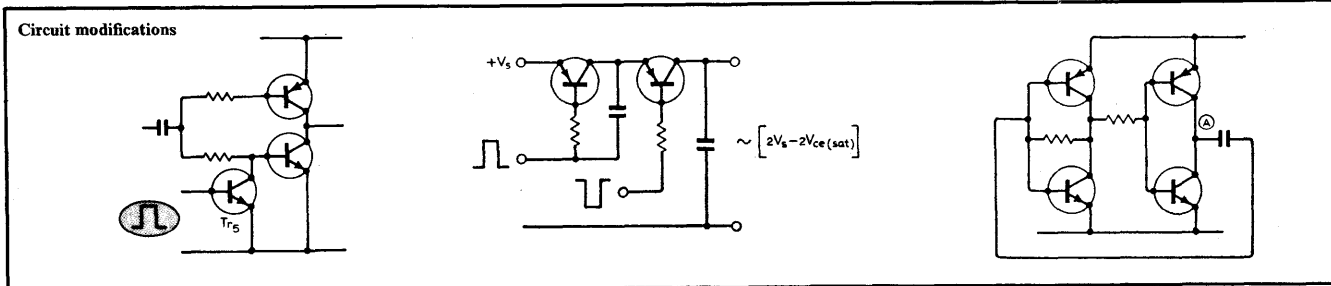
- By adding Tr₅, which may be an n-p-n device fed with a positive-going pulse with respect to the zero line or a p-n-p device driven negative with respect to positive line, the circuit may be either inhibited if the pulse is of long duration or synchronized if the pulse is of short duration and repetitive.
- The circuit is ideally suited to dc-dc convertor applications and efficiency can be high. One possible configuration is shown making use of antiphase outputs of the oscillator to switch transistors in and out of conduction. (middle)
- As with other dual inverter astables, alternative circuits can be produced (see series 1 cards in which a single capacitor controls both parts of the cycle. One output waveform A is normally distorted by the unbalanced loading. As shown, the upper voltage limit is restricted to ≈1.2V.

Further reading

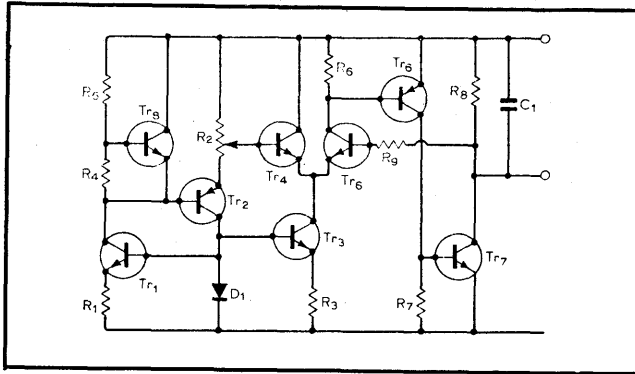
Poirier, N., Cochran, B. L., *et al*, Unique hybrid transformerless dc-to-dc converter suitable for biomedical applications, *IEEE International Solid-state Circuits Conference Digest* 1971, pp. 110/1.
 Ho, C.f., Zero quiescent current monostable multivibrator, *Radio and Electronic Engineer*, 1969, pp. 22-4.

Cross references

Series 10, cards 4 & 8.

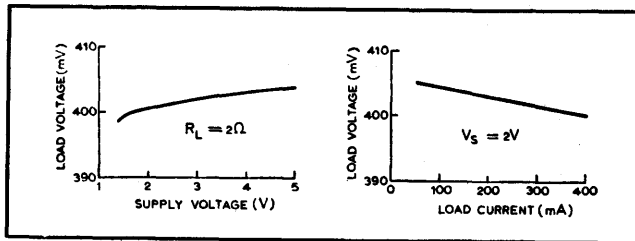


Low-voltage regulators



Typical performance

- | | |
|-----------------------------|------------------------------------|
| R_1, R_2, R_3 1k Ω | Tr_6 BC126 |
| R_4 12k Ω | Tr_7 MJE521 |
| R_5, R_6 3.3k Ω | Supply 2V |
| R_7, R_8, R_9 1k Ω | R_L 2 Ω |
| C_1 100 μ F | V_L 400mV |
| Tr_1, Tr_3 2N1302 | Output res. 0.01 Ω |
| Tr_2 2N1303 | For $\Delta V_B/R_L \pm 20\%$, |
| Tr_4, Tr_5, Tr_8 B2125 | $\Delta R_L/R_L \pm 20\%$, |
| | $\Delta V_L/V_L \approx \pm 0.5\%$ |



Circuit description

This circuit is relatively complex because the problems of voltage regulation at very low voltages are severe. Transistors Tr_1, Tr_2 form a low-voltage ring-of-two reference (see series 6 card 9) providing the dual function of a variable temperature-stabilized reference voltage for the base of Tr_4 , with a bias voltage for Tr_3 . This latter provides a constant current for the long-tailed pair Tr_4, Tr_5 , which compare the reference voltage with the regulator output. Transistors Tr_6, Tr_7 amplify any imbalance controlling the output. It is important that the output transistor operates in common-emitter to minimize the wasted p.d. For low $V_{ce(sat)}$ -transistors the regulated output can be maintained for supply voltages down to a few tens of millivolts above the required output. Choice of reference circuit limits the minimum supply voltage; if D_1, Tr_1, Tr_3 are replaced by a current mirror (series 6 card 4) supply voltages down to 1V can be accommodated. At

slightly higher supply voltages a monolithic micropower op-amp may be used with one or more additional output transistors to increase the current capabilities. In principle circuits such as this are able to operate at supply voltages up to the breakdown voltages of the transistors as each stage is designed to work at a current that is largely independent of the supply voltage. The circuit as shown is capable of handling relatively large currents but may be used at currents down to $< 1mA$ if required.

Component changes

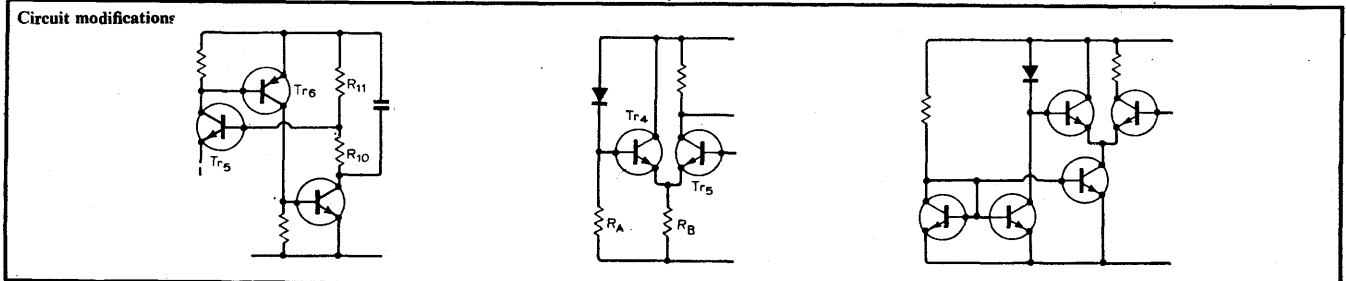
- R_{1-3}, R_{7-9} : Not critical but should be comparable in value and depend on output current required. Typical range 330 Ω to 2.2k Ω .
- R_4, R_5 : Ratio used to adjust temperature coefficient of reference voltage. Keep ratio R_5/R_4 low to minimize supply voltage e.g. $R_5 \rightarrow 0, R_4 \infty$ leading to temperature drift of +0.1% deg. C.
- Tr_{1-3} : Must be germanium for these low voltages. Any general-purpose complementary germanium transistors with Tr_4, Tr_5 at equal temperature for low drift.
- $Tr_{4,5,6}, Tr_8$: General-purpose silicon e.g. 2N2926, ME4103 for n-p-n; ME0413, 2N3702 for p-n-p.
- Tr_7 : High current n-p-n for low saturation voltage; voltage rating unimportant.
- C_1 : $> 10\mu$ F.
- D_1 : General-purpose silicon diode.

Circuit modifications

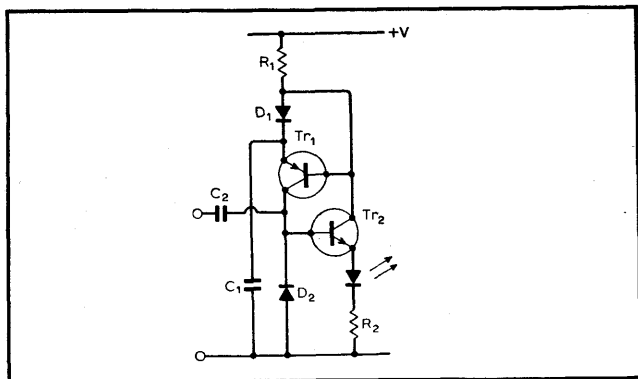
- For output voltages greater than the reference, the reference is compared with a fraction of the output i.e. $V_1 = (R_{10}/R_{11} + 1)V_{ref}$. Generally the reference voltage must be significantly less than the minimum supply voltage, so that the current in the reference element may be sustained with some accuracy. The output may need to be close to the supply minimum. (left)
- The circuit may be simplified where the output is required to have a negative temperature coefficient, as for card 9. It becomes more susceptible to supply changes but by returning R_A (and R_B if possible) to the stabilized line. (middle)
- An alternative method for improving stability against supply variations while minimizing the required supply voltage, is to use matched transistors to form current mirrors, defining both the reference current and that for the long-tailed pair. (right)

Further reading

- Widlar, R. J., New developments in i.c. voltage regulators, *IEEE I.S.S.C. Digest*, 1970, pp. 158/9.
- Williams, P., D.C. regulator for low voltages, *Electronic Engineering*, March 1970, pp. 41-3.



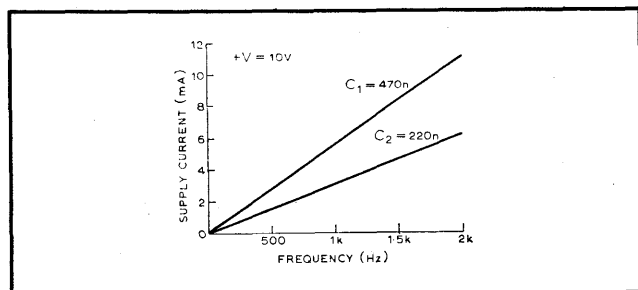
Optical link with low standby power



Typical performance

R_1 47 Ω , R_2 10 Ω
 R_3 , C_1 470nF, C_2 2.2nF
 D_1 , D_2 1N914

Tr_1 , BC126
 Tr_2 BC125
 Supply +10V, 5.4mA
 f 1kHz
 Standby current <0.1 μ A



Circuit description

One method for minimizing power in systems is to have as many circuits as possible working close to zero quiescent current, conducting only on the appearance of some signal it is desired to detect. For example, the c.m.o.s. Schmitt trigger of series 2 has virtually zero quiescent current but can deliver a sharp transition to a load or hold that load on until the disappearance of the signal. The present circuit uses complementary bipolar transistors with D_1 holding the circuit in the non-conducting state and with C_1 storing charge. On receipt of a positive-going transition Tr_2 begins to conduct, pulling the base potential of Tr_1 below the emitter potential as stored on C_1 . The cumulative increase in current in Tr_1 and Tr_2 discharges C_1 (with D_1 reverse-biased) through the transistors and the l.e.d. The short duration pulse may be at a relatively high current ensuring high intensity from the l.e.d., without excessive mean dissipation. After C_1 has discharged to the level at which Tr_1 , Tr_2 come out of conduc-

tion it is recharged through D_1 , R_1 and is ready to detect the next positive transition. D_2 absorbs the intervening negative transition. The charge lost by C_1 is C_1V for each pulse and with a repetition frequency f , the supply current is around fC_1V (V is the supply voltage).

Component changes

R_1 : This limits rate of recharge of C_1 and hence recovery time. Typically 39 to 150 Ω with higher value better at low repetition rates.

R_2 : Limits peak current in l.e.d. may be lowered to 3 Ω for l.e.d.s capable of >1A peak current. If high intensity not required may be increased to >100 Ω .

D_1 , D_2 : Not critical, any general-purpose Si diode, though high-speed preferable for short pulses.

Tr_1 , Tr_2 : Complementary n-p-n/p-n-p silicon transistors. Parameters not critical but peak current rating must be sufficient for required l.e.d. peak current.

Supply: +5 to +20V. At low voltages, high-intensity light pulses difficult to achieve.

L.e.d.: Any low-cost l.e.d. may be used, but optical coupling between l.e.d. and detector must be good, as light intensity is relatively low and spectrum differs from that of standard silicon photodetectors. Alternatives include infra-red emitters (TIL31/32 etc.).

Circuit modifications

● An identical circuit may be used as the detector, again with monostable characteristics, such that an output pulse is delivered to a load on receipt of a short duration optical pulse. Detector and l.e.d. may be optically linked by light fibres, but R_3 , R_4 may have to be one or more orders of magnitude higher than R_1 , R_2 with inefficient coupling. (left)

● Alternatively a phototransistor may replace Tr_1 and the photodiode D_2 (middle). This applies equally if the l.e.d. and phototransistor form an opto-isolator. Standby current remains low in each of these circuits.

● The low-power characteristics of c.m.o.s. provide an alternative solution. Using one non-inverting buffer (or two inverters) with positive feedback through R_2 a Schmitt trigger action is obtained as on Circards series 2 card 3. Sensitivity is set by R_1 , hysteresis by the ratio of R_2/R_1 and the output has good rise- and fall-times for operating following circuits.

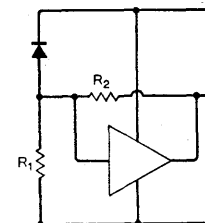
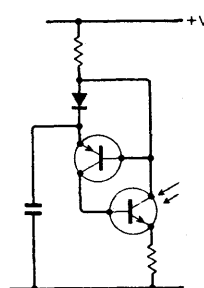
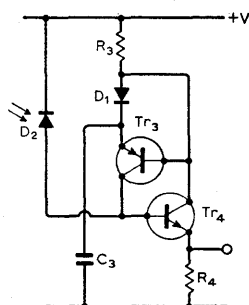
Further reading

Van der Geer, C. A. J., Optical-trigger-link scheme minimizes stand by power, *Electronic Design*, ref. 18, 2 september 1972, p. 62.

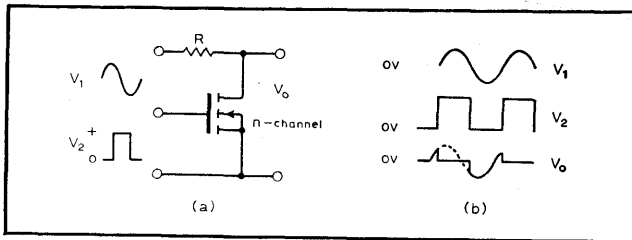
Cross references

Series 9, cards 3, 4 & 11.

Circuit modifications

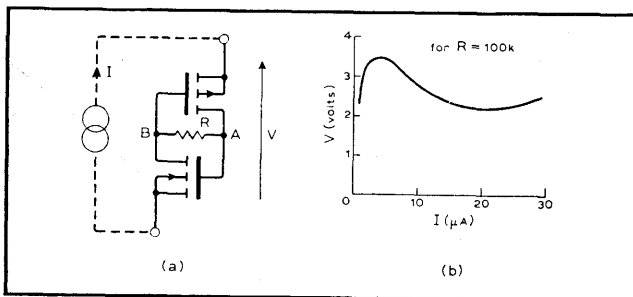


Signal-powered circuits



Circuit description—1

Chopper circuits, phase detectors and wave-shaping circuits are among the types that can be designed without a separate d.c. supply. They are not intrinsically micropower circuits though they can be, but are useful in locations where the provision of a separate supply voltage is difficult. Any amplifying device (bipolar, j.f.e.t. or m.o.s.f.e.t.) can be switched into and out of its conducting state by application of a square/rectangular waveform to its control electrode. An enhancement-mode m.o.s.f.e.t. can be driven from near open-circuit down to an on-resistance $\ll 1k\Omega$ by a square-wave applied to its gate. When fed via resistor R with a sine-wave of the same frequency, the output appears only for $V_2 \neq 0$, $V_1 \neq 0$. Applying this output to a high-impedance meter, e.g. a sensitive moving-coil instrument, gives a mean reading dependent on the phase-difference of the signals. It is also proportional to the amplitude of V_1 which needs to be controlled if the output is to indicate phase difference accurately (see circuit 4). The m.o.s.f.e.t. may be a single device from an existing c.m.o.s. i.c. such as CD4007.

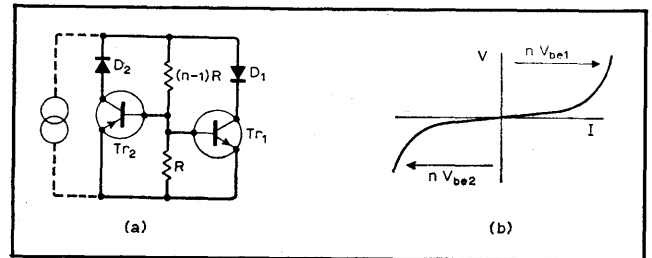


Circuit description—2

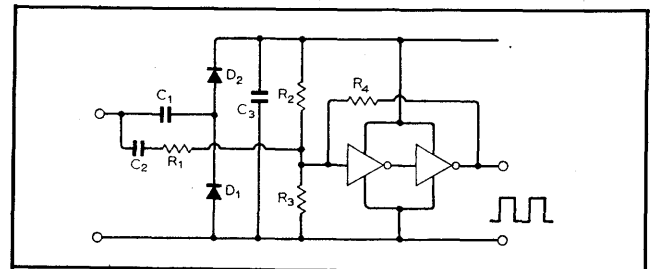
Devices may be combined to give non-linear V/I characteristics which may be used for wave-shaping. A particular example is the generation of a controlled negative-resistance region. This may be used in both astable and sinusoidal oscillators as well as in amplifiers. The major difficulty is the accurate control of the slope and end-points of the negative resistance region. In the circuit shown, the use of enhancement-mode m.o.s.f.e.t.s (again from CD4007 or similar for economy) ensures that gate-current effects can be ignored even where negative resistance is required at micropower levels. At low currents the total p.d. is of the order of the sum of the device threshold voltages ($\approx 3.4V$), while at higher currents this is reduced by the resulting p.d. developed across R. Similar circuits using bipolar transistors have been developed which can also operate without a separate power supply provided they are embedded in a system that can bias them into the negative-resistance region.

Circuit description—3

Other non-linear devices have application to a.c. systems, where it is required for example to change a sinusoidal



waveform into an approximate square-wave without providing a d.c. supply to the shaping device. Simple passive components such as zener diodes offer one solution, but circuits such as the "amplified-diode", current-mirror may be used, if complementary-pairs of transistors are used. As shown, Tr_1 conducts for positive currents with the terminal p.d. being an amplified version of its V_{be} . The potential divider resistances are a compromise, minimizing current flow at low-terminal p.d.s while avoiding excessive loading by base currents at higher levels. Diodes D_1 and D_2 prevent conduction via the collector-base paths of Tr_1 , Tr_2 respectively when the applied polarity is that intended to bring the other device into conduction. The complete circuit approximates to a two-terminal a.c. zener of variable breakdown voltage, though with limited accuracy.



Circuit description—4

If a low-impedance alternating voltage is available in a system it can be used to provide a low-power d.c. supply via a voltage doubler such as that formed by C_1 , D_1 , D_2 and smoothed by C_3 . The voltage may then be processed in any way desired provided that the circuit used to do this can operate from this supply voltage while consuming the minimum power. The example shows a pair of inverters (as in card 10) connected as a Schmitt trigger circuit, giving an output square-wave whose amplitude could be stabilized if the supply voltage is large enough to be fed via a zener diode or other voltage regulator. Similarly micropower op-amps can be used in any of their normal measurement or signal-processing configurations. A second voltage-doubler giving a negative supply voltage may be required.

Further reading

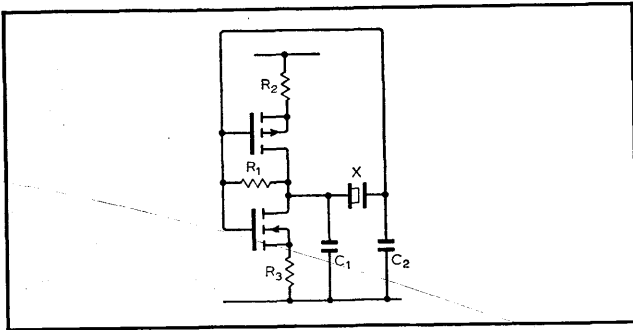
Worstell, G., Phase-lock detector requires no external power supply, in 400 Ideas for Design, Hayden vol. 2, 1971, p. 223.
Williams, P., The Amplified Diode, *Design Electronics*, Jan. 1968, pp. 26-9. Hart, B. L. & Barker, R. W. J., Minimum-power waveform generation circuit technique, *Electronics Letters*, vol. 8, pp. 585/6.

Grothe, L. E., Sine-to square-wave converter is self-powered, in 400 Ideas for Design, Hayden vol. 2, 1971, p. 78.

Cross references

Series 10, cards 3, 7 & 10.

Micropower crystal oscillator



Typical performance

IC 1/3 × CD4007AE

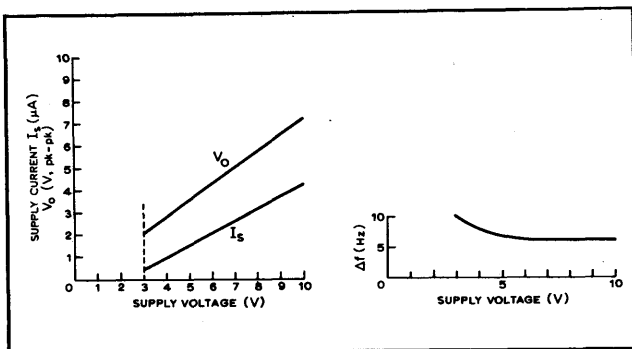
R₁ 15MΩ

R₂, R₃ 100kΩ

C₁ 15pF, C₂ 10pF

Crystal 256kHz nom.

Supply +5V, 2.8μA



Circuit description

High-frequency oscillators present particular problems at low currents and voltages. In general, the gain of the active device falls, while the shunt reactance of stray capacitance becomes more important as the load impedance has to rise to maximize the gain at low currents. For the best frequency stability quartz crystal oscillators are obligatory and a convenient configuration is a π -network containing the crystal and two capacitors, interconnected with an amplifier having high input and output impedances. A c.m.o.s. inverter meets the requirements for the amplifier, particularly with resistors R₂, R₃ providing series-derived series-applied negative feedback. The loading on the π -network is controlled by R₁ and the voltage gain of the inverter via the Blumlein (Miller) effect. The quiescent current is very low, limited further by R₂, R₃, and at low frequencies the current necessary to charge and discharge the capacitors is minimal. At higher frequencies this effect dominates and the consumption becomes an almost linear function of frequency. Fine trimming of frequency is

achieved by varying C₁, and while the precise temperature coefficient is thereby adjusted the effect is dominated by the cut of the crystal in use. Where high stability is not required C₁ and C₂ may be omitted with oscillation being dependent on the presence of stray/circuit capacitance. The data is based on an industrial grade i.c., but specially developed circuits are available for supply voltages down to 1.5V or less.

Component changes

R₂, R₃: should be equal. Max. value 470kΩ; minimizes current drain, makes circuit more critical of loading, applicable only at frequencies \ll 1MHz. Can be reduced to zero with waveform distortion, frequency shift, higher current.

R₁: not critical. Determines bias voltage condition but input current >0 so R₁ may be very large. Lowering R₁ to 10 to 100kΩ controls amplitude of oscillation at expense of loading π -network.

C₁, C₂: 1 to 100pF. Varies considerably with crystal used; may be eliminated if stability not critical and significant strays present. Varying C₁ gives fine frequency control for optimum stability.

Crystal: Oscillators possible from <100 kHz to >2.5 MHz. Reduced amplitude at high frequencies. If R₂, R₃ too low, second or higher harmonic may be excited. Use passive LC circuit to select fundamental/harmonic.

Supply: As low as 3V with CD4007AC, down to 1.1V with TA6178 development type.

Circuit modifications

● The simplest circuit eliminates R₂, R₃ retaining R₁ for bias with possibly a series resistor R to control the amplitude of oscillation. Amplifier characteristics are no longer controlled by series negative feedback and performance is not as good. As before stray capacitance may be sufficient to allow oscillation, but not recommended for good stability. (left)

● Still simpler circuits using a single m.o.s. inverter gives reasonable performance. Using depletion mode device, R₁ may be replaced by gate resistor to ground with series resistor in source to limit gain. (middle)

● For coupling to following logic circuits a single inverter may suffice. At 5V and with total load capacitance of ≈ 20 pF, the current consumption at 250kHz is ≈ 60 μA and rise and fall-times ≈ 0.3 μs. For sharper transitions at low frequencies, replace single inverter by Schmitt trigger. (right)

Further reading

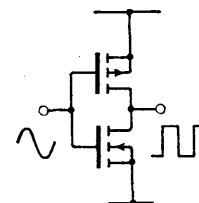
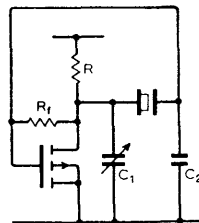
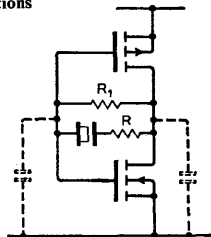
Eaton, S. S., Timekeeping advances through c.o.s.m.o.s. technology, RCA application note ICAN-6086.

Meindl, J. D., Micropower Electronics, Wiley, 1969, p. 94.

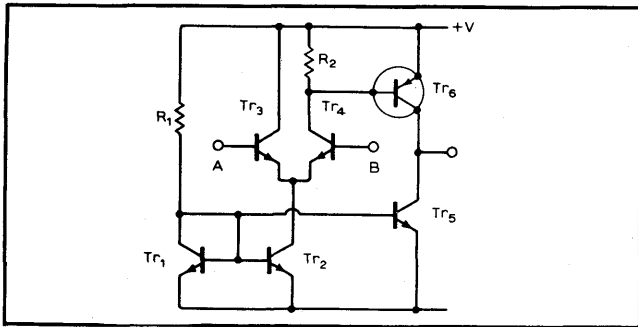
Cross references

Series 10, cards 2 & 8.

Circuit modifications



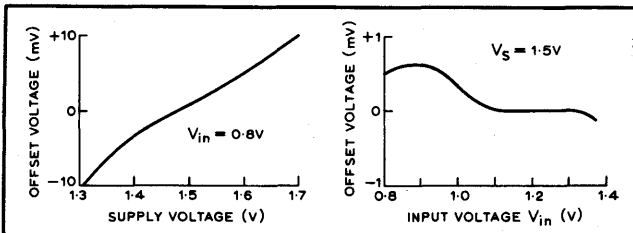
Micropower d.c. amplifier



Typical performance

Tr₁ to Tr₅ CA3046
 Tr₆ BC126
 R₁ ≈ 47kΩ (vary for minimum offset)
 Supply +1.5V
 Input current 0.13μA

Common-mode range 0.8 to 1.3V used as voltage follower with A to output and B to input.



Circuit description

Monolithic i.c. micropower amplifiers are becoming increasingly available. They have performance comparable with conventional designs but at supply voltages down to ±1V and with currents presettable to very low values. For still lower voltages, to ±0.75V with restricted performance variants of the circuit shown may be used. Transistors Tr₁ to Tr₅ are from a general-purpose i.c. CA3046; Tr₁, Tr₂ and Tr₅ constitute a current mirror system that define the currents in the long-tailed pair Tr₃, Tr₄ and the output stage Tr₆ in terms of the current in R₁. At low supply voltages the p.d. across R₁ will be little greater than that across R₂ (≈0.6V) while R₂ should carry rather less current. Hence R₁ = R₂ is a reasonable compromise. The quiescent current in the output stage is low, and the circuit as it stands is suitable for use with grounded loads—with a centre tapped supply, the negative load current would be restricted to the quiescent current. This amplifier may be used as a d.c. amplifier, comparator, etc. just as a normal op-amp but with restricted performance. In particular,

the loop gain parameters are such that with heavy feedback the frequency response is restricted by the compensation required to avoid oscillation. A second disadvantage is that there is no current regulation against supply voltage changes, though this is shared by some micropower op-amps. The amplifier is indicative of low-cost micropower design.

Component changes

Tr₃, Tr₄: Any matched pair of silicon n-p-n transistors—preferably on a single monolithic chip for best matching (at low currents).

Tr₁, Tr₂, Tr₅: Matched triple as above. All five are conveniently available in i.cs such as CA3045, CA3046. Alternatively, resistors inserted in emitters of Tr₁, Tr₂, Tr₅ dropping 50 to 500mV equalize and define currents at expense of increased minimum supply voltage.

R₁, R₂: These will be comparable in value, with R₁ increased at higher supply voltages such that the current in R₂ is ½ × current in R₁.

Supply voltage range: May be as low as 1V with increased drift via R₁. See Circards series 6, constant-current circuits.

Circuit modifications

- There are many complete i.cs appearing capable of operating down to ±1V or less. These will give better performance than the simple circuit shown. The advantage of flexibility and even lower supply voltages can be enhanced by additional transistor(s) to improve equality of currents in Tr₃, Tr₄. In circuit left Tr₇ absorbs much of the current from Tr₃, leaving smaller and more nearly equal currents in R₂, R₃ and hence in Tr₃, Tr₄.

- Middle circuit shows how a p-n-p current mirror may be used if matched pairs are available. Similar techniques are commonplace in the monolithic i.cs mentioned above. As Tr₇ and Tr₈ carry the same current, then provided Tr₆ base current remains small, so also must Tr₃ and Tr₄.

- As with any other d.c. amplifier negative or positive feedback can be applied to the gain or the hysteresis if used as a level sensing circuit (Circards series 2). (right)

Further reading

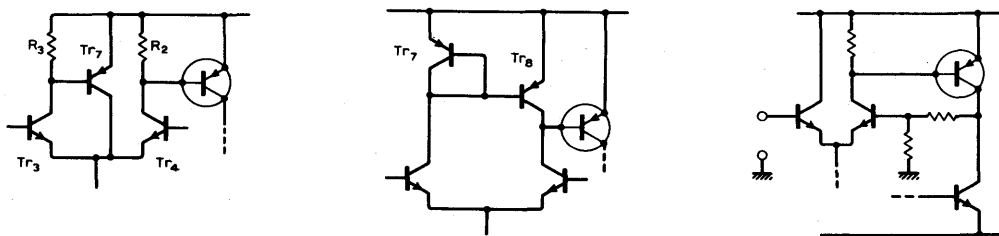
Harden, W. R. & Hellstrom, M. J., Triple-channel micropower operational amplifier, *IEEE J. Solid State Circuits*, vol. SC-4 1969, pp. 236-40.

Bittmann, C. A. *et al*, Technology for the design of low-power circuits, *IEEE J. Solid State Circuits*, vol. SC-5, 1970, pp. 29-37.

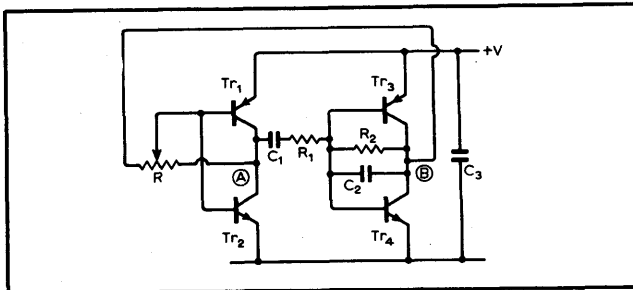
Cross references

Series 10, cards 1, 3 & 9.

Circuit modifications

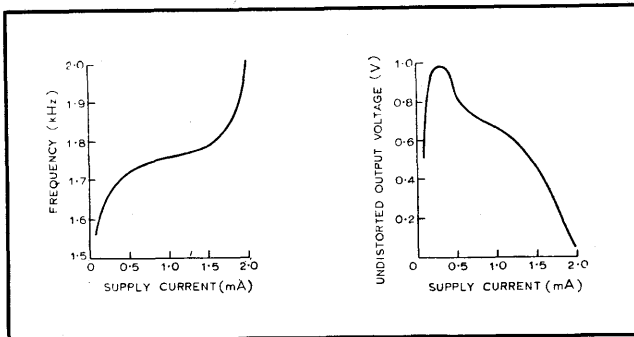


RC oscillator for low voltages



Typical performance

R_1 4.7k Ω
 R_2, R 10k Ω
 C_1 22nF, C_2 10nF
 C_3 100 μ F (optional)
 Tr_1, Tr_3 matched p-n-p pair from CA3084
 Tr_2, Tr_4 matched n-p-n pair from CA3046
 I_s 1mA
 R set for 0.4Vpk-pk at A
 f 1.77kHz



Circuit description

The same inverting stages are used as in the astable oscillator but they are now operated in the linear mode as with the output stage of the class B amplifier. Careful biasing is again necessary to minimize distortion without excessive quiescent current. This may be achieved by supplying from a voltage source having a negative temperature coefficient of $\approx -4\text{mV}/\text{deg. C}$ i.e. two p-n junctions. A novel alternative is possible for equal swings at A and B which for high gain transistors will result when $R_1 = R_2/2$, $C_1 = 2C_2$. Then R is centre tapped for oscillation to just commence. Under these conditions and with class A operation Tr_1, Tr_3 and Tr_2, Tr_4 can be considered as long-tailed pairs receiving antiphase signals and a constant-current supply can act as the "tail" i.e. this is a circuit that can work from either a constant voltage or a constant current source. The RC network is a Wien network,

and finite gains in the transistors prevent the ideal balance conditions applying, with some resulting distortion and a frequency of oscillation that departs from the ideal value $f = 1/2\pi R_1 C_1 R_2 C_2$. R is adjusted to sustain oscillations at the required amplitude.

With the circuit as described antiphase outputs are available at A and B which may also be used to drive a load connected between them.

Component changes

R_1, R_2 : 100 Ω to 100k Ω , $R_1 = R_2/2$.

C_1, C_2 : 1nF to 10 μ F, $C_1 = 2C_2$.

Frequency of oscillation that can be sustained falls with falling current. Setting of R controls the onset of oscillation.

$R \approx R_1$.

Tr_1 to Tr_4 : Choice restricted to pairs from same monolithic chip or matched pairs kept at equal temperature.

Circuit modifications

● By adding a non-linear network to the feedback path the loop gain may be reduced as the oscillation amplitude exceeds a given level. With the low supply voltage and current lamp/thermistor methods are inappropriate but germanium or Schottky diodes can be used. The non-linear transconductance of the transistors together with the collector-emitter saturation characteristics complicate the design. (left)

● An alternative low-voltage oscillator uses a complementary pair as a unity negative-impedance-converter. To ensure oscillation $C_1 > 2C_2$ and $R_1 < R_2/2$. Fixing C_2 at 0.01 μ F, C_1 at 0.022 μ F for example allows R_1 to be varied about $R_2/2$ as required. The supply is from a constant-current generator. (middle)

● As with other oscillators, by setting just below oscillation a band-pass filter can be obtained (see Circards series 1). The signal should be injected from a reasonably high impedance as the input point is an imperfect virtual earth in this particular case. (right)

Further reading

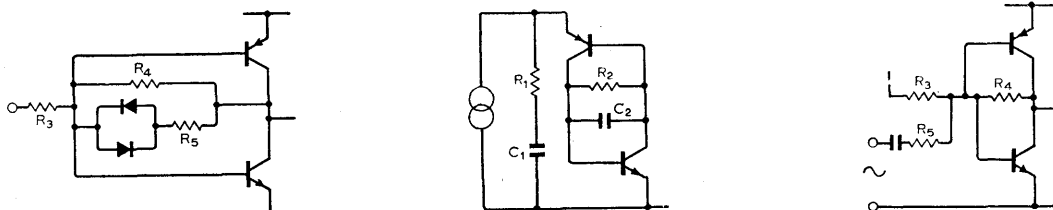
Williams, P., Wien oscillators, *Wireless World*, Nov. 1971, pp. 541-7.

Hachtel, G. D., & Pepper, R. S., Synthesis of integrable nearly-sinusoidal potentially bistable oscillators, *IEEE Journal of Solid State Circuits*, vol. SC-1, 1966, pp. 111-17.

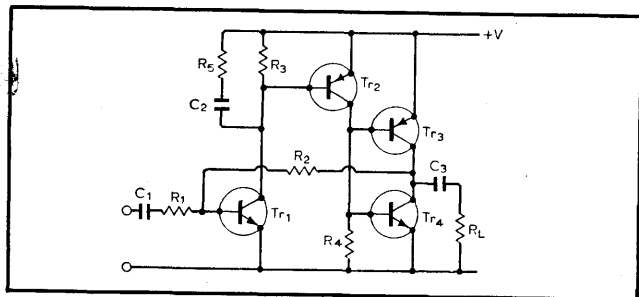
Cross references

Series 10, cards 1, 2 & 9.

Circuit modifications



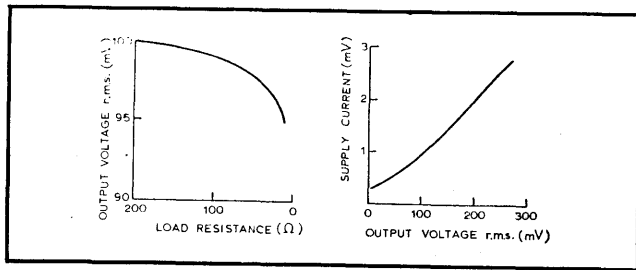
Class B low-voltage amplifier



Typical performance

R_1, R_2, R_3 100k Ω
 R_4 4.7k Ω , R_5 560 Ω
 C_1 1 μ F, C_2 3.6nF
 C_3 100 μ F
 Tr_1 BC125, Tr_2 BC126
 Tr_3 BFR81, Tr_4 BFR41

Supply 1.2V, 2.1mA
 R_L 100 Ω
 V_O 370mV r.m.s. (1kHz)
 Distortion 5%
 Efficiency 51%
 Quiescent power 0.43mW



Circuit description

This circuit has strong similarities with the standard "complementary emitter-follower" output amplifiers used for audio systems. The difference is critical. By changing the output transistors into the common-emitter mode the output swing and efficiency at low supply voltages dramatically improve. This is because the output can swing to within $V_{ce(sat)}$ of each supply line while the common base drive voltage is small and restricted to the mid-supply region. With emitter follower output, saturation of the output transistors would require a base drive voltage extending by $\approx 0.6V$ above and below the supply lines i.e. an auxiliary supply would be needed. There remains a serious disadvantage to this form of circuit. For high efficiency class-B operation demands that the supply voltage be fixed at $2V_{be}$, and more particularly at a value of V_{be} that keeps the transistors just on the edge of conduction. If higher crossover distortion is permissible supply voltages down to 1.1V are possible. Where the temperature and/or supply voltage is variable the circuit quiescent current may increase too greatly. One solution is to provide a simple voltage regulator whose temperature coefficient

matches that of the transistor V_{be} 's (see card 3). The preceding stages are basically those of the amplifier of card 1 and the same constraints on feedback/compensation apply.

Component changes

Tr_1, Tr_2 : Any general-purpose n-p-n/p-n-p silicon planar transistors preferably with high h_{fe} at low currents.
 Tr_1 : ME4103, BC109, 2N2926, 2N930, 2N3707, etc.
 Tr_2 : ME0413, BC179, 2N2904 etc.
 Tr_3 : Low saturation voltage p-n-p: BFR80, BFR81 etc.
 Tr_4 : Low saturation voltage n-p-n: BFR40, BFR50 etc.
 R_1, R_2 : Set gain and input impedance. Increasing R_2/R_1 increases gain, reduces feedback upsets output quiescent if taken too far. R_1, R_2, R_3 : 10 to 470k Ω .
 R_4 : Typically $R_3/5$ to $R_3/50$.
 C_1, C_3 : Standard input/output coupling. C_1 : 0.1 to 10 μ F, C_2 : 10 to 500 μ F.
 R_5, C_2 : Depends on open and closed-loop gains as well as transistor types. More complex CR networks necessary to maximize bandwidth and ensure stability.

Circuit modifications

- An alternative output stage applicable at slightly higher voltages has been published as the basis of a monolithic i.c. hearing aid amplifier. Tr_2 can only be produced in the standard i.c. as a "lateral" transistor, having low gain. Tr_1 having its collector to the most negative supply line i.e. a "vertical p-n-p" has higher gain, and the combination is a reasonable match to Tr_3 . Supply voltage required is raised to $\approx 1.5V$ compatible with single dry cells. (left)
- Additional bias networks may be added (such as amplified diode) if of low dynamic resistance and/or suitably bypassed so that output bases receive some a.c. drive. These networks should ideally have a positive temperature coefficient, and to avoid excessive quiescent current under adverse conditions the bias may have to be set into the class C region with exaggerated crossover distortion prior to the application of feedback. (middle)
- Some improvement in the output capabilities for the emitter-follower output circuits can be obtained by bootstrapping the preceding collector load. This allows the base drive to be taken past the positive supply voltage and halves the wasted voltage. (right)

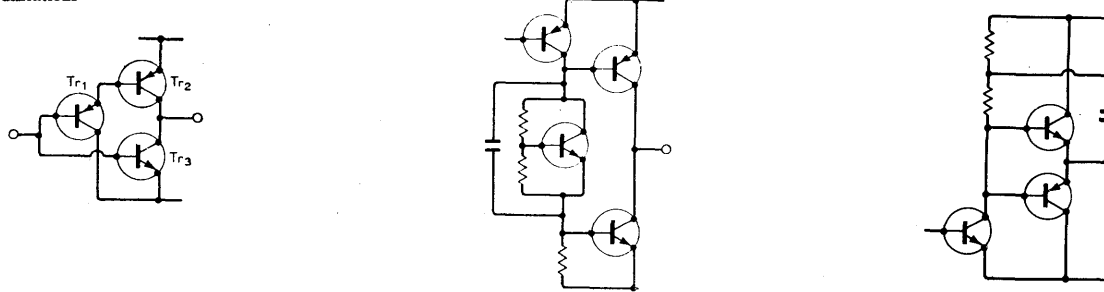
Further reading

Getreu, I. E., Integrated class-B hearing-aid amplifier, *JSSC Digest*, 1971, pp. 182/3.
 Meindl, J. D., *Micropower Circuits*, Wiley, 1969, pp. 59-61.

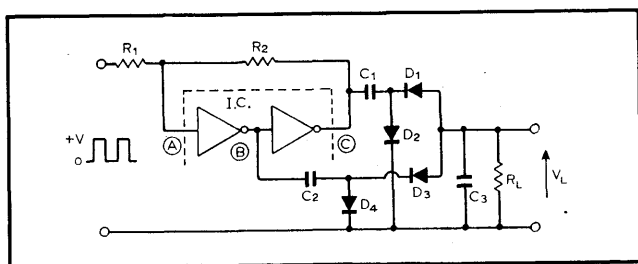
Cross references

Series 10, cards 1 & 8.

Circuit modifications

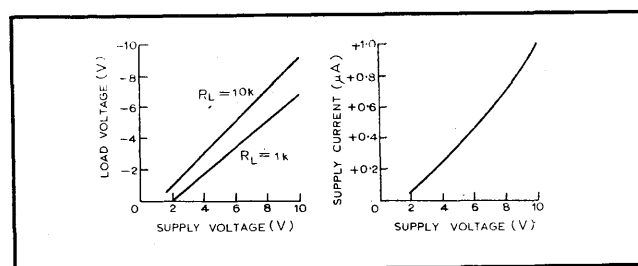


Low-voltage d.c. converter



Typical performance

R_1 470k Ω , R_2 1M Ω	f 100Hz
C_1, C_2, C_3 22 μ F	Supply 5V
IC 1/3 \times CD4009AE	V_I -3.8V
D_1 to D_4 bridge rectifier	Open-circuit current
R_L 10k Ω	1.6 μ A



Circuit description

The circuit is used where a supply of given polarity is available, and a reverse polarity is required e.g. for biasing op-amps or m.o.s. logic circuitry. It simplifies the problems of battery operation by removing the need for a separate negative rail, but efficiency is a key parameter particularly for micropower operations and/or low supply voltages. Basically the circuit is related to the diode pump, with antiphase square waves generated at B and C, with R_1 and R_2 providing hysteresis to speed up the transitions and make them less dependent on input amplitude and wave shape. For micropower applications c.m.o.s. inverters are the obvious choice as standby power is minimal, with a high ratio of available loadpower to standby power.

As point B switches positive C_2 is charged via D_4 to a p.d. less than the supply voltage by one diode forward voltage. When the B returns to zero, the output of C_2 tries to swing negative bringing D_3 into conduction and transferring charge from C_2 to C_3 . At low repetition rates the continuous loss of charge to R_2 prevents a significant accumulation of charge in C_3 . As the repetition rate increases (or C_2 increases) the p.d. across C_3 approaches a magnitude of $|V_s - 2V_f|$ where V_f is the forward voltage across each diode. Output polarity depends only on the diode configuration and not on the original supply voltage. By adding the inverted output at C

and a second diode pump the output capability is increased and ripple reduced (equivalent to full-wave rectification). The output is unregulated but the drift is largely that due to the diode losses i.e. the output change is $\ll 1V$ for temperatures between say 20 and 50°C. Minimum supply voltage is dictated by that of the c.m.o.s. stages, and by the proportionally larger diode losses.

Component changes

Supply: +3 to +15V (for AE series)
 R_L : 470 Ω upwards
 D_{1-4} : Bridge rectifier or 4 \times 1N914, etc.
 R_1 : 10k Ω to 10M Ω .
 R_2 : 22k Ω to 22M Ω .
 C_1, C_2 : 100nF to 100 μ F.
 C_3 : 1 to 100 μ F.
 IC: Any c.m.o.s. containing 2 inverters (CD4009, CD4049) or NAND/NOR gates used as inverters (CD4001).
 f : 10Hz to 100kHz.

At low values of R_L , the on-resistances of the m.o.s. devices within c.m.o.s. inverters limit output. Choose R_1, R_2 high to minimize standby current or with high load resistance. Too high values allow possibility of hum pickup. High frequencies simplify smoothing, allow lower C values, but increase standby current.

Circuit modifications

- Only one of the diode/capacitor networks need be used where max. output not required. If c.m.o.s. output already available in system only one inverter needed for full-wave system.
- Both positive and negative outputs can be obtained simultaneously, each approaching pk to pk input less diode losses (left). Similarly voltage multiplier circuits may be used to obtain increased voltage outputs, but method is difficult to apply as supply voltage becomes comparable with diode p.d.s. Germanium or Schottky-barrier diodes give smaller losses, increasing efficiency at low supply voltages.
- One diode may be replaced by n-channel enhancement mode m.o.s. device (middle). For low on-resistance device, one diode-loss is eliminated. Method is not applicable to devices within normal c.m.o.s. package as these have in-built protection diodes which prevent negative output swings.
- Output currents can be increased by buffer stages such as emitter followers, but with increased voltage losses. (right)

Further reading

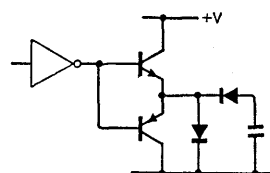
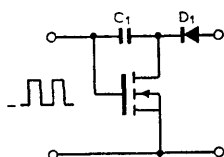
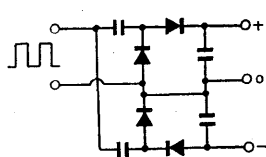
Poirier, N., Unique hybrid transformerless d.c.-to-d.c. converter suitable for biomedical applications. *ISSCC Digest* 1971, pp. 110/1.

Kauffman, R., Transformerless converter yields plus/minus voltages, in *400 Ideas for Design*, Hayden, 1971.

Cross references

Series 10, cards 2, 5 & 12.

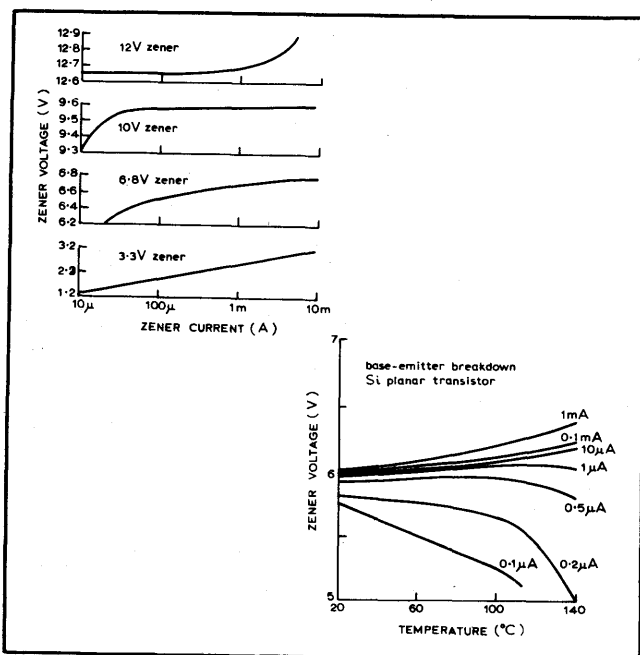
Circuit modifications



Low-current use of diodes

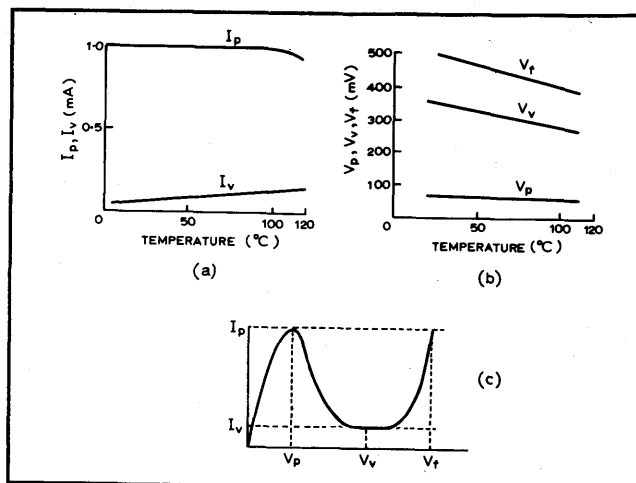
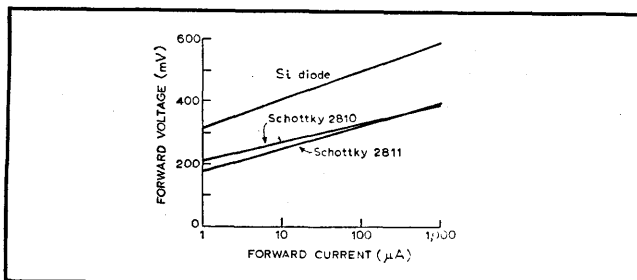
Zener diode

1. These are normally silicon diodes processed to have a reverse breakdown voltage in the range 2 to 100V (though most lie between 3V and 15V). In addition the base-emitter junction of silicon planar transistors have very sharp breakdown characteristics that make them suitable for micropower applications. Broadly the higher the breakdown the sharper the knee with a low slope resistance, together with a positive temperature coefficient of up to $+0.1\%/deg. C$. Low-voltage zeners have poor low-current stability and a temperature coefficient of $\approx -2mV/deg. C$. The second graph shows that those diodes breaking down in the 5 to 7V region may have positive or negative temperature drifts depending on current. Some planar transistors exhibit stable base-emitter breakdown characteristics at currents below $1\mu A$.



Schottky diode

2. A metal-semiconductor junction has a diode characteristic similar to that of silicon p-n junction, but for a given p.d. the current flow may be two orders of magnitude greater. While developed for their low charge-storage characteristics, this low p.d. makes them useful in low-voltage rectifiers and as temperature compensation elements. The p.d. is intermediate between Si and Ge diodes and the temperature drift comparable at $\approx -2mV/deg. C$, increasing in magnitude at lower current densities.

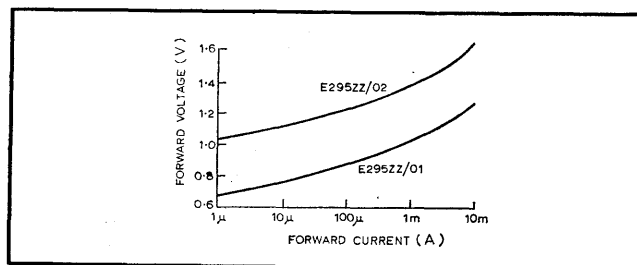


Tunnel diode

3. Tunnel diodes have a non-linear V/I characteristic with a pronounced negative resistance region. The voltage levels are low and currents may also be low. Because of their intrinsically high switching speed they are useful in applications where high-speed sensitive triggering isn't necessary. This is facilitated by the high temperature-stability possible for the peak-current, I_p (drift of $<1\%$ over a wide temperature range possible with selected devices). The ratio I_p/I_v worsens with temperature rise, reducing the output swing available when used as oscillator. Devices have been produced with I_p values below 1mA to above 100mA. Now a limited production device for special applications only.

Assymmetric voltage-dependent resistor

4. Certain polycrystalline materials have non-linear V/I characteristics resulting in zener-like behaviour but with breakdown-voltage intermediate between silicon diodes and zener diodes i.e. in the region 0.6 to 1.6V depending on type and current. Temperature coefficient of the types shown is positive, but the devices are only specified for temperatures up to 40 to 50 $^{\circ}C$. Voltage stability even at currents below those for which the devices are specified is generally adequate for biasing transistors.



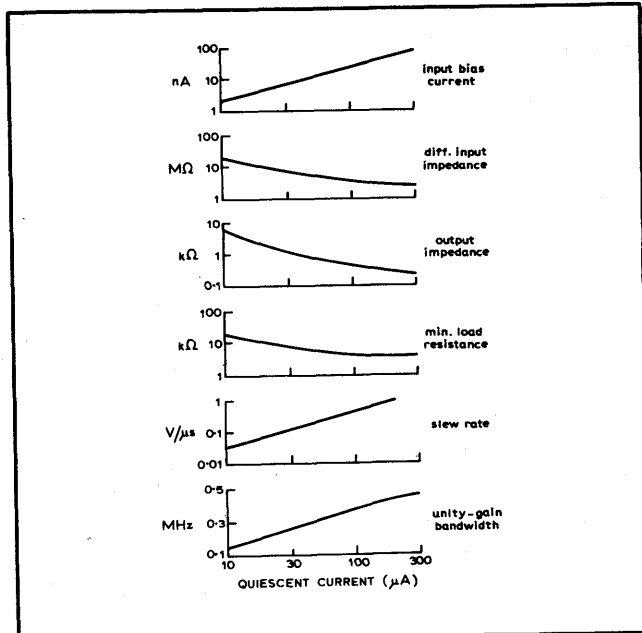
Further reading

Ferranti, Use of the base-emitter junction of the ZTX300 as a zener diode, E-Line Transistor Manual, 1969.
Preis Joachim, High-performance low-cost "active zener" regulators, *Wireless World*, vol. 1969, pp. 484-6.

Cross references

Series 10, cards 3, 5 & 12.

Micropower active devices



Operational amplifiers

It is difficult to define the operating currents of micropower op-amps by internal circuitry. It is advantageous to be able to control the currents by an external setting. Recent designs ($\mu\text{A}722$, UC4250, ICL8021C, CA3060) have internal currents controlled by a single external resistor. Quiescent current is the dominant factor in controlling many amplifier parameters; input bias currents, slew rate and output currents are broadly proportional to the quiescent current, while input and output impedances fall at higher currents though not as sharply. The minimum level of supply voltage at which satisfactory performance is sustained is determined by the amplifier configuration, and supply voltages to $\pm 1\text{V}$ or even less may be used. Output voltage swing into high resistance loads is to within 0.7V of each supply line with types listed.

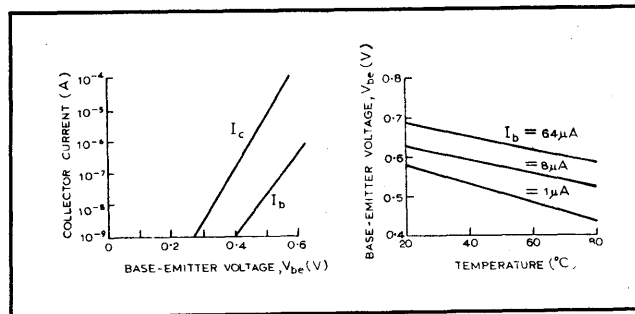
Silicon transistors

At low currents the relationship between V_{be} and I_C for fixed V_{ce} is very accurately defined for silicon planar transistors: $I_C = I_S \exp(qV_{be}/kT)$ where q is the charge on an electron, k Boltzmann's constant, and T the absolute temperature. The collector current increases tenfold for an increase in V_{be} of just short of 60mV at room temperature (doubling for collector current increases tenfold for an increase in V_{be} of

$\approx +18\text{mV}$). The temperature drift in V_{be} is around $-2\text{mV}/\text{deg. C}$ being greater in magnitude at lower values of V_{be} . Again the relationship is well defined and is given by $dV_{be}/dT = (V_{be} - 1.2)/T$. Hence transistors operating at the same V_{be} have the same temperature drift in V_{be} .

The current gain of a silicon planar transistor including those in monolithic integrated circuits is sustained to currents in the nA region though the frequency response tends to be inverse to the current because the capacitive element of the output/load is independent of current. Because the base current falls almost as rapidly as does collector current, the input resistance rises at low currents. While the current gain varies, the transconductance is defined by qI_C/kT .

The characteristics of germanium transistors are similar but the corresponding values of V_{be} are 300 to 500mV below those of silicon transistors operating at comparable currents. When the V_{be} s differ by $\approx 430\text{mV}$ the temperature drifts cancel. This is the basis of a number of low-voltage reference circuits for voltage regulators.



The collector saturation voltage of small-signal transistors at low currents varies widely but is within the range 10 to 200mV, and depends largely on the I_C/I_B ratio rather than the currents and temperature.

Further reading

Sah, Chik-Tang, Effect of surface recombination and channel on p-n junction and transistor characteristics, *IRE Trans. on Electron Devices*, 1962.

Wittlinger, H. A., Applications of the CA3080 and CA3080A high-performance operational transconductance amplifiers, RCA application note ICAN-6668.

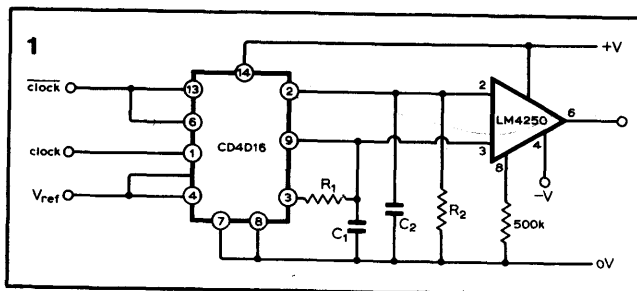
Cross references

Series 10, cards 7 & 11.

Micropower circuits

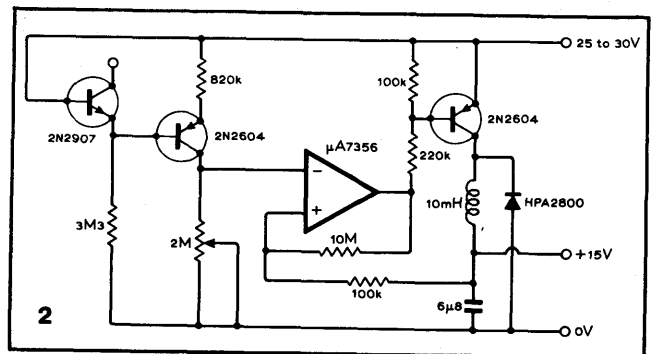
1. Bilateral switches within the CD4016 are clocked at a frequency of $1/2R_1C_1$ with $R_2C_2 = 2.2R_1C_1$. Via pins 3 and 9, C_1 swings between ground and V_{REF} , while C_2 swings between V_{REF} and $V_{REF}(1 + e)$. Output of the comparator remains low at all times. If V_{REF} increases, the positive peak of VC_1 exceeds the trough of VC_2 and the comparator changes state briefly. Micropower operation depends on the use of a suitable operational amplifier operated at the minimum current consistent with the required speed of response. For correct operation the clock frequency is precisely defined, but an alternative application is to use a constant reference voltage, in which case a pulse train is obtained when the frequency falls below a critical value. Variation of V_{REF} can then control the pulse width.

Holzschuh, J. E. Versatile circuit detects changes in voltage or frequency, *Electronics Design*, vol. 21, Nov. 22, 1973, pp.192-4.

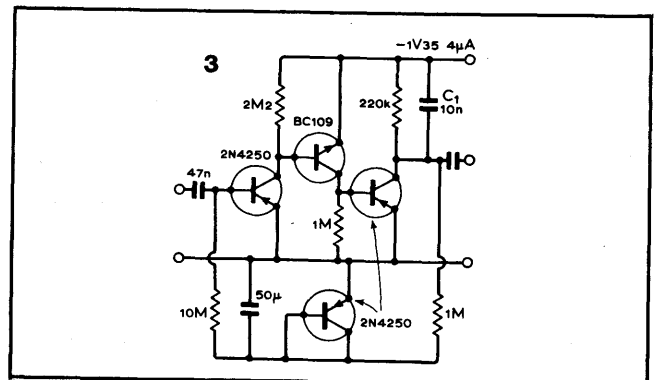


2. Switching regulators have the advantage of high efficiency even when the output voltage is much lower than the supply voltage. This, together with a low standby power, makes a micropower design of considerable interest. In form this circuit is similar to standard designs. The micropower amplifier has sufficient positive feedback to give a small hysteresis. The series transistor switches current to the inductor until the capacitor charges sufficiently to switch the amplifier, and hence the series transistor off. The inductor attempts to sustain the current during this off-period, by conducting via the diode. The left hand pair of transistors provide a constant current through the variable $2M\ \Omega$ resistor. Cater, J. P. Micropower switching regulator has a low

standby current, *Electronic Design*, vol. 20, Mar. 16, 1972 p.166.



3. The circuit is that of the amplifier for a long-life implanted transmitter for e.c.g./e.e.g. work. As such it is isolated from external wiring and requires no additional circuitry for differential measurements. The overall voltage gain is >2000 for low source impedances, with 3 to 80Hz bandwidth (this is extended to around 2kHz with capacitor C_1 removed). Input noise is $\approx 3\mu V$ pk-pk with input s/c rising to $10\mu V$ pk-pk for input o/c. The bias technique is to use an additional transistor, diode-connected to decouple the overall feedback while providing temperature compensation for input V_{be} changes. The article also describes a low-voltage complementary astable which can be modulated by the amplifier output. Horwitz, C. M. Low-drain EEG amplifier, *IEEE Trans. on Biomed. Eng.*, 1974, pp.60-1.



circuit designs



PWilliams / J Carruthers / JHEvans / JKinsler