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" × 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
Collected Circards

P Williams / J Carruthers / JHEvans / JKinsler
Paisley College of Technology, Renfrewshire

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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.

Wireless World Circuits

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.e. 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.

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Adjustable-Q twin-T notch filter  9
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Compound filters  11
N-path filter  12
Basic Active Filters

Like the doting 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-tyrator 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 $\omega^2$ as well as $\omega$. 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 $Q = \frac{f_0}{2\Delta f}$, 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 band-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 equalizing the former, or a negative-impedance converter in which an impedance 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 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
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. Gyrators 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.
Wien-bridge bandpass filter

Typical performance at 1kHz

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

\[ Q = \frac{1}{2(R_2/R_1)} \]

Voltage gain, \( \frac{V_{out}}{V_{in}} \approx -3Q \)

To find \( Q \) measure voltage gain and divide by \(-3\).

IC : 741; supplies ±15V
C = 1nF ± 1%
\( R = 150k\Omega \) ± 5%
\( (R_2 + R_1) = 10k\Omega \)
Source resistance: 600Ω
\( V_{out}(\text{max.}) = 9-20 \text{V r.m.s.} \)
Max. load current: 12mA
\( Q \) is constant for supplies between ±3V and ±18V if clipping is avoided.
\[ Z_{in} \approx 1/Q; Z_{out} \propto Q. \]

Circuit description
The Wien network—Fig. 1—gives a maximum response at
\[ f_0 = 1/(2\pi RC) \text{ Hz} \]
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 Rs or both Cs, 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 \( \pm 5\% \).
- Circuit will oscillate when \( R_2/R_1 \gg 2 \).
- 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 \( Z_2 \) and/or \( Z_3 \) and \( Z_5 \). Note that the output may not then be taken from the op-amp output.
- Since \( Z_{in} \ll 1/Q \) 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—on a cascode stage.

Further reading

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

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

$\lambda = \frac{R_1}{R_2}$

For unity voltage gain

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

IC : 741; supplies ±15V

$C = 1\mu F \pm 1\%$

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

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

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

Source resistance: 60Ω

With unity voltage gain $V_{out(max)} = 25V$ pk-pk.

No significant change in performance for supplies in the range ±3 to ±18V.

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 $R_s$ and $R_o$ 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.

Further reading

1. Williams, P. Allpass networks using Wien's bridge 
2. As 1: Alternative allpass networks using Wien's bridge, 
   p.188.

Cross references

Series 1, card 1.
**Voltage-controlled filters**

![Diagram of 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 ±6V

$VVR_3$: 1/4 CD4016AE

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

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

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

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

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

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

$V_C = 0$ to $+6 \text{V}$

$VVR_3$ variable within the range approximately 680 to 2800.

**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 $R_3$) 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 $R_i/(R_1 + VVR_3)$.

**Component changes**

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

- The greater the ratio $VVR_3/R_3$, $n = 1$ or 2, the wider the range of $f_0$ variation for given values of $C$ within the audio band.

- For a given value of $R_3$, $Q$ 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.

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**Further reading**


4. RCA Databook SSD-203, 1972, p.67.

**Cross references**

Series 1, cards 1, 6-10.
Low, high & band-pass triple amplifier

Typical performance

\[ f_0 = \frac{1}{2\pi \sqrt{C_1 C_2 R_x R_y}} \]

Drive: 400mV pk-pk.
Supply: +6V.
\[ R_x, R_y, R_z, R_3 = 10k\Omega \pm 5\% \]
\[ C_1, C_2 = 10nF \pm 1\% \]
\[ R_x = 470k\Omega \pm 5\% \]
\[ R_1 = \infty \]
\[ f_0 = 1\,583\,Hz \quad Q = 49.5 \]

Circuit description

Three inverting amplifiers form the filter. A fourth amplifier buffers the output. There are two integrators, one damped by resistor \( R_4 \), 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.e. 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 10kΩ, with \( R_4 = 50R_x \) for correct biasing. In addition, all outputs have a d.c. content.

Component changes

Supply voltage + 6V: \( f_0 = 1\,583\,Hz \), gain = 1.15, \( Q = 49.5 \).
Supply voltage + 5V: \( f_0 = 1\,562\,Hz \), gain = 0.95, \( Q = 40 \).
For best \( Q \) values, capacitors should be matched within 1%.
For low-pass and modified high-pass filters \( R_1 = 6.8k\Omega \),
\[ Q = \frac{R_1}{R_x} = 0.68 \]
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
\[ R_1 = 3.3k\Omega \pm 5\% \]
\[ C_1 = 0.1\mu F \pm 10\% \]
\[ R_3 = 330k\Omega \pm 5\% \]
\[ f_0 = \frac{1}{2\pi R_3 C_1} = 482Hz \]

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
\[ R_1 = 4700 \text{ to } 1M\Omega \]
\[ R_3 = 4700 \text{ to } \infty \]
\[ C_1 = 330pF \text{ to } 10\mu F \text{ (non-polarized)} \]
Amplifiers: N5741V, SN72741P, MC1741, etc.
Supply \( \pm 5 \text{ to } \pm 18V \) (\( \pm 3V \) in some cases).

Alternative amplifiers: any op-amp compensated for integrator operation.
\[ f_0 = < 1Hz \text{ to } > 100kHz \]
\[ Q < 1 \text{ to } > 100 \]

Circuit modifications
- Variable damping at high Q inconvenient because of large value for R3. Replace by fixed resistor fed from tapping across amplifier as shown e.g. \( R_T = 10k\Omega \) pot., \( R_3 = 100k\Omega \).
- 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

Cross references
Series 1, cards 1, 3, 6, 8, 11 & 12.
Multi-feedback filter

![Circuit Diagram]

Typical performance

For $C_1 = C_2 = C$

$$A_0 = \frac{R_2}{2R_1}$$

$$C = 0.01\mu F \pm 10\%$$

$$R_1 = R_3 = 1.5k\Omega \pm 5\%$$

$$R_2 = 470k\Omega \pm 5\%$$

Amplifier: 741

$$f_0(\text{obs}) = 845Hz$$

$$Q(\text{obs}) = 11.3$$

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$\Omega$ to 1M$\Omega$

$R_1$, $R_3$: 1k$\Omega$ to 100k$\Omega$

$C_1$, $C_2$: 150pF to 10$\mu$F

$Q = 1$ to 50 (At high $Q$ values, sensitivity to component variations is excessive)

$f_0 < 1Hz$ to $> 100kHz$

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$\Omega$ for passive network given. Alternatives include Darlington-pair amplifiers.

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

Further reading


Cross references

Series 1, cards 1, 3, 6, 7, 8, 11 & 12.
Wireless World Cocard  Series 1: Basic active filters

Adjustable-Q twin-T notch filter

![Adjustable-Q twin-T notch filter diagram]

Typical performance

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

where \( n = 2C_1/C_2 = R_1/2R_2 \) (≈ 1 usually)

\( R_1 = 324k\Omega \)
\( R_2 = 162k\Omega \) (trimmed)
\( f_1 = 33Hz \)
\( f_2 = 76Hz \)

\( C_1 = 10nF \)
\( C_2 = 20nF \) (trimmed)
\( R_1 = 16.2k\Omega \)
\( R_1 = 8.1k\Omega \) (trimmed)
\( f_1 = 680Hz \)
\( f_2 = 1650Hz \)
\( f_0 = 50Hz \) with
\( C_1 = 10nF, C_2 = 20nF \) (trimmed)

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 (\( k \to 1 \)) 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 \( C_1 = 270pF, R_1 = 600k\Omega, n = 1, f_0 = 1kHz \). Very large resistors (>10M\( \Omega \)) should be avoided.

![Circuit modifications diagram]

\[ f_0 = \frac{\sqrt{3}}{2\pi RC} \]

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-up. 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

2. N. B. Rowe, Designing a low frequency active notch filter, Electronic Engineering, April 1972.

Cross references

Series 1, cards 3 & 10.
Easily-tuned notch filter

Typical performance

\( Q \) of resonant branch is

\[
\frac{1}{\sqrt{R_1 + R_2} \sqrt{C_2 R_1 R_2}}
\]

\( f_0 = \frac{1}{2\pi} \sqrt{\frac{C_1 C_2 R_1 R_2}{C_1}} \)

\( f_0 = 50 \text{Hz} \) with

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

\( C_1 : 4 \text{nF} \), \( R : 100 \text{k}\Omega \)

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

\( R_1 \) : 50k\Omega to 1M\Omega

\( C_2 \) : 10\mu\text{F} to 0.1\mu\text{F}

\( C_1 \) : 40 n\text{F} to 100p\text{F}.

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. \( f_1 = 5kHz, f_2 = 1.9kHz \) and \( f_3 = 13kHz \).

(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.

\[ v_0 = (v_1 + v_2)/2, \]
\[ f_1 = 3kHz, f_2 = 0.76kHz, \]
\[ f_3 = 12kHz, \]
\[ f_4 = 2.1kHz \]
\[ f_5 = 4.5kHz. \]

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

N-path filter

Typical performance
IC₁ ¾ 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.

\[ f₀ = \text{clock frequency}/N \]

Bandwidth: \[2/NRC \text{ Hz},\]

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

Peak on graph:
460mV pk–pk.

\[ R = 1k\Omega \pm 1\% \]

\[ C = 0.1\mu F \pm 10\% \]

\[ f₀ = 14.9\text{kHz} \quad Q = 30 \]

Drive signal 600mV pk–pk.

Resonant frequency

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. \[R = 10k\Omega \pm 10\% \quad Q = 200.\]

IC₁ ¾ 7479; IC₂ ¾ 3001; IC₃ ¾ 7403.

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 \[f₀\] is cascaded with a 3-path filter with a fundamental resonant frequency \[2f₀\], a single bandpass characteristic is obtained, centred on \[2f₀\] (ref. 2).

Further reading
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-kΩ 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.


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 $\approx 31$kHz, centre-frequency gain $\approx 115$ 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.

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.c.s, 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 converters to simulate lossy indicators, 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.

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
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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 ±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 $\beta V$, by a small amount of the order $(V_+/\beta A)$, where $\beta$ 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 $\beta AV$, where $\beta AV$ is the change in the output, and provided $\beta A$ 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 $\beta$, the latter for an accurately defined product $\beta AV$ and either large $A$ or at least a defined value of $\beta$. 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.e. 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: ±5V; IC: 748
- Tr1: BC125; Tr2: BC126
- Diodes: IN914
- R1, R2, R3: 10kΩ
- With \( V_{in} = 1V \) pk-pk
- at 10kHz, \( V_{\pm\text{out adjust}} = \pm1.6 \text{ to } \pm4.4V \)
- Rise and fall times \( \approx 700\text{ns} \)
- Supply currents
  - Max \( V_{in} = \pm V \)
  - For \( 0 < V_1 < 3V \)
  - \( V_{\pm\text{out}} = V_3 + 3V_{be} \)
  - \( V_{\text{clamp}} = V_{\text{out}} - V_{be} \)
- For \( -3V < V_5 < 0 \)
  - \( V_{\text{out}} = V_3 - 3V_{be} \)
  - \( V_{\text{clamp}} = V_{\text{out}} + V_{be} \)

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 \( R_2 \) and \( R_3 \) set the positive and negative clamping levels; \( R_1 \) determines hysteresis at the clamping levels. Diodes provide base-emitter breakdown protection.

Component changes

- Useful range of supplies: ±5 to ±18V.
- Transistors: general-purpose silicon types.
- Useful range of \( R_1 \): 1 to 100kΩ.
- Useful range of \( R_2, R_3 \): 1kΩ (increases supply current drawn) to 100kΩ (produces error in \( V_3 \) and \( V_5 \) 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 \( R_2 \) and \( R_1 \).

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.
- \( R_2 \) and \( R_3 \) could be connected as shown on left allowing the base potentials of both \( Tr_1 \) and \( Tr_2 \) to be set positive or negative independently. It would then be possible for \( Tr_1 \) base to be positive and \( Tr_2 \) base to be negative, which would allow excessive conduction in \( Tr_1, Tr_2 \).
- Fig. on right shows a modification which allows the mean level of \( V_{\text{out}} \) to be set positive or negative by the potentiometer with its peak-to-peak value still determined by \( \pm3V_{be} \).

Further reading

- IC op.amp beats fets on input current, National Semiconductor application note AN-29, 1969, p.15.

Cross references

- Series 2, cards 4 & 6.
Basic Schmitt circuit

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

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

Circuit description
Emitter coupling between $T_1$ and $T_2$ introduces positive feedback causing a regenerative switching action into one of two states. When $V_{in}$ is below threshold level $V_1$, $T_1$ is non-conducting and $T_2$ conducts, the base voltage being determined by $R_1$, $R_2$ and $R_5$. The emitter potential is then well-defined. As input voltage exceeds $V_1$, $T_1$ begins to conduct, reducing its collector potential, and hence that of the base and emitter of $T_2$. This drop in potential is fed back to the emitter of $T_1$, thus further increasing the conduction of $T_1$ until $T_1$ is on and $T_2$ 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 $R_4 = 1kΩ$, $R_1 = 100Ω$, $R_2/R_3 = 0.5$, 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. $R_1$: 680Ω, $R_2$, $R_4$: 1.5kΩ, $R_3$: 3.3kΩ. Drive signal: 2V pk-pk; output swing: 5.2 to 10V. Hysteresis: 110mV.

On right: useful range of $R_2$: 10 to 500Ω. Useful range of $R_2$: 1 to 2kΩ. $R_1$: 1kΩ. Typical performance:

$V_{CC} = 5V$, $R_2 = 10Ω$, $R_1 = R_2 = 1kΩ$. 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

Further reading

Cross references
Series 2, cards 3, 7 & 8.
Complementary m.o.s. Schmitt

Typical performance
IC: CD4007AE (connected as triple inverter)
Supply: 10V
R1: 1MΩ
R2: 10MΩ
V1: 5.9V
V2: 5.1V
Output swing: 10V
Input current: ±0.5μA

Circuit description
Two c.m.o.s. inverters are cascaded with positive feedback defined by ratio R2/R1. 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 R1, R2 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).
- R2/R1 may be varied between 1 and 100. At upper end of range positive feedback may be too little to guarantee switching. At lower end hysteresis is comparable to supply voltage.
- To minimize capacitive effects/hum pickup reduce R1, R2 to ~ 10kΩ. 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 ~ 10pA). Resistor R1 may be dispensed with, the output resistance of buffer taking its place, with R2 reduced to range 1 to 30kΩ. Resulting hysteresis in range 2.5 to 0.2V.
- Use spare inverter self-biased by large resistor (~ 10MΩ) (see Fig.) to bias input terminal of first inverter via second resistor (~ 10MΩ). 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

Cross reference
Series 2, cards 1, 2 & 8.
High-power comparator/Schmitt

Performance data
IC: 741; Supplies ±5V
Tr1: BFR41, Tr2: BFR81
R1, R2: 220Ω, R3: 270Ω
R4, R5: 68Ω, 3W

Minimum \( V_{in} = 800mV \)

pk-pk,

Max. \( \pm V = \pm 7V, \)

for full output swing into

68Ω 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
current 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 milli-
amperes 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 (180Ω 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.c.s. *Control Engineering*, December 1969,
p.91.

Widlar, R. J., IC op-amp beats f.e.t.s 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  
$V_{\text{in}}$ (off) = 3.0V  
$V_{\text{out}}$ (on) = 2.5V  
$V_{\text{out}}$ (on) = 3.8V

$R_1$, $R_2$: 1kΩ  
$R_3$: 4.7kΩ

Supply current: 8mA  
(on), 1mA (off)

Circuit description

The transistors together have properties similar to those of a unijunction transistor. When $V_{\text{in}}$ is low, $Tr_1$ and $Tr_2$ do not conduct and $V_r$ is defined by $R_3$. For $V_{\text{in}} \approx V_r + 1.3V$, $Tr_1$ and $Tr_2$ begin to conduct, regenerative switching via $Tr_2$ clamping $V_{\text{out}}$ close to 0V. Reversal of switching occurs when $V_{\text{in}}$ falls. Significant current is drawn from the source unless a limiting resistor ($R_i$) is included.

Component changes

Tr$_1$: any general purpose p-n-p silicon transistor.  
Tr$_2$: any general purpose n-p-n silicon transistor.  
Maximum useful frequency $\approx$ 100kHz.  
Range of $V_r$ source resistance (seen by Tr$_4$ base): about 2.2kΩ to 33kΩ.  
$R_1$ (max) $\approx$ 3.3kΩ. For large $R_1$ 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 $Tr_1$ 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 $V_{\text{in}}$ its input voltage. If $V_{\text{in}}$ (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. $R_2$ and $R_3$ may then need to be reduced.

Further reading

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

Cross references

Series 2, cards 2 & 12.
Variable hysteresis level detector

Component changes
IC: 748 or LM301A.
For $V_{\text{ref}}$ of $-1V$ to $-14V$, $R_3 = 0$, supplies: $\pm 15V$;
  hysteresis: $180mV \pm 2\%$;
  trip level: $V_{\text{ref}} + 200mV$.
For $V_{\text{ref}}$ of $-1V$ to $-4V$, $R_3 = 0$, supplies: $\pm 5V$;
  hysteresis: $700mV \pm 5\%$;
  trip level: $V_{\text{ref}} + 100mV$.
Hysteresis: $10mV$, $R_1 = 1M\Omega$, supplies: $\pm 10\rightarrow \pm 15V$; trip level: $V_{\text{ref}} + 100mV$.
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 $R_2$ 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

Cross references
Series 2, cards 9-11.

Component data
Supplies: $\pm 15V$
$R_1$, $R_4$: 2.2k$\Omega$
$R_2$: 100k$\Omega$
$R_3$: 100M$\Omega$
$R_4$: 3.3k$\Omega$
$R_5$: 1k$\Omega$
$R_7$: 1k$\Omega$
$D_1$, $D_2$: general-purpose diodes.
$IC$: 741
$Tr$: ME4103 (in general determined by load current requirements)
$R_2 + R_3$: All resistors $\pm 5\%$.

Circuit description
$V_{\text{ref}}$ 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.
High-speed Schmitt circuit

Typical performance

Supplies: ±5V
$V_T = -1.0V; V_{in\,(Tr_{2on})}$
$T_{R1}, T_{R2}: $ BSX20
$D_1 = BZX55, C3V9$ $V_{in\,(Tr_{2off})}$
$R_1, R_2, R_3, R_4, R_5: 100\Omega$
$R_4: 3.3k\Omega; R_5: 1k\Omega$
$C_1, C_2: 0.1\mu F; C_3: 27pF$
$L_1, L_2: 0$
$V_{in\,(on)} = 0.2V$
$V_{in\,(off)} = 2.02V$
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$V_{in\,(on)} = 0.2V$
$V_{in\,(off)} = 2.02V$
$V_{in\,(on)} = 0.2V$ thresholds are negative and the hysteresis is variable but is not independent of the threshold levels.

Component changes
Useful range of $V_T$: 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.l.l-compatible with levels of 0V and +5V. The output from $T_{R2}$ may be used to feed a high-speed t.l.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.l.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
MECL Integrated Circuit Schmitt Triggers, Motorola application note AN-239.

Cross references
Series 2, card 2 & 8.
TTL Schmitt circuit

Performance data
Graph obtained with
R2: 250Ω, R1: 30Ω
Supply: 5V
ICs: 7400
Frequency 0 to 1MHz.
Threshold values and hence hysteresis may be changed slightly by varying R1 and R2.

Lower limit (1V as shown) is affected by R1 and R2.

With R1 = 0 there is no positive feedback and switching is not clean.
With R2 = ∞ upper limit is reduced from 5V.

Circuit description
Each NAND gate with one input gate disabled behaves as an inverter. The circuit with positive feedback via R1 is very similar to the basic Schmitt trigger as each gate is essentially identical. This results in the potential across R1 being constant and independent of which inverter is enabled. This results in an offset voltage compensated by R2.

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

Cross references
Series 2, cards 2, 3 & 7.
Low-voltage level sensor

**Typical performance**

Supply: -12V  
$V_{ref}$: -1V  
$V_{out}$

$R_1$: 3.3kΩ  
$R_2$: 100kΩ  
$R_3$: 470kΩ  
$R_4$: 82kΩ

Switching levels

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 $\sim 120\mu A$.
- Reference voltage -200mV to -3V.
- Load currents up to 100mA possible with no change in circuit. Replacing $T_4$ by higher rating transistor, and scaling all resistors down by factor of 5 allows load currents of up to 0.5A (BF441, BF480 etc).
- $T_1$-$T_4$ 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 $R_2$ 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 $R_3$ 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 $T_4$ may be omitted.

**Further reading**


**Cross reference**

Series 2, cards 6, 10 & 11.
Reference-controlled hysteresis circuit

Typical performance

Graphs obtained with
Supplies: ±15V
Tr: Motorola 2N4092
D: 1N914, IC: 741
R1: 5.6kΩ ±5%
R2: 27kΩ ±5%
C: 100pF

Lower threshold (l.t.):
Upper (u.t.):
Vref
Vref (R1 + R2)/R2
Hysteresis: Vref R1/R2
Max. frequency: 300Hz
Vref must remain positive.

Circuit description

With a low V_in, V_out is initially negative and the f.e.t. switch if off. V' is then given by V_in R1/(R1 + R2). 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 R1 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. R1 is chosen such that the f.e.t. on-resistance is much lower than R1 and the off-resistance is much higher than R1. Varying R1 and R2 hysteresis of 0.1Vref and 10Vref 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 ±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 R1 in this case being about 10kΩ.
- 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


Cross reference

Series 2, cards 4 & 6.
'Window' detector

Typical performance
IC: 711
Supplies +12V, -6V
$V_{strobe} = +5V; V_{bias} = -12V$
$R_1, R_3 = 5.6k\Omega$
$R_2, R_4, R_5, R_6 = 470\Omega$
$R_7, R_8 = 22k\Omega$
$V^+_{L} = 953mV; V^-_{L} = 886mV$ (i.e. $V_{OL}$ hysteresis = 67mV)

$V^+_{U} = -946mV; V^-_{U} = -879mV$ (i.e. $V_{OL}$ hysteresis = 67mV)
$V_{OU} = +4.4V; V_{DL} = -0.4V$ (inputs commoned)
Supply current: +11.5mA, -5mA
Strobe current: 76mA
Bias current: 2.2mA

Circuit description
$A_1$ is a non-inverting comparator having a positive reference level ($V_r$) set by $R_1$ and $R_2$. Amplifier $A_2$ is an inverting comparator having a negative reference ($V_{r2}$) set by $R_3$ and $R_4$. $V_{in}$ remains at a low level when $V_{r1} < V_{in} < V_{r2}$, and $A_2$ is capable of switching its output to $+V_{OU}$ when $V_{in} < V_{r1}$. As the outputs of $A_1$ and $A_2$ are common, $V_{out} = +V_{OU}$ when $V_{in} < V_{r1}$ or when $V_{in} > V_{r2}$. Hysteresis is introduced by the positive feedback on $A_1$ (by $R_7$ and $R_8$) and $A_2$ (by $R_4$ and $R_4$). See transfer characteristic above. From an output level viewpoint the circuit is t.t.l. compatible.

Component changes
Maximum useful frequency $\approx 1$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 $R_3 = 7000$
Minimum load resistance (for 10% reduction of $V_{OU}$) $\approx 6800$

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 $V_r$ and $V_{r2}$. Typically a resistance of 470Ω provides 5mA which is sufficient to illuminate the i.e.d. without excessively loading the IC.

Further reading

Cross references
Series 2, Card 6.
Complementary Schmitt

Typical performance

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{in\ (on)}$</td>
<td>$V_T + 2V_{be}$</td>
</tr>
<tr>
<td>Supply:</td>
<td>$+12\text{V}$</td>
</tr>
<tr>
<td>$R_1$, $R_3$:</td>
<td>$1k\Omega$</td>
</tr>
<tr>
<td>$R_2$:</td>
<td>$10k\Omega$</td>
</tr>
<tr>
<td>$R_4$:</td>
<td>$1000\Omega$</td>
</tr>
<tr>
<td>$C_1$:</td>
<td>$0.1\mu\text{F}$</td>
</tr>
<tr>
<td>$V_{in\ (on)}$</td>
<td>$2.16\text{V}$</td>
</tr>
<tr>
<td>$V_{in\ (off)}$</td>
<td>$1.59\text{V}$</td>
</tr>
<tr>
<td>Hysteresis</td>
<td>$0.57\text{V}$</td>
</tr>
<tr>
<td>$V_O\ (on)$</td>
<td>$1.0\text{V}$</td>
</tr>
<tr>
<td>$V_O\ (off)$</td>
<td>$1.1\text{V}$</td>
</tr>
<tr>
<td>Supply current:</td>
<td>10.5mA (on), 1mA (off)</td>
</tr>
</tbody>
</table>

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 $V_T + 2V_{be}$, 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$\Omega$ 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 $T_2$ 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 $R_3$ produced by $T_2$ base current.
- A speed-up capacitor of about 27pF across $R_2$ 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$\Omega$ resistor included in series with the 'free collector' of $T_2$ 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. $T_1$ will still saturate and $T_2$ 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


Cross references

Series 2, cards 2 & 5.
Wireless World Circuit

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.


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 Vin 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.


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 \( \approx 5 \)V 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.
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 waveforms 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 price of some small distortion of 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.e. 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 wave shapes 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-ampl 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 seesaw 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.
Basic op-amp square/triangle generator

![Basic op-amp square/triangle generator diagram]

**Data**

- IC<sub>1</sub>, IC<sub>2</sub>: 741
- Supplies: ±6 to ±15V
- V<sub>2</sub>: ±0.9V
- V<sub>1</sub>: 1 to 20V pk-pk
- Frequency: zero to 3kHz
- R<sub>1</sub>: 10kΩ; R<sub>2</sub>: 8.2kΩ
- R<sub>3</sub>: 1MΩ; R<sub>4</sub>: 220kΩ
- R<sub>5</sub>: 1.2kΩ; C: 0.1μF

**Circuit description**

When a positive voltage is applied to the input of the inverting integrator, consisting of R<sub>4</sub>, R<sub>3</sub>, C and IC<sub>2</sub>, 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<sub>4</sub> and R<sub>3</sub> the output voltage rises linearly with time. At some value of V<sub>1</sub> the negative current fed back through R<sub>2</sub> will overcome the positive current in R<sub>3</sub> and R<sub>4</sub>, and the resulting negative current in the non-inverting input of IC<sub>1</sub> initiates a negative going transition in V<sub>2</sub>. 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<sub>2</sub> back to its original positive value.

Hence V<sub>2</sub> is a square wave and V<sub>1</sub> a triangular wave. Resistor R<sub>4</sub> gives independent frequency control and R<sub>3</sub> varies the frequency and the magnitude of V<sub>1</sub>.

**Component changes**

- The low frequency of operation of this circuit is due mainly to the limited slow-rate of a 741 op-amp as the active element IC<sub>1</sub>. A 301 op-amp will permit frequencies of up to 10kHz to be achieved, the square wave degenerating visibly before the triangle.
- R<sub>1</sub> and R<sub>3</sub> limit the current drawn from IC<sub>1</sub> when R<sub>4</sub> and R<sub>5</sub> are in their minimum positions and could possibly be omitted.
- R<sub>2</sub> may be varied widely but must not be so low that IC<sub>2</sub> is heavily loaded and not so high that IC<sub>1</sub> fails to switch before IC<sub>2</sub> reaches saturation.
- C can also be changed, bearing in mind that the slope of the triangle is inversely proportional to C(R<sub>4</sub> + R<sub>5</sub>).

**Circuit modifications**

- The triangular wave can be given a d.c. offset of either polarity by applying a bias signal V<sub>b</sub> as shown left in which R<sub>6</sub> = 10kΩ. The bias can be increased to the point at which the integrator saturates without changing the state of IC<sub>1</sub>.
- A sawtooth waveform can be achieved by adding a d.c. signal V<sub>3</sub> to the integrator output, as shown middle. The magnitude of V<sub>3</sub>/R<sub>3</sub> must be less than [(V<sub>2</sub>)/(R<sub>5</sub> + R<sub>4</sub>)] otherwise the integrator output will not change direction as V<sub>2</sub> changes sign. Time t<sub>1</sub> is greater than t<sub>2</sub>/2 if V<sub>3</sub> is positive and t<sub>1</sub> is less than t<sub>2</sub>/2 if V<sub>3</sub> is negative. The ratio V<sub>3</sub>/R<sub>3</sub> must be comparable to V<sub>2</sub>/(R<sub>5</sub> + R<sub>4</sub>) if a large mark-space ratio is required. Independent frequency control through R<sub>4</sub> 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<sub>1</sub> < t<sub>2</sub> but t<sub>1</sub> > t<sub>2</sub> if the diode is reversed. As shown, the integration rate on the negative-going side of the triangle is controlled with R<sub>4</sub> + R<sub>5</sub> and on the positive-going side by R<sub>4</sub> in parallel with R<sub>3</sub>.

- The output of IC<sub>1</sub> 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<sub>1</sub> more nearly equal to t<sub>2</sub>/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**


Linear Applications Handbook, National Semiconductor application note AN31-6, 1972.

**Cross references**

1. Series 3 cards 2 & 11.
2. Series 2 cards 1 & 3.
Emitter-coupled triangular wave generator

Typical performance

\[ V_3 = V_4: \text{Antiphase triangular waveforms} \]

\[ V_{CC} = +15V \]

0.4V pk-pk on a d.c. level of 4V. As \( R_5 \) 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 \( R_3 \) and \( R_4 \) 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 \( Tr_1 \) fully conducting, \( Tr_2 \) off. The charging circuit is then as shown in the above diagram. Provided the conditions \( v < v_3 \) and \( v < v_4 \) are maintained, the capacitor charges linearly, but in any case for \( R_3 = R_4 \) any rise in \( v_3 \) must be accompanied by an identical fall in \( v_4 \), to maintain a constant total current. Hence there are two outputs \( v_3 \) and \( v_4 \) 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 \( v \approx -0.5V \) at which condition \( Tr_2 \) 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 \( \approx 1V \) pk-pk at a mean level of \( \approx 5V \), depending on the controlled supply current. Supply current is defined by the constant-current source connected between X and Y, where \( R_5 \) determines the value of I.

Component changes

Useful range of C: 1\( \mu \)F to 3.3nF

Frequency range: 0.5 to 130kHz

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

Circuit modifications

- With \( R_5 = 0 \), \( C = 0.1\mu \)F, 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 \( \approx 15V \). 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 \( C = 0.1\mu \)F, \( R_3 = 2.2k\Omega \), \( C_x = 1\mu \)F. The waveform across \( R_4 \) is doubled, the frequency change being \( < 1\% \).

- Resistors \( R_3 \) and \( R_4 \) replaced by a 10-k\Omega potentiometer \( R_7 \) as shown right. As \( R_7 \) is varied, the output triangular wave peak amplitude remains unchanged, though the slopes alter asymmetrically. Typically for \( C = 1\mu \)F, \( R_3 = 2.2k\Omega \), \( f = 0.5kHz \) when \( R_7 \) is set at mid-point. For a setting of 2:1, frequency reduces by 10\%.

Further reading


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 ≈ \( \frac{(V_{in} - 1)C_1}{C_2} \)  
No. of steps = \( \frac{\text{height}}{\text{step size}} \)  
Ramp height dependent on unjunction 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


Cross references

Series 3, cards 4 & 5.
Unijunction sawtooth generator

Data

Typical output waveforms obtained with Tr: 2N2160
R1: 100Ω ± 5% Supply range can be 4 to 20V at least.
R2: 10kΩ ± 5% 20V at least.
C: 1μF ± 5%

V: +15V

Circuit description

Circuit is used as a sawtooth \( (V_2) \) or a trigger pulse generator \( (V_1) \). Capacitor C charges through R1 until the unijunction transistor \( V_1 \) is reached and then discharges via R1 until the transistor changes state at approximately \( 0.5V_2 \) (sat). C then starts charging through R2 again. Waveform frequency \( \propto 1/R2C \). 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, \( V_1 \) 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 \( V_2 \) becomes rounded.

Emitter leakage current modifies the charging waveform and places an upper limit on the value of \( R_2 \) 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 \( R_2 \). The rise in the \( B_1 \), \( B_2 \) 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 \( R_1 \) to zero causes \( V_1 \) to become zero but has little effect on \( V_2 \). Any standard unijunction transistor may be used. Motorola 2N2664 will produce a smaller lower limit on \( V_2 \) and consequently reduced frequency for the same \( C \) & \( R \).

Circuit modifications

- Discharge time through \( R_1 \) may be greatly reduced by the modification, shown left, in which \( V_2 \) is used to short the capacitor to ground. This makes the pulses of \( V_1 \) much narrower and alters the frequency slightly.

- The unijunction transistor may be replaced by the two transistor version, shown middle, with \( R_1 \): 1000, \( T_1 \): BC126, \( T_2 \): BC125 and the potentiometer 2.2kΩ. The lower value of \( V_2 \) in this case comes much closer to zero. The potentiometer is set to the maximum value of \( V_2 \) 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 \( V_2 \) is at 0V rather than the minimum value it later achieves on the first discharge cycle. Resistors \( R_3 \) and \( R_4 \) are chosen so that the transistor conducts, charging \( C \) rapidly to this minimum value (ref 2).

- Charging resistor \( R_2 \) 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).

Further reading


Cross references

Series 3, cards 2 & 8.
Voltage-controlled square/triangle generator

while the tap on $R_1$ 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 $R_1$ 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 $R_1$ while the inverting input is still maintained at a constant negative voltage. A convenient setting, if the switch is ideal, is for $R_1$ to be centre-tapped with the non-inverting input tapped onto $R_2$ at $\frac{1}{2}V_C$. 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 $R_1$ to $V_C$.

Component changes
- Frequency is linearly related to control voltage $V_C$ up to -4V.
- Useful range of $C_1$: 100 pF to 0.1 µF.
- Positive feedback via $R_3$ must be <75% to maintain triangular shape, because of saturation of the $IC_1$ for the low supply voltage used.
- Adjustment of $R_2$ 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 1:5 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 $C = 1nF$, $n = 1$, $k = \frac{1}{2}$ and a supply of ±5V.

Triangular output is 4V pk-pk. Operation at 1Hz is easily achieved, but some readjustment of $R_3$ necessary to retain symmetry. Effect of supply voltage on frequency for the above components is also indicated.

Further reading

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: $(n + 1)k = 1 - k$

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

$\therefore$ For $n = 1$ $k = \frac{1}{2}$. 
Complementary transistor sawtooth generator

Typical performance

- $V_{CC} = +15V$
- $C = 10nF$
- $V_{out}:$ excursion is 2.8 to 7V at 1kHz
- $R_1: 2.2k\Omega$, $R_2: 22k\Omega$
- $R_3: 13k\Omega$, $R: 1M\Omega$ (pot)
- Supply current: 0.5mA

Component changes

- Minimum $V_{CC} = 4V$, oscillation ceases at 3.4V. With $C = 1nF$, $R_{min} \approx 47k\Omega$, $R_{max} \approx 2.6M\Omega$. Useful range of $C$: 47pF to 32µF (tantalum bead). Maximum useful frequency $\approx 70kHz$.
- Changing the ratio $R_1/R_2$ alters the voltage to which $C$ charges.

Circuit modifications

- A resistor may be included in $Tr_2$ 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 $Tr_1$ collector.
- Resistor $R$ may be returned to the $V_{CC}$ rail instead of $Tr_1$ collector, as shown middle. To increase $R$ above 2.6M$\Omega$, current gain of $Tr_1$ 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


Cross references

- Series 2 card 12.
- Series 3 card 9.
Wireless World Circuit

Series 3: Waveform Generators

D/A converter waveform generator

Typical performance

1C1: 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 R1 = 2R2. 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 ±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: 4 (CD4016) c.m.o.s. transmission gate. Resistors R4 and R5 are 100kΩ, R6: 84kΩ and R7: 16kΩ.

Diagram on bottom indicates the overall connection, where

Circuit modifications

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


Cross references

Series 3, cards 3, 11 & 12.
Triggered ramp/trapezium generator

Voltage mode and the output voltage remains constant at a value that may be controlled by the ratio \( R_3 / R_1 \). 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 \( R_3 \) and \( R_4 \) 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 \( T_{11} \) driven from a pulse source is one example.

**Component changes**

Maximum useful frequency \( \approx 100 \text{kHz} \).

With \( R_3 \) equal to 100\( \Omega \), variation of \( R_2 \) over the useful range 22 to 150\( \Omega \) varies \( V_{out} \), between 8.5 and 3V. Output voltage waveform becomes a ramp either when \( R_1 \) is increased to about 47\( \Omega \) or \( C_1 \) increased to about 32\( \mu \)F.

With \( C_1 \) equal to 22\( \mu \)F, max. useful \( R_1 \) is about 10\( k \Omega \) (ramp amplitude no longer defined by regulator feedback resistors \( R_2 \) and \( R_3 \)).

With \( R_1 \) equal to 1\( \Omega \), max. useful \( C_1 \) value is about 3000\( \mu \)F.

With \( R_1 \) set at 22\( \Omega \), \( V_{out} \) becomes a square wave with \( C_1 \) of 1\( \mu \)F.

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 \( R_1 : 330 \Omega \), \( C_1 : 1000 \mu \)F & \( R_3 : 100 \Omega \).

**Circuit modifications**

To give a higher output current rating and to provide a feedback (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:

- \( V^+ : +10V \), \( R_1 : 50 \Omega \), \( R_2 : 3.9k \Omega \), \( R_4 : 1k \Omega \), \( R_5 : 15 \Omega \)
- \( R_6 + R_7 : 1k \Omega \), \( C_1 : 47 \mu \)F, \( C_2 : 100 \mu \)F, \( C_3 : 1000 \mu \)F, \( T_{11} : BFR41 \), \( T_{12} : BFR81 \), \( V_p^+ : +3.6V \). Period: 45ms, pulse width: 26ms, pulse source resistance: 50\( \Omega \).

With \( R_6 + R_7 \) equal to 1k\( \Omega \), \( R_6 \) should not be greater than about 200\( \Omega \) 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 \( T_1 \) and \( V_p^+ \) amplitude.

**Further reading**


**Cross reference**

Series 3, card 4.
Stable waveform generator using single i.c.

**Typical performance**

- **IC**: NE555V (Signetics)
- **V**: +5V
- **R_A**: 1 kΩ ± 5%
- **R_B**: 100 kΩ ± 5%
- **C**: 10nF ± 5%
- **f**: 710Hz

Charge time ~ 0.69 x (R_A + R_B)C
Discharge time ~ 0.69 x R_B C
Period ~ 0.69(R_A + 2R_B)C
Duty cycle: R_B/(R_A + 2R_B)

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**Component changes**
- **V**: +4.5 to +16V
- **R_A**: 1kΩ to 1MΩ
- **R_B**: 1kΩ to 1MΩ
- **C**: 100 pF to 100μF

- 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 ± 1. 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 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 wave shape can be modified by changing the charge/discharge circuit. Consider the flip-flop in the state that leaves Tr1 non-conducting. The capacitor charges through R_A + R_B until the high-level comparator reverses the flip-flop. Transistor Tr1 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 << R_A the flyback time is very short, and sawtooth waveforms are possible. Conversely, for R_B >> R_A, the time-constant 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 ≫ 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 ≫ 100mA 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.

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**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 Tr1 (left) is set by current in Tr1, with only small influence of collector-emitter p.d. Any alternative current generator with p.d. < 1V 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 D1 to the capacitor i.e. using output current capability to reduce flyback time. Fall-time of < 1μs for C = 0.1μF is possible at V = +10V.

**Cross references**

Series 3 cards 2, 4 & 6.
Simple multi-waveform generator

Typical performance

Supply: +5V
A1, A2: LM3900
R1: 1MΩ; R2: 100kΩ
R3: 1.2MΩ; R4: 470kΩ
R5, R6, R7: 1MO pot.
R8: 22kΩ; C1, C2: 1nF

D1: PS101

With R5, R7 = 1MΩ & R8 = 0 output waveforms are typically as shown right.

Component changes

Maximum useful frequency ≈ 20kHz.
Useful C1 range ≈ 10pF to 22nF.
R8 min ≈ 100kΩ (Vout, switches to +4.2V).
With R2 = 105kΩ, Vout = 3.7V pk-pk, Vout = 3.3V pk-pk,
t1 = 28µs, t2 = 1.26ms, t3 = 800µs
R6 max ≈ 770kΩ (Vout, becomes +4.3V d.c.).
With R2 = 105kΩ, R6 ≈ 490kΩ, Vout is triangular;
R6 > 490kΩ, Vout is a trapezoid clamped at 4.3V;
R6 = 0, R6 = 433kΩ to 1MΩ, Vout is a ramp;
R6 = 0, R7 = 11 to 433kΩ, Vout is a trapezoid;
R6 = 0, R7 = 5kΩ, Vout is a square wave (anti-phase with
Vout) and may be made a trapezium or a ramp waveform by
selecting C3 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 R4 above should be reduced to 1MΩ and connected to 0V instead of the +V rail.
- For certain values of R4 and R7, Vout 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


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: \( \frac{1}{4} \times \) CD4007AE
C.m.o.s. switch: \( \frac{1}{2} \times \) CD4007AE
\( R_1 + R_2 \); \( R_3 + R_4 \): 47kΩ
\( R_5 \): 100kΩ, \( R_6 \): 1MΩ pot.
C\(_1\): 1nF

Variation of frequency as function of \( C_1 \) (\( V_C = +1V \)) and of \( V_C \) (\( C_1 = 1nF \)) shown right, with
\( R_1 = 2R_2 \), \( R_3 = 2R_4 \) and
\( R_6 = 1MΩ \).
At 10kHz, \( V_{\text{out}} \), is symmetrical 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 (\( C_1 = 1nF \); \( V_C = +1V \))
\( R_6 \) min \( \approx 60 \) kΩ (loss of triangular wave).
Increasing \( R_1/R_2 \) increases +ve slope of \( V_{\text{out}} \), without changing its -ve slope, increases \( V_{\text{out}} \), and increases mark-space ratio of \( V_{\text{out}} \) (22:1 when \( R_2 \) zero).
Decreasing \( R_1/R_2 \) reduces +ve slopes of \( V_{\text{out}} \), without affecting its -ve slope, reduced frequency (\( f_{\text{min}} \approx 600\text{Hz} \)), and reduces mark-space ratio of \( V_{\text{out}} \) (1:32 min at \( f \approx 700\text{Hz} \)).
Increasing \( R_3/R_4 \) reduces +ve and increases -ve slopes of \( V_{\text{out}} \), reduces frequency (\( f_{\text{min}} \approx 300\text{Hz} \)), and reduces mark-space ratio of \( V_{\text{out}} \) (1:60 min at \( f \approx 400\text{Hz} \)).
Reducing \( R_3/R_4 \) increases +ve and decreases -ve slopes of \( V_{\text{out}} \), reduces frequency (\( f_{\text{min}} \approx 20\text{Hz} \)), and increases mark-space ratio of \( V_{\text{out}} \) (500:1 max. at \( f \approx 100\text{Hz} \)).

Circuit modifications
- Variable voltage control input (\( V_C \)) may be derived from a potentiometer (\( R_7 \approx 5kΩ \), 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, C1 10pF, C3 150pF and R4 5kΩ.
- In place of a c.m.o.s. device, the switch may be realized by a discrete transistor e.g. BC125. Inclusion of \( R_6 \), typically 10kΩ, 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 \( V_{\text{out}} \).

Further reading

Cross references
Series 3 cards 1, 2, 5, 7 & 10.
Simple wave-shaping circuits

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.

Using an ideal switch and $R_1 = R_2$, gain is exactly inverted. Switch may be driven by the sawtooth of a square/triangle generator, and the circuit then inverts alternate ramps to give square wave of a square/triangle wave as desired. Cross reference Series 3, card 7.
Waveform generators

1. Ramp/triangle generators can be constructed using an integrator and a level-sensing circuit to determine the endpoints of the waveform. The 555 timer is ideal for such applications since it has internal comparators that provide sharp switching at well defined levels of $\frac{1}{3}$ and $\frac{2}{3}$ of the supply voltage. The main output can be used to switch an FET that discharges the integrator capacitor when the integrator output reaches $\frac{2}{3}V$. The second output of the timer is an open-collector device that can have TTL-compatible swings if its load resistor is returned to $+5V$. The circuit has good linearity up to 10kHz with a conversion ratio of 1kHz/V.

2. A means is shown to increase the frequency range of a standard square/triangle function generator by applying control voltages to the emitters of a pair of matched complementary transistors. The resulting current flow is a good approximation to an exponential function of the control voltages $\pm V_c$, which can be provided by a single control voltage and an amplifier of voltage gain $-1$. The transistors switching the control voltages are alternately saturated by the drive resistors. The frequency is an exponential function of the control voltage making it useful for plotting the frequency characteristics of other circuits. Hiscocks, P. D. Function generator mod. for wide sweep range, Wireless World, vol. 79, 1973, p.374.

3. The conventional astable multivibrator has an approximate ramp waveform at the base of each transistor only if the base return voltage is greatly in excess of the collector supply. Alternatively if the ramp amplitude is restricted better linearity results. Diode feedback across each transistor collector-base clamps the negative base swing to $\approx -0.4V$. The base forward voltage is $\approx +0.6V$ giving a 1-V ramp much less than the supply voltage. If the collector ramps are summed in a diode OR gate, a sawtooth waveform results. The amplitude is reasonably constant against supply changes, but since the collector resistors also provide the return paths for the capacitors the frequency is to a first order proportional to the supply voltage – i.e. a simple VCO. Hari, S. Minor changes convert an astable multivibrator into a sweep generator, Electronic Design, vol. 21, 1973, p.72.
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 I.e.d. Source loading is minimal with this circuit, which can operate at frequencies up to a few megahertz. (Card 6.)

Measurement of r.m.s. values is usually achieved in inexpensive equipment, such as the ubiquitous multimeter, by measuring mean or peak values and calibrating them in r.m.s. values. For sinusoidal waves, the relation \( V_{\text{rms}} = 1.11 V_{\text{mean}} \) 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).

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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 impressively 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 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
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 current 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 $T_2$ collector and $T_1$ emitter, or between $T_2$ emitter and $T_1$ 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 designs 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 (<<1Hz) an alternative method is the use of an integrator during a single complete half-cycle or complete cycle with separate measurement of the time to allow determination of the mean value of the waveform during that cycle.

The amplitude of the 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-
Basic diode rectifiers

![Diode Rectifier Diagram]

**Typical performance**

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

Circuit right:
- **C:** 56nF; diode: PS101
- Useful frequency range: up to 25MHz

**Circuit description**

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

**Component changes**

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

**Circuit modifications**

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

**Further reading**


**Cross reference**

Series 4 card 2, 5, 10
Peak/mean/r.m.s. calibrated rectifier

Components
IC₁: LM301A*
IC₂: LM302
Supply: 5V
R₁, R₃, R₅: 100kΩ
R₄: 22kΩ; R₅: 47kΩ
R₆: 6.8kΩ; R₇: 100MΩ
C₁: 68nF
D₁, D₂: 1N914
* needs 30PF compensating capacitor

Typical performance at 1kHz (all ranges)
Input res.: 100kΩ
Output res. < 50
Load current: 0-10mA
Stability: < 0.5%
(Vᵣ ± 7 to ±15V, Vin > 250mV r.m.s.)
Accuracy: ±0.5%
± 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 R₂ = R₁. 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 D₃ 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 1kΩ to 1MΩ 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 > 1pF 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 120kΩ.
• 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 C₃, C₄ in conjunction with R₁ may be necessary with some combinations of amplifiers to avoid risk of high-frequency oscillation. For R₁ = 100kΩ C₃, C₄ 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 D₄ still provides clamping to avoid saturation of IC₁ and R₆ 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

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 Vout: +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 -Vin/R causing Vout to be an inverted (positive) version of Vin. 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 -Vin. Amplifier A2 therefore receives the sum of the two input currents having values of Vin/R and -Vin/R. The resultant current at the input to A2, and in its feedback resistor, is -Vin/R which therefore makes Vout = Vin. Hence for any input signal Vout 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 slow-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 Vout (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 Vout (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 R3 and R4 is at -Vin and as A2 acts as a unity-gain inverter (non-inverting input of A2 is virtually grounded), Vout = Vin. 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 -2Vin/3. Amplifier A2 now acts as a follower with a gain of 1 + R5/(R3 + R4) = 3/2 for R3 = R4 = R5, hence Vout = -Vin. Thus Vout is a full-wave rectified version of Vin. 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 Vin 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 Vout = Vin and as R4 = 2R3, Vout = -2Vin + 3Vin = 2Vin. For negative inputs D1 is reverse biased and D2 conducts so that A1 acts as a follower with a gain of 2 making Vout = 2Vin and again as R4 = 2R3, Vout = -2Vin + 3Vin = -4Vin + 3Vin = -Vin. Hence Vout is the absolute value of Vin.

Further reading

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  
$R_3$: 39kΩ; $R_4$: 470Ω  
$R_5$: 270Ω  
Diodes: PS101  
$C_1$: 10μF (tantalum)  
$C_2$, $C_3$: 22μF (tantalum)

**Component changes**

Meters requiring different full-scale deflection currents can be accommodated by a suitable choice of $R_3$ for a given $V_{in}$. For $R_2$ greater than about 100Ω and a full-scale deflection sensitivity of around 100mV for a 1mA movement, the transfer function is defined by $R_4$ 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 $T_{R2}$ and possibly a resistor of around 100Ω in series with the source.

**Circuit modifications**

- Replacing the input stage by a long-tailed pair, as shown left, decreases the loading on the feedback resistor $R_3$ and allows the transconductance to approach closer to the ideal value $1/R_5$. 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 $R_2$, 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-derived feedback for low input impedance ensuring that the a.c. component of the collector current of $T_{R3}$ is diverted into the meter. The emitter of $T_{R4}$ 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 $T_{R3}$ 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**

Class-B economy rectifier

![Image of Class-B economy rectifier circuit](image)

**Typical performance**
- IC: 741
- Supplies: ±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 ~25mA and the current in the output stage may be ~1mA. This latter current ensures that the supply current changes little for signals up to 5-10% of maximum. The circuit has a number of advantages to offset these limitations: the input impedance is very high; the circuit is uncritical of supply voltages, lower values minimizing dissipation and consequent change in meter readings; sensitivity is easily adjusted; amplitude-frequency response up to 500kHz with suitable low-cost amplifier; output of amplifier available for oscilloscope monitoring.

**Component changes**
- Use range of R: 50 to 5000
- Use range of supplies: ±5V to ±15V
- There is no advantage to increasing negative supply if meter is in positive line and vice-versa. Use minimum supply voltages possible to minimize heating effect, which changes standing current in class AB stage. Any other feedback configuration can be used such as follower with gain, see-saw amplifier. Reduction of feedback increases sensitivity and allows reduced value of compensating capacitor with op-amps such as 301, 748. Bandwidth may be increased to 500kHz with gain of 10 using compensating capacitor of 3pF for 301 type op-amp. Bandwidth is already improved over some circuits, since small output voltage swing minimizes slew-rate limitations.

**Circuit modifications**
- The output stage of 741-type op-amps is basically a complementary pair of emitter followers. Driving current into the output uses the transistors instead as common-base stages, i.e. with unity current gain, but developing any desired unipolar output voltage across the meter (circuit left). Effective input resistance → 0; same limitations on minimum signal registered as before.
- Using an existing amplifier, add a separate emitter follower to the output, monitoring the current only in the collector of one transistor. A second transistor may be added to retain both polarities of output and present undistorted waveform to oscilloscope if required, and to maintain closed feedback loop, avoiding saturation on negative-going inputs (middle circuit).
- Alternately add clamping diode if amplifier has suitable access point, normally at base of one output transistor (right circuit). To avoid class A operation, which prevents detection of small signals in this mode of operation, while retaining linearity the novel approach to class B proposed by Blomley is indicated (see Further reading).

**Further reading**
- National Semiconductor application note AN31-11, 1972.

**Cross references**
- Circard series 4, cards 1, 2 & 8.
Potentiometric peak-sensing circuit

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

Circuit description
A comparator is a high-gain amplifier specifically designed for minimum response time. It is therefore a good choice for detecting specific amplitudes of 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, it 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 R₂: 10k to 100Ω.
- Same principle can be applied to any other comparator (e.g., 311) to give equal positive or negative output states (over, right).

Circuit modifications
To obtain a square-wave output, drive a divide-by-two t.t.l. bistable. Load the comparator output with 220Ω to ensure a sufficient current sink in the off condition. A moving-coil meter or light-emitting diode may be used to give a visual indication that the pulse has reached the reference level (circuit left).
- Introduce positive feedback to improve switching speed but at the expense of switching level, as shown centre. The diode ensures that the positive feedback only affects one of the switching levels. If it is arranged that the on-level is unaffected, the switch-on will occur at an accurately-controlled amplitude, but the circuit does not switch off for small transients or ripple on the pulse and only when the pulse has fallen by a defined amount.
- For the circuit shown right and with the reference voltage positive an input that is greater than the reference will cause the output voltage to be positive. When the input falls below the reference the output will be negative. Resistor R₃ has to be less than R₂ for small hysteresis, but R₃ 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
Low-frequency measurement of a.c. waveforms

Typical performance

<table>
<thead>
<tr>
<th>IC1, IC2: 741</th>
<th>Vin: 8V pk-pk</th>
</tr>
</thead>
<tbody>
<tr>
<td>A1 to A4: CD4011AE</td>
<td>Waveform: square, sine, triangle</td>
</tr>
<tr>
<td>N1 to N4: CD4001AE</td>
<td></td>
</tr>
<tr>
<td>G1 to G3: CD4016AE</td>
<td>Output: 1.00V, 0.72V, 0.51V respectively</td>
</tr>
<tr>
<td>C1, C2: 1nF; C3: 4.7nF</td>
<td>Stability: &lt;1% change</td>
</tr>
<tr>
<td>R1: 3.3kΩ; R2: 120kΩ</td>
<td>for V1 ± 4 to ± 7.5V</td>
</tr>
<tr>
<td>R3: 150kΩ</td>
<td>f: 0.9 Hz</td>
</tr>
</tbody>
</table>

Circuit description

Circuit is used to determine the integral of an input waveform over the duration of the first complete positive period following the receipt of a reset pulse. It does this using IC1 as a comparator and NOR gates N1, N2 as a schmitt trigger circuit to generate pulses via C1, C2 at each zero-crossing of the input. Gate G1 is opened and G2 closed, discharging C3 and leaving the initial output of IC2 at zero. The first positive-going zero-crossing acts via C2 and the RS (set-reset) flip-flop composed of NOR gates N3, N4 to open G1 and close G2. The input is then integrated by IC2 until a negative-going zero-crossing changes the state of the NAND gate RS flip-flop A1, A2. This opens G1 and no further integration takes place, while leaving G2 open so that the final integral can continue to be read. Gate G2 is used to suppress the negative-going step that may occur prior to the start of an integrating period, as when the original reset pulse is fed in during a positive period. The non-zero input current of IC2 together with leakage effects in G1, G2 cause the output voltage to drift and readings should be taken as soon as possible after the end of the positive period normally a single half-cycle of a repetitive waveform. Similar circuits can be produced using all NAND or all NOR elements but with no reduction in package count. Normal gating techniques could be used in place of G2 but in the present version it avoided the use of a further logic circuit.

Component changes

IC1: Any operational amplifier/comparator as speed is not critical—low offset voltage an advantage.
IC2: Minimum input current for lowest drift, e.g. LM308. Alternatively use drift compensation methods.
G1 to G3: Any m.o.s. gates; possibly reed switches for minimum drift at low frequencies.
A1 to A4, N1 to N4: Any RS flip-flops, though c.m.o.s. convenient as compatible with gates and op-amps, at V1 ± 5V.
R1/R2: 5/1 to 100/1; R1 > 1kΩ, R2 < 1MΩ.
R3: dependent on signal being integrated. Choose for output between 0.5 and 2.5V for given waveform—higher if higher supply available.

Circuit modifications

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

Further reading


Cross references

Series 4 Cards 2, 3 & 12.
High-current peak/mean rectifier

Typical performance

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>IC</td>
<td>741</td>
</tr>
<tr>
<td>R1</td>
<td>1kΩ</td>
</tr>
<tr>
<td>C</td>
<td>0.47μF</td>
</tr>
<tr>
<td>Vrms</td>
<td>1.44V</td>
</tr>
<tr>
<td>Vdc</td>
<td>2.0V</td>
</tr>
<tr>
<td>Supplies</td>
<td>±10V</td>
</tr>
</tbody>
</table>

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 R1 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.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1</td>
<td>100Ω</td>
</tr>
<tr>
<td>R2</td>
<td>10kΩ</td>
</tr>
<tr>
<td>R3</td>
<td>1MΩ</td>
</tr>
<tr>
<td>C</td>
<td>1nF</td>
</tr>
<tr>
<td>Vrms</td>
<td>100V</td>
</tr>
<tr>
<td>Vdc</td>
<td>2.0V</td>
</tr>
</tbody>
</table>

Raising R1 further is likely to leave TR1 conducting permanently; too low a value requires excessive signal drive.

Circuit modifications

- Replacing the capacitive load by a low-value resistance results in a precision half-wave rectifier. Values of R down to 15Ω may be used with input voltages of around 1V r.m.s.
- Higher peak currents may be used depending on transistor current gain/peak current rating. A d.c. milliammeter used instead of R reads mean current of half-wave rectified signal i.e. circuit may be used as a.c. mean-reading meter with moving-coil movements of low sensitivity (circuit left).
- For voltage gain R may be replaced by potential divider in usual way. For rectification giving negative-going outputs, an n-p-n transistor can be driven from other amplifier supply line. Voltage gain as shown is \( R_2 + R_1/R_1 \) (middle circuit). Inverting action is possible but requires diodes to maintain the output at zero for the input polarity for which an output is not intended. However full-wave rectification is possible with circuits such as that shown right where the input is applied in common to the two inputs with outputs also commoned. For positive-going input IC1 drives TR1 into conduction and any current through resistors R is absorbed. For negative-going input IC2 similarly drives TR2.

Further reading

National Semiconductor Linear Brief LB-8, 1969.

Cross references

Series 2 card 4.
Series 4 card 5 & 10.
Simple precision rectifiers

Typical performance
IC: 741
Diode: PS101
Supplies: ± 15V
R3: 10kΩ; R2: 3.3kΩ
R1: 6.8kΩ
Signal level: 5V pk-pk
Amplitude response: see graphs

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

Useful value of R3/R1:
\[ V_{out} = \frac{V_{in} \cdot R_3}{R_1 + R_2 + R_3} = \frac{V_{in} \cdot R_2}{R_1} = \frac{R_1}{R_2 (R_1 + R_2 + R_3)} \]
Let \( R_1 + R_2 = R' \)
\[ R_2 = R_1 \cdot R_3 \quad \text{If } R_3 = R' \]
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, \( D_2 \) 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 \( R_3: R_1 \). One ratio of \( R_3: R_1 \) to ensure that the alternative positive half-cycles are equal is deduced in the analysis shown. Note that the effect of the diode forward voltage drop is minimized as it is within the feedback loop. As the amplifier supplies the output only during one half-cycle, the amplitude response for this condition and that when only the resistors are in circuit must be different. Further, no mechanism is shown for limiting amplifier saturation for negative outputs i.e. the recovery time from this saturated condition is long, and includes the slew-rate limitation, which may be well below 1V/μs.

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

Circuit modifications
- Reverse diode \( D_3 \) to obtain negative voltages.
- Use 301 with feedforward compensation capacitors to improve response; \( C_1: 150\mu F, C_2: 1500\mu F \).
- 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 0mV above zero level.
- Variation of load is possible independently of \( R_2: R_3 \) using the circuit shown right. \( T_1: BC125, R_1: 10kΩ, R_2: 9.8kΩ \). Useful range of \( R_3 \): 3 to 10kΩ, \( R_4: 6.8kΩ \). When input is positive diode \( D_1 \) conducts, IC acts as amplifier with unity gain. When input is negative, \( D_2 \) is reverse biased, \( T_1 \) conducts and signal is applied across the load. Trimming of \( R_2 \) necessary to equalize the peaks of \( V_{out} \).

Further reading

Cross reference
Series 4, cards 2 & 3.
Positive/negative peak detector

Typical performance
IC1: LM311
IC2: LM310
Supplies: ±15V
R1: 2.2kΩ
R2: 1MΩ
C: 1μF (35-V tantalum)
R4: 10kΩ
Ripple <1% down to 200Hz
Peak detection for inputs <100mV to >20V pk-pk.
For supplies of ±10V, max $V_{in}$ for accurate
peak detection reduced to 16V pk-pk.
Max frequency >500kHz

Circuit description
The peak detector shown is based on a particular comparator
though the method is similar to that described for other peak
detectors. The difference lies in the output stage of this
comparator which may be considered as equivalent to a
switch controlled by the relative potentials of the amplifier
input, but where the switch may be effectively floated with
respect to the supply lines. While the polarity of the switch
p.d. must be defined, it allows either end of the switch to be
connected to the appropriate supply point (in the given circuit
pin 7 is taken to the positive line with pin 1 as the output;
if pin 7 is used as output, pin 1 is taken to the negative line).
This change in connection introduces an additional inversion
in the loop, equivalent to changing from common-emitter to
common-collector output configuration. This necessitates
interchange of the input to which the feedback is returned.
When the input voltage goes positive the output stage goes
into conduction supplying a large current to the capacitor C,
which charges until the potential returned to the inverting
input matches the signal. As the input signal falls, the output
stage cuts off and the capacitor C holds this peak potential
until it receives current on any succeeding positive input peak
greater than its stored potential. To ensure that the capacitor
discharges at a controlled rate and is capable of responding to
lower peak inputs within some defined time, resistor R2 is
included. The voltage follower allows normal load resis-
tances 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 4.7kΩ.
R2: 220k to 3.3MΩ; too high a value allows output stage
leakage to charge capacitor beyond range at which positive
peaks may be sensed; too low a value increases ripple at low
frequencies.
C: Determines low-frequency limits in conjunction with R2.
Use of tantalum for higher values ensures that risk of r.f.
oscillation on peaks is minimized.
Supply voltage: ±5 to ±18V.

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

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

Cross references
Series 4, cards 1, 2, 6 & 8.
Square-law meter circuit

Typical performance

- $V_s = \pm 10V$
- IC1: 741
- IC2: 741
- $T_{r1}, T_{r2}, T_{r3}$: matched n-p-n transistors, CA3046 (RCA) using three out of the five transistors in package
- $R_1: 10k\Omega$, $R_2: 220\Omega$
- $R_3, R_4: 1000\Omega$
- $R_5: 220\Omega$, $R_6: 10k\Omega$

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 $T_{r1}, IC2$ and $T_{r3}$ comprise amplifiers whose outputs are proportional to the logarithm of the input current. As these currents can be made proportional to voltage sources $V_1$ and $V_2$, the p.d. applied between base and emitter of $T_{r2}$ is of the form $\log A - \log B$ where $A$ and $B$ depend on $V_1, V_2$ respectively. Collector current of $T_{r2}$ is then proportional to the antilog of the p.d. between its base and emitter (see Theory). By tapping the base of $T_{r1}$ onto $R_3$ the effective output voltage from this first circuit can be made any desired multiple of $T_{r1}$ base-emitter p.d. provided that the base current is much less than the current in $R_3$. The output voltage fed to the emitter of $T_{r2}$ then becomes proportional to $2\log V_1$ or $\log V_2^2$ when $R_3$ tap is set to its centre value. The second logarithmic stage is required for temperature compensation even where $V_2$ is made a constant reference voltage. All the transistors should be well matched and operate at equal junction temperatures. Circuit is designed to use n-p-n types throughout and these are available in a standard low-cost multi-transistor package. As shown, the circuit deals only with positive-going voltages for $V_1$ and $V_2$, but a modification is shown that extends the operation to bipolar form. The second amplifier may also be used to provide power-law action such that the output current becomes proportional to $V_1^n / V_2^m$ with the restriction $(n - m) = 1$ for temperature-compensated operation.

Component changes

- 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.
- $R_2$: may be omitted if other means used to obtain oscillation (they reduce loop gain in conjunction with $R_3, R_4$).
- $R_5$: May be omitted subject to precautions or may be tapped as with IC1.
- $R_3, R_4$: 1k to 100k $\Omega$, setting sensitivity of circuit. At both high ($\leq 1mA$) and low ($\geq 1mA$) 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 $R_4$, the output of the amplifier is some multiple of the transistor $V_2$, i.e. is proportional to a multiple of $\log V_2$. This results in $T_{r2}$ receiving a base-emitter p.d. depending on different power laws of $V_1$ and $V_2$: $IC2 \propto \frac{V_1^n}{V_2^m}$. For the single junction of $T_{r2}$ 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 control-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 equally feedback, the output is proportional to square-root of input. Overall function performed is thus r.m.s. value of input.

Further reading

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

Typical performance
Supplies: ±15V, ±2mA
A1: 741
Diodes: CA3019 (part)
R1: 20kΩ, ±5%
R2, R3, R4, R5:
10kΩ ±5%
Source res: 60Ω
Error in mean V_{in}:
≈ +3.7%

Circuit description
This circuit is basically an inverting amplifier having its gain defined by R5/R1. All of the amplifier's output current flows in the bridge rectifier and then divides between R2 and R4. For negative input voltages the output of the operational amplifier goes positive, producing a current in R7 from node B to A via diodes D4 and D1. For positive values of V_{in} the amplifier's output swings negative bringing diodes D2 and D3 into conduction, with D1 and D4 reverse biased, again producing a unidirectional current in R7 from B to A. The p.d. across R1 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 R1 to read r.m.s. values directly. Full-scale current will be determined by R3 in parallel with R4. By making R5 about ten times R4 the full-scale current can be set by R4, which allows the r.m.s. scaling factor of the movement to be determined largely by R2 for a given value of R1. 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 R1 is sufficiently smoothed. This can be achieved by the use of a sufficiently large capacitor across R1, or by replacing R1 with a circuit of the form shown over (left). In either case, the value of R4 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, R1 and hence R2 and R3 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 × PS101 silicon diodes typically produces the response shown in curve 2: low-frequency error in V_{out} (mean) ≈ +3.7%. Using the PS101 diodes with the 741 operational amplifier replaced by a 301 with a 33-pF compensation capacitor typically produces the response shown in curve 3: low-frequency error in V_{out} (mean) ≈ +3.7%. Useful range supplies: ±3 to ±18V. Useful range of V_{in}: 350mV to 4.2V r.m.s. R4 min for V_{in} in min and no significant peak clipping ≈ 15Ω.

Circuit modifications
- To measure alternating voltages on a differential-input digital voltmeter the resistor R4 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, R4: 1MΩ, R7, R8: 10kΩ C1: 4.7μF, C2: 22μF, A1: 741, D1 to D2: 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 R4 = (2V_{in})/π.
- 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

Cross references
Series 4, cards 3 & 7.
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 mains supply being monitored, is applied via a 1-kΩ 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 uni-junction 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 uni-junction 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 uni-junction transistor.


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.


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-MΩ 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.

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.e. 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.e. audio circuits (using LM 3900) 10
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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 < 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, 100μV 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 1MΩ, possibly up to 10MΩ 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 employing 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 or pickup 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 BS192B accommodates the large 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

\[ T_0 = k \frac{(1+j \omega T_s)}{(1+j \omega T_r)} \]

Where \( T_r = 75, T_s = 318 \) and \( T_i = 3180\mu s \). 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 50kΩ. 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 3180μs, 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 intensities, 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 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 I). 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 (co-ordinateetetet). 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 quadraphonic operation surely an export boom must be the result.

<table>
<thead>
<tr>
<th>Frequency range (Hz)</th>
<th>Title</th>
</tr>
</thead>
<tbody>
<tr>
<td>10 - 30</td>
<td>grunter</td>
</tr>
<tr>
<td>30 - 100</td>
<td>boomer</td>
</tr>
<tr>
<td>100 - 300</td>
<td>roarer</td>
</tr>
<tr>
<td>300 - 1k</td>
<td>crooner</td>
</tr>
<tr>
<td>1k - 3k</td>
<td>howler</td>
</tr>
<tr>
<td>3k - 10k</td>
<td>screamer</td>
</tr>
<tr>
<td>10k - 30k</td>
<td>screecher</td>
</tr>
</tbody>
</table>
Magnetic cartridges/R.I.A.A. equalization

**Typical performance**

- **IC:** 741
- **R₃:** 680kΩ
- **R₄:** 76kΩ (82kΩ/1MΩ)
- **R₅:** 15kΩ
- **R₆:** 68kΩ
- **C₃:** 0.33μF
- **C₄:** 4.7nF
- **C₅:** 4.7nF
- **Supply:** ±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

\[
T = K \frac{(1 + j\omega T₅)}{(1 + j\omega T₄)}
\]

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 \( (R_4 + R_6)/R_6 \) provided the input time constant is low enough. As frequency rises \( X_{29} \) becomes comparable with \( R_4 \) and the gain falls, until \( X_{29} \) becomes comparable to \( R_6 \). 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 \( R_4 \) 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 \( R_5, C_5 \). To obtain high-accuracy capacitors at low cost, the time constants may be obtained using resistors for \( R_6, R_5 \) that result in a d.c. offset due to input current that conveniently cancels that due to \( R_6 \). 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.
- **R₃:** 6.8 to 680kΩ
- **R₄:** 680Ω to 100kΩ
- **C₄:** 470Ω to 47kΩ
- **C₅:** 0.1 to 10μF (may be dispensed if head allowed to carry small input current).
- **C₆:** 1nF to 1μF
- **C₇:** 1nF to 1μF

Constraints:
- \( R_1C_1 = 75\mu s, R_2C_2 = 3180\mu s \) and
- \( R_2R_3C_2/(R_3 + R_6) = 318\mu s \).

Within these constraints the input resistance should be around 50kΩ 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 \( > R_1 \) if the 75-μs 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.
Wireless World Circuit

Series 5: Audio circuits

Tone control circuits/Baxandall

![Circuit diagram]

Typical performance
IC: 741
Supplies: ±15V
R₁, R₂: 4.7kΩ ±5%
R₃, R₄: 100kΩ
R₅: 39kΩ
R₆: 5.6kΩ
C₁: 47nF ±5%
C₃, C₅: 2.2nF ±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 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 R₄ and R₅ may be increased to double their present value, or reduced to zero, e.g. for R₄ zero bass boost and cut increased by approximately 3dB for frequencies below 1kHz; for R₅ 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 = (R₃ + R₆)/R₅. Useful values for a gain of 10: R₅: 47Ω, R₆: 4.7kΩ. 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 R₁ and R₄ 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.

Further reading

Cross references
Series 5, card 6.
Rumble filters

Typical performance
IC: 741
Supplies: ±15V
C1, C2: 100nF ±10%
R1, R2: 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-audible-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 produce intermodulation distortion and even audible tones by mixing with unwanted signal components. These rumble signals may be produced from many sources such as turntable units, tape decks and microphones, all of which are driven by frequencies outside the audible range. 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 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 cores 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 R1 and R4, which could be in the form of a ganged potentiometer.

Component changes
Useful range of supplies: ±3 to ±18V. For a required cut-off frequency C1 and C3 could be made 1% components and R3 and R4 matched to give the same time constants. Ratio C1/C3 can be changed to alter response, Operational amplifier 741 can be changed to a 748 or 301 with a 30-pF compensation capacitor. Resistors R1 and R4 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 C9, R9, R10 and A9. Components C9 and R9 form an additional first-order section and R4 is connected between the output of this section and the original second-order section. Amplifier A9 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 R9 = R4, Vout is in the form of a first-order high-pass filter response. When R9 = 0, Vout follows the signal at the output of A9, which is in the form of a third-order response as the passive first-order section C9, R9 is then in cascade with the active second-order section; R4 is then, ideally simply a high resistance shunt path. As R9 changes from R4 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.

- Circuit right shows an alternative high-pass section having a Bessel response and a cut-off frequency of 0.7/2πC1R9 when C9 = C4 = C5 and R9 = 3R4.

Further reading

Cross references
Series 5, card 6.
Series 1, cards 3, 5 & 6.
Tape-head preamplifier

Typical performance

<table>
<thead>
<tr>
<th>IC</th>
<th>741*</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supplies</td>
<td>±5 to ±15V</td>
</tr>
<tr>
<td>$R_1$</td>
<td>220Ω; $R_4$: 10kΩ</td>
</tr>
<tr>
<td>(9.5kΩ preferred)</td>
<td></td>
</tr>
<tr>
<td>$R_5$: 47Ω</td>
<td></td>
</tr>
<tr>
<td>$C_1$: 0.33μF ±1%</td>
<td></td>
</tr>
<tr>
<td>values suitable for 9.5cm/s</td>
<td></td>
</tr>
<tr>
<td>$R_2$: 100kΩ; $R_3$: 22Ω</td>
<td></td>
</tr>
<tr>
<td>$C_2$: 47nF; $C_3$: 1μF</td>
<td></td>
</tr>
<tr>
<td>*not optimum for low noise performance</td>
<td></td>
</tr>
</tbody>
</table>

Circuit description

The circuit is closely related to that of the magnetic pickup preamplifier having a transfer function of the form

$$k(1 + jωT_1)/(1 + jωT_2).$$

Time constants are determined in accordance with appropriate standards. The circuit shown achieves two time-constants using a single capacitor with $T_1 = C_1(R_5$ in parallel with $R_4)$, $T_2 = C_2R_4$. For widely-spaced time constants $R_4 \gg R_5$ and $T_1 \approx C_1R_5$ i.e. independent control of the time-constants is possible in practice by varying $R_2$ and $R_4$ separately. At low frequencies the gain rises to a maximum of $(R_4/R_5) + 1$ 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 $C_3$ (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 $R_4$.
- Approximately 6dB of treble boost achieved when $C_2/R_4$ network inserted, the upper limitation being determined by $R_5$.

Circuit modifications

These are concerned with the higher frequency range where $XC_1 < R_2$.

- 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 $(R_2 + R_4)/R_4$.
- Variable gain possible 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.

Further reading

Linear Integrated Circuits and MOS Devices, RCA Databook series SSD-201.

Cross references

Series 5, card 1.
Audio mixer/Operational amplifier

Typical performance

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

Circuit description

This is a conventional summing amplifier using a 741 op-amp, with the gain defined by the ratio of the feedback resistor $R_b$ 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-kΩ inputs, the effective resistance to ground is 100Ω. Hence with a feedback resistor of 100kΩ, 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, $NK \times$ bandwidth $= 1$MHz. If the bandwidth is 50kHz, $NK$ must be 20, and if $N = 10$, $k$ is 2.

Component changes

Effects of parallel source resistance on magnitude and phase at 1kHz and 10kHz for nominal gain magnitudes of 100 and unit ($R_4 = \infty$) are tabulated below $R_2 = 100k\Omega$, $R_1 = 1k$ or 100kΩ.

<table>
<thead>
<tr>
<th>Nominal</th>
<th>Measured</th>
<th>Frequency (kHz)</th>
<th>$R_4$ (kΩ)</th>
<th>Phase Lag (deg)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100</td>
<td>95</td>
<td>1</td>
<td>$\infty$</td>
<td>$-180$</td>
</tr>
<tr>
<td>100</td>
<td>70</td>
<td>10</td>
<td>$\infty$</td>
<td>$-240$</td>
</tr>
<tr>
<td>100</td>
<td>90</td>
<td>1</td>
<td>1</td>
<td>$-200$</td>
</tr>
<tr>
<td>100</td>
<td>45</td>
<td>10</td>
<td>1</td>
<td>$-260$</td>
</tr>
<tr>
<td>1</td>
<td>unity</td>
<td>1</td>
<td>$\infty$</td>
<td>$-180$</td>
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<tr>
<td>1</td>
<td>within</td>
<td>10</td>
<td>$\infty$</td>
<td>$-180$</td>
</tr>
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<td>resistor</td>
<td>1</td>
<td>1</td>
<td>$-180$</td>
</tr>
<tr>
<td>1</td>
<td>tolerances</td>
<td>10</td>
<td>1</td>
<td>$-215$</td>
</tr>
</tbody>
</table>

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

Circuit modification

To mix two signals, use voltage follower as shown below. Output voltage is $V_{out} = kV_{in1} + (1 - k)V_{in2}$.

An advantage is that the degree of mixing is continuously variable, with equal mixing when $k = 0.5$. However, $R_4$ output impedance of the signal sources, and suitable drivers would be op-amps with $R > 10k\Omega$.

Further reading


Circuit modification
Multi-section tone control system

Typical performance
IC: 741
Supplies: ±15V
Ra to Re: 10k
Rc: 12k
Cl: 100nF, 50nF, 20nF
10nF, 5nF, 2nF, 1nF
Ra: 1kΩ
n: 7 stages
For Ra — Ra set to max.
output: +0dB/−3dB
from 120Hz to 15kHz.
centre frequency: 1/2π
CR.

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 100kΩ within above constraints.
C: 100 pF to 10μF chosen to give centre frequencies of 1/2πCR.
Rc: 1k to 10kΩ
Rf: 1 to 10kΩ

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 Cirdard 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

Cross references
Series 5, cards 2 & 10.
Series 1, cards 1, 3, 8 & 11.
Scratch filters

Typical performance
IC: 741
Supplies: ±15V
C1, C2: 1nF ±1%
R1, R2: 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 ±18V.
For defined cut-off frequency C1 and C2 may be changed with corresponding change in R1 and R2 to give same time constants.
Ratio C2/C1 may be varied to change response.
741 operational amplifier may be replaced by a 748 or 301 using a 30-pF compensation capacitor.
R1 and R2 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 A1, R4, R5, C4 and C5 form the basic second-order active filter and R3 and C3 form an additional first-order passive section. Potentiometer R4 is connected between the outputs of the first-order and second-order networks. When R3 = R4, Vout provides a first-order response cascaded first-order and second-order networks. R4 must be reasonably large to prevent significant changes in the time constants at the extreme values of R3 and A5 serves as a high-input-impedance buffer to avoid excessive loading of the network. Between the extreme settings of R3 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 A5, A4:741; supplies of ±15V; C5, C4:1nF;
R1, R2: 10kΩ; C5: 10nF; R3: 1kΩ and R4: 100kΩ.
- 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 R2 = R4 = R1, and C4 = 3C2 with a cut-off frequency of 1.4/2πC1R4 Hz.

Further reading

Cross References
Series 5; card 6.
Series 1; cards 3, 4, 6 & 11.
Microphone preamplifiers

Typical performance

IC: 741
Transformer: see note
R₁: 1kΩ
R₂: 100Ω
Source resistance: 200Ω
Response: 45Hz to 20kHz, ±0dB to -3dB
(Application of small bass-boost as in card 2 gave 20Hz to 20kHz ±1.5dB)
Equivalent input noise: 0.5μV 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 (<1mV and 100-500Ω). 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.c.s 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 n²R₂ where n is the step-up turns ratio and R₂ 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 10⁶Hz 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 X the microphone e.m.f. though 100 to 200 is more likely. Keep R₁, R₂ low from low-noise standpoint e.g.

R₁: 10 to 5000Ω.
R₂: 100 to 5kΩ.
R₄/₅: 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 R₄/₅ is relatively large (high closed-loop gain) then p.d. across R₄ is Vₛ of Tr₄, ≈0.6V for silicon. Convenient resistors might have R₁ = 5R₂, R₄ = 10R₂, R₃ = 10R₂, R₄/₅ = required gain, 0.6V/R₂ = standing current in Tr₄; all for Vₛ = 5 to 15V. The values may be chosen to give a standing current in Tr₄ 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

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 $10$Hz < $f$ < 1MHz, 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 ($Z_{in}$ ~ 11000Ω at $f = 10$kHz for $R = 100$kΩ). If d.c. conditions not important, common-base amplifier has low input resistance (~ 25Ω for $I_e = 1$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 $V_{in}$, $R_e$ determine quiescent conditions. Input resistance non-linear and input current should be restricted to small fraction of standing current. Input resistance $\propto 1/I_e$.

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 ~25mA. 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 (>100kΩ) 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. (>1MΩ) 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 $R = 10$kΩ, 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
Series 5, cards 1, 8 & 11 high $Z_{in}$
4 & 5 low $Z_{in}$
5 fixed $Z_{in}$
Wireless World Circard

Series 5: Audio circuits

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 $R_2 = 2R_1$, 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 the d.c. at the output. Although large capacitances are then required (>500uF) 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 tapping $R_2$ and decoupling to ground.

Circuit 2. Alternative capacitance feedback for inputs such as piezoelectric pick-ups (see card 11). Resistor $R_1$ necessary to provide bias, reduces low-frequency gain. Possibly center-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 $R_3$ 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 $\leq R_1$ 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 $V_{u}(d.c.) \approx [(R_2/R_1) + 1.06V$. 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 ~5V dissipation in the output transistors is <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 nonlinear 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 $R_7/R_1$ 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 $1/2\pi R_1 C_4 R_2 C_5$ while 8 represents a simple notch filter of the same frequency, adjusting $R_4$ to obtain the notch. $R_4$ 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: ±15V
R1, R2: 10kΩ; R3: 220kΩ; R4: 22kΩ;
C1: 22μF (tantalum),
Cx: 1.5nF (see circuit description)
Input: 100mV r.m.s.
Source resistance: 60Ω;
(see curve opposite)
Gain 1 + \frac{R_0(R_3 + 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 1MΩ. As the input impedance falls so also does the response at the bass end of the spectrum, leading to an approximation to an unequalized magnetic cartridge response. In the circuit shown, Cx represents the capacitance of a crystal or ceramic cartridge, normally in the range 400pF to 1nF. Bootstrap capacitor C1 allows the cartridge to be lightly loaded by the high input impedance at the non-inverting input of the operational amplifier. To achieve this C1 must be a high-value, high-quality capacitor such as a tantalum bead type. Then, the p.d. across R1 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 R1 and the cartridge sees it as an open circuit. At low frequencies, the bootstrapping becomes imperfect and the inductive input impedance results in the resonant peak shown above. In the circuit shown below(left) e = V(1 + jC2R4) and i = e/R1, hence Zin = V/i = R1(1 + jC2R4) which contains an equivalent inductance L = C1R1R4. This inductance resonates with Cx at f0 = 1/2 \pi C_1R_1R_4C_x (88Hz for above component values). For good transient performance f0 must be well below the audio band. See below for suitable values.

Component changes
Useful range of supplies: ±3 to ±18V.
To make resonant peak occur below 10Hz with C1=22μF use R1 = R3 = 180kΩ for Cx = 400pF and R1 = R3 = 120kΩ for Cx = 1nF.
Ratio R2/R4 may be adjusted to give desired gain.
Lower values of R3 and R4 make gain less dependent on Cx, R1 and R4.
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 f0 from 88Hz to 78Hz.

Circuit modifications
The circuit shown centre demonstrates a method of obtaining a flat amplitude response by making the time constant CxR2 and CxR4 equal, where Cx is the capacitance of the piezo-electric cartridge. With R4 = kR3 and C1 = C2/k the gain of the amplifier is given by: G = (kR3 - j\omega C2)/(R3 - j\omega C2) = k. Resistive loading of the cartridge is determined by R4 which should not be less than about 1MΩ. The d.c. negative feedback provided by R1 and R2 will not significantly affect the response of the amplifier provided that R1 and R2 are both very much greater than R4 and that C2 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 = VoOT/(1 + jωC4R2) and i = I'/R4. Hence the equivalent impedance (Zeot) of the d.c. feedback network is given by Zeot = Vout/i = R3(1 + jωC4R2). Thus, Zeot contains an equivalent inductance L' = C1R1R4. This inductance will resonate with C2 (with which it is in parallel) at a frequency f0 = 1/2\pi C_1R_1R_4C_x Hz. To demonstrate this effect, component values were chosen as: A: 741; Cx: 1.5nF; R1: 1kΩ; R4: 1.5kΩ; C2: 1nF; R5, R6: 10kΩ; C3: 22μF and the resonant peak found to occur at 108Hz compared with a predicted f0 of 107Hz.

Further reading
A preamplifier for use with crystal and ceramic pickups, Design Note 14, SGS-Fairchild, 1965.
Multi-input preamplifier

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 R11 chosen as a fixed value to simplify the switching arrangement. Components R5, R6, C4 and C5 provide an RIAA equalization when using the magnetic pickup input and R4 was chosen to load the cartridge with 47kΩ.

Circuit description
Useful range of supplies: ±3 to ±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. C4 and C5 may be included to roll off the flat response at high frequencies. The C4 to R4 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.

Typical performance

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>IC:</td>
<td>741</td>
</tr>
<tr>
<td>Supplies:</td>
<td>±15V</td>
</tr>
<tr>
<td>Transformer:</td>
<td>AKG 0204</td>
</tr>
<tr>
<td>R1, R2:</td>
<td>150kΩ; R3: 47kΩ</td>
</tr>
<tr>
<td>R4, R5:</td>
<td>70kΩ; R4: 8.2kΩ</td>
</tr>
<tr>
<td>R6, R7:</td>
<td>560kΩ; R6: 47kΩ</td>
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<tr>
<td>R8, R9:</td>
<td>120Ω; R9: 47kΩ</td>
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<tr>
<td>R10:</td>
<td>120Ω; R11: 1.2kΩ</td>
</tr>
<tr>
<td>R12:</td>
<td>820Ω</td>
</tr>
<tr>
<td>C1:</td>
<td>22μF tantalum</td>
</tr>
<tr>
<td>C2:</td>
<td>5.6nF</td>
</tr>
<tr>
<td>C3:</td>
<td>1.8μF</td>
</tr>
<tr>
<td>Sensitivities:</td>
<td></td>
</tr>
<tr>
<td>Tuner</td>
<td>For 350mV</td>
</tr>
<tr>
<td>Tape</td>
<td>output at 1kHz</td>
</tr>
<tr>
<td>Magnetic pickup</td>
<td>mic: 3.5mV</td>
</tr>
<tr>
<td>Piezo</td>
<td>piezo: 308mV</td>
</tr>
<tr>
<td>Tape</td>
<td>mag: 4.82mV</td>
</tr>
<tr>
<td>Tuner</td>
<td>tape: 310mV</td>
</tr>
<tr>
<td>Tuner</td>
<td>tuner: 141mV</td>
</tr>
</tbody>
</table>

Further reading
High Fidelity Audio Designs, Ferranti Ltd 1967.

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.


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.


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-kΩ 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 ≈ 800mW dissipation with a quiescent collector current of 125mA and requires a heat sink thermal resistance of 50deg C/W.

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.

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Constant-current Circuits

For every circuit using a constant-voltage element or a 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.
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-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.
Hybrid constant-current circuit

Typical data
Supply: ±15V
IC1: 741; Tr1: BFR41
R1: 1kΩ ±5%
R2: 680Ω ±5%
R3: 100Ω ±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

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 Vref/R1, 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
- Supply: useful range down to ±9V. Typically variations of current better than 0.05% over this range, when Vref is independent of the supply.
- If oscillation exists, connect a capacitor across R1.
- Useful range of R1: 330Ω to 3.3kΩ. At 2mA load current variations less than 0.05%.
- At 2mA, variations are less than -2% with BFR41 hFE 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 Vref 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 Vref/R1.

- 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%. R2: 1.1kΩ, R3: 100. Vref adjusted to give load current of 2mA. R3 varied from 4.7kΩ (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.
Silicon Zener Diode and Rectifier Handbook, Motorola.
**Constant-current use of voltage regulators**

Typical performance

- $V_{IN}$: ±15V
- $V_L$: 3V
- IC: LM309H
- $R_1$: 245Ω ±1%
- $R_L$: 120Ω ±5%
- $I_L$: 25mA

For 1.5V pk-pk input ripple at 100Hz, load current ripple is approx. 16μA pk-pk.

Dynamic output res: 90kΩ

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 $R_L$ 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 $R_L$ 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) ≈ 10mA: lower limit due to quiescent current at regulator terminal 3.
- $I_L$ (max) ≈ 200mA: power dissipation limitation of 2W in regulator without heat sink.

For $L_1$, values of 50, 100 and 200mA typical values of $R_L$ with $V_L = 3$V are 109, 51.35 and 25.2Ω respectively.

If regulator is placed some distance from the d.c. supply filter, a capacitor of about 0.1μF 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 $T_1$ and $T_2$ as a long-tailed pair with $T_3$ and $T_4$ 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 $R_s$. If the voltage regulator is good and $R_s$ 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 $T_3$ and $T_4$ is accessible in the voltage regulator, $R_s$ 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 $R_s$ 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.

Cross references

Series 6, cards 13, 10 & 11.
Simple current limiting circuits

Typical performance
- IC: N5741V (Signetics)
- Supply voltage: 10V
- Current: 26-30mA
- Voltage for limiting: 15-7-9V
- 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.

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:

<table>
<thead>
<tr>
<th>Inv. input</th>
<th>Non-inv. input</th>
<th>Output</th>
<th>Current (mA) at 10V</th>
</tr>
</thead>
<tbody>
<tr>
<td>+</td>
<td>-</td>
<td>+</td>
<td>30</td>
</tr>
<tr>
<td>-</td>
<td>+</td>
<td>-</td>
<td>29</td>
</tr>
<tr>
<td>+</td>
<td>-</td>
<td>o/c</td>
<td>12</td>
</tr>
<tr>
<td>+</td>
<td>+</td>
<td>o/c</td>
<td>1.4</td>
</tr>
<tr>
<td>-</td>
<td>+</td>
<td>+</td>
<td>0.9</td>
</tr>
<tr>
<td>+</td>
<td>+</td>
<td>o/c</td>
<td>0.85</td>
</tr>
</tbody>
</table>

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

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 TR2 as in a section of an i.c. shown above). Transistor TR2 is one of the output transistors and if the output current flowing through R tries to exceed the value at which the Vbe of TR3 reaches 0.5V, TR3 comes into conduction, diverting the base current supplied by TR1 and preventing further increase in output. In general, the limit current falls with increasing temperature because the Vbe of TR3 required for conduction rises, 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.

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, TR5 depletes TR3 of base-current, monitoring the p.d. across an external resistor R4. This allows boosting of the output current via external transistor TR1, 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, TR1 is a constant-current stage biased by D2 acting as a high-impedance load for the error amplifier (not shown). As the output current increases so does the p.d. across R4, bringing D1 into conduction and diverting current from TR1 i.e. limiting base current of TR3.
- 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 TR4 which senses the p.d. across R4. In this case it is the total current that is limited i.e. load current plus circuit quiescent current.

Further reading

Cross references
Series 6, cards 2, 5 & 9.
**Current mirror**

![Current mirror diagram]

**Typical performance**

Supply: +6V
Tr1, Tr2, part of CA3046
R1: 0-10kΩ decade resistor, ±0.05%
Ib: 0.5mA from commercial current generator, ±0.05%
Ic: 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 IMR1 maintained constant.

![Current mirror performance graph]

**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.

\[ I_R = I_B + I_E = I_{BE}(1 + \beta) + I_{BE} = I_{BE}\left(\frac{1}{1 + \beta} + 1\right) \]

\[ I_M = aI_{BE} = \frac{\beta I_{BE}}{1 + \beta} \]

\[ I_{BE} = I_R \cdot \frac{1 + \beta}{2 + \beta}; \quad I_M = \frac{\beta}{1 + \beta}. I_R \cdot \frac{1 + \beta}{2 + \beta} \approx I_R. \]

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_BE of Tr4 constant.**

**Circuit modifications**

- **Output impedance of the current mirror is increased by negative feedback via resistor R4 (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 Tr4 and R4 must be overcome. The resulting transfer ratio of IM/I_R can be shown to be (β + 2β)/(β + 2β + 2) indicating an improvement dependent on the β term, the (3β + 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**


**Cross references**

Series 6, cards 5 & 12.
Wireless World Circard Series 6: Constant-current circuits

Ring-of-two reference

Typical performance
Minimum terminal p.d.
\( V_{2a} + V_{21} - 0.5V \)
Constant current
\( V_{2a} - 0.5 \)
\( + \)
\( V_{21} - 0.6 \)
\( + \)
\( R_{3} \)
\( \frac{R_{3}}{R_{T1}} \)
\( R_{T1} : 2N3702 \)
\( R_{T2} : 2N3707 \)
\( R_{1}, R_{2} : 470k\Omega ; R_{3} : \infty \)
\( D_{1}, D_{2} : \) 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.

Component changes
\( T_{11}, T_{12} : \) General purpose silicon e.g. n-p-n types ME4103, 2N706, BFR41; p-n-p types 2N3702, ME4013, BFR81.
\( D_{1}, D_{2} : \) 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 \( \pm 6V \) 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.
\( R_{1}, R_{2} : 330k\Omega 1M\Omega \) At higher currents, self-heating effects vary current as terminal p.d. changes. At lower currents, low-leakage transistors used for \( T_{11}, T_{12} \). 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).
\( R_{2} : \) Typically 330k\Omega to 10M\Omega. Use highest value that ensures self-starting. 1M\Omega adequate with all except high leakage zeners.

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.ds 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 \( V_{be} \) 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 \( T_{11} \) and hence a constant p.d. across \( R_{3} \) (\( V_{z} - V_{be} \)). The resulting constant emitter current ensures that the collector current of \( T_{11} \) 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 \( R_{4} \) between the bases (or from \( T_{11} \) base to +ve line for example).

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 offshore without contributing current in one state. Identical zener diode with high resistance drive brings \( D_{2} \) into conduction-preferred method in some i.e. regulators but current in R flows in load if used as two-terminal constant-current circuit.
A.C. constant-current circuits

Typical performance

\[ T_{r1}, \ T_{r2}: \text{BC125} \]
\[ R_3: 1.5k \Omega \]
\[ BD_1: \text{A154} \]
\[ R_1: 47k \Omega; \quad R_2: 3.9k \Omega; \quad R_3: 100 \Omega; \quad R_4: 47k \Omega \]
\[ 5.8mA \pm 1\% \text{ for direct} \]
\[ \text{voltage of 6 to 18V.} \]

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 \( T_{r1} \) in parallel with a 100\( \Omega \) resistor \( R_3 \), maintaining a current through \( R_3 \) of about 6mA with the feedback loop closed via \( T_{r2} \). Although the current in \( T_{r1} \) varies when the applied voltage varies, this current is appreciably less than that in \( T_{r2} \) 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 \( T_{r2} \) 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 \( T_{r2} \) is substantially constant, and hence p.d. across \( R_5 \) must fall, and hence the current i.e. a relatively large increase in the current in \( R_3 \) (which is small) is balanced by a small decrease in the relatively large current through \( R_2 \). By suitable choice of \( R_3 \), \( R_4 \), the dynamic resistance can be controlled to be positive or negative, and with a critical value of \( R_3 \) 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 \( T_{r1} \) 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, \( D_3 \) conducts, clamping gate close to source voltage, and \( T_{r1} \) current is near maximum value and invariant 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


Cross references

Series 6, cards 9 & 12.
Switching current regulator

**Typical performance**
- IC: 301
- Tr: TTP3055
- D: 1A 25V diode
- C₁: 1nF
- C₄: 22μF 6.3V Tantalum
- C₅: 22μF 20V Tantalum
- R₁: 1kΩ; R₄: 5.6kΩ
- R₄: 470kΩ; R₅: 220Ω
- R₆: 150Ω
- L: 5mH (Ferrite core)*
- For Rₑ = 2Ω, Vₑ = −10V
- Load voltage: 12V,
- Supply current: 150mA
- Switching frequency: 4kHz

*See component changes

**Ripple voltage: 100mV**

**Stability: output change**

- < ±4% for supply 5 to 20V
- Output change

- < ±1% for load resistance
- 2 to 15Ω

**Circuit description**

The key difference between switching regulators and conventional types lies in the discontinuous operation of power stages 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 provided by the capacitor, the ripple voltage across R₄ can be small compared with its mean p.d. The circuit may be alternatively viewed as a simple stable in which the inversion due to the output transistor interchanges the functions of the op.amp. input terminals, while an LC circuit replaces the conventional CR version. Hysteresis provided by R₅ defines the pk-pk swing that will occur across R₆. The smaller this hysteresis, i.e. the larger R₅, the smaller the resulting ripple.

This brings it increased frequency of operation, as the rate-of-change of voltage is a function of L, C₄, R₄ as outlined above. Mean level across R₅ is fixed by that across R₄, 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 R₆. 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 200μH to 10mH.
- IC₁: Uncompensated op.amp. 748, etc. Possibility of 741,301 compensated amplifiers at low frequency with suitable choice of ferrite.
- Tr₅: For currents < 500mA: BFR41, BFY50 with reduced efficiency; somewhat higher frequencies at moderate currents: MJE521.
- R₃, R₄, R₅: Set reference voltage/hysteresis. R₃, R₄ 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 Vₑₑ 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.

**Cross references**

- Series 6, cards 8 & 10.
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, Tr2: BC 125
Tr3, BC 126
Tr4: 50V 1A (mean d.c.)
thyristor (2N1595 etc)
R1: 10Ω; R2, R3, R4: 10kΩ
R5: 25kΩ; R6: 470Ω
R7: 500Ω; 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 ±25%
≈ ±4%
Change in current for
terminal p.d. changed by
±2V ≈ ±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,
R3). Any increase in the mean current causes the mean p.d.
across R4 to increase and via R2, smoothed by C1, bring
Tr1 into conduction. This bypasses some charging current
from C3 delaying the onset of firing of the unijunction
equivalent composed of Tr2, Tr3, R4, R5 (see Series 3, card 4)
The minimum p.d. wasted across current-sensing resistor R1
need only be ≈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 R2, Tr4 would define control poin
tmore 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/R1. Setting RV1 to 50%
doubles mean current, and p.d. across R4, quadrupling power
in R4.
C1: Smooths bias to Tr1, 50 to 100μF low-voltage electrolytic.
R3: Increased value allows lower C1 for given smoothing but
decreases accuracy of current. Typical range: 2.2 to 47kΩ.
R5, C2: To give free running frequency ≈ 100Hz so that
firing can occur early in each cycle. R7: 47 to 470kΩ; C2: 10 to
100nF.
Tr1, Tr2, R4: Can be replaced by single unijunction
transistor e.g. 2N2646, 2N2160, etc. Any other general-purpose
silicon transistors in place of Tr1, Tr2.
R2, R3: Reduce R2 to 100 for some unjunctions. R3 not
critical.
Thyristor: Any medium sensitivity, low-voltage thyristor. For
higher peak currents reduce R3, R4 proportionately.
Resistor R4 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 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 R4 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 R4, R5, delaying and eventually preventing firing. Addition
of a series resistor R4 to the junction of Tr1 base/Tr2 collector
prevents excessive current flow via Tr1, Tr2.
- Alternative coupling methods including pulse transforma-
tion, 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
15/6.
400V constant-current source, Electronic Circuit Design
Nowicki, J. E., Power Supplies for Electronic Equipment,

Cross references

Low-voltage current regulators

Typical performance

\[ D_1: 1S130 \]
\[ D_2: 1S130 \]
\[ TR_1: 2N1308 \]
\[ TR_2: 2N4404 \]
\[ R_1, R_2: 220\Omega \]
\[ R_3, R_4: \infty \] Typically leakage current of Ge transistors sufficient for self-starting. To increase dynamic resistance \( R_4 \) may be in range 100\( R_1 \) to 1000\( R_1 \).

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%\( \text{deg} \) C is normally negligible. Leakage currents of the Ge transistors are enough to ensure start-up in most cases and \( R_2 \) may be dispensed with. Resistor \( R_4 \) may be added to neutralize the effect of \( R_1 \) 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 \( R_4 \) 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

\[ D_1, D_2: \text{Any silicon p-n junction including diodes (1N914, etc.) base-emitter junction of transistors (2N3707, BC125, BC126, ME4103, ZTX300, etc.) diode-connected transistor i.e. collector-base short.} \]
\[ TR_1: \text{n-p-n germanium transistor (OC139, 2N1302, 2N1304, 2N1306, 2N1308).} \]
\[ TR_2: \text{n-p-n germanium transistor (2N1303, -05, -07, -09, OC42, OC44) for optimum temperature performance with reasonably high gain transistors, diode/transistor combination should result in 400-450mV across emitter resistor.} \]

Circuit modifications

- Diodes may be placed in one limb of the circuit, over-compensating the temperature induced change in \( TR_1 V_{BE} \). By keeping \( R_1 \) and \( R_4 \) low, resulting decrease in the p.d. across \( R_1 \) is insufficient to compensate for the change in the \( V_{BE} \) of \( TR_2 \). Hence currents in the two limbs change in opposite senses and approximate cancellation is possible. Once this has been achieved, \( R_1, R_2 \) 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 \( TR_1 V_{BE} \) and suffers from variation of current in \( R_4 \) as supply varies in addition to temperature dependence (\( \approx 0.3\%\text{deg} C \)).
- Replacing \( R_n \) by a junction f.e.t. \( TR_1 \) improves the constancy of current against supply voltage while the introduction of \( D_1 \) 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 \( D_1 \). Resistors \( R_3, R_4 \) compensate for current variations in \( R_1 \) by causing the p.d. across \( R_4 \) to fall as the supply voltage rises. Typically \( R_1 = 10 R_n, R_3 = 100 R_n, R_4 \) is varied to optimize slope resistance, but is in the region 0.5 to 5\( R_n \).

Further reading


Cross references

Series 6, cards 3, 4, 5 & 6.
High-power current regulators

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 feedback 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 C3 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**

R4 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 Tr1. Basic voltage regulator normally has its output voltage set by connecting the tap on a potential divider to the feedback terminal. The resistance seen by this terminal should be around 2.2kΩ to minimize drift caused by the bias current at this terminal. This explains the values shown for the voltage regulator divider, and need for R4 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 R1 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.
**Constant-current applications**

The Zener diode $D_3$ sets the base potential of $T_{R_1}$ and hence the P.D. across its selected emitter resistor $R_9$ to $R_{f9}$. Current in the selected resistor is therefore defined as the current in the load or device under test. Transistors $T_{R_1}$ and $T_{R_2}$ 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 $T_{R_1}$. The base current of $T_{R_1}$ 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 $R_9$, the load current can be set to be 1mA by adjustment of $R_{f9}$. Constants currents of 100μA, 10μA and 1mA are then also defined for an accuracy, depending on the tolerances of $R_{f9}$, $R_9$ and $R_{f10}$ respectively.

**Typical performance**

Supply: +12V
$T_{R_1}$: 2N3702
$T_{R_2}$: BFY50
$D_3$: HS7062
$R_1$: 50kΩ, $R_2$: 270kΩ
$R_3$: 100kΩ, $R_4$: 56kΩ
$R_5$: 27kΩ, $R_6$: 12kΩ
$R_7$: 5.6kΩ, $R_8$: 2.7kΩ
$R_9$: 1.2kΩ, $R_{f9}$: 560Ω
$R_{f1}$: 470Ω, $R_{f2}$: 100Ω
$R_{f9}$, $R_{f10}$: 1kΩ

$\text{I}_{\text{supply}}$: 14.5 to 24.3mA. With load of 1kΩ all preset currents within +8% of nominal values and decade values, e.g. 10μA, 100μA, 1mA, 10mA within ±1% of each other. Dynamic output resistance/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 of 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 100μA ($S_1$ in position 1) to 10mA ($S_1$ in position 10), with an overall stability of less than 1% at any preset value. The accuracy of the preset currents is not as high as for preamble values, 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.

**Circuit modifications**

- Errors in the constant currents will be due to drift in the zener diode, drift in $V_{be}$ of $T_{R_1}$ 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 $T_{R_1}$ (see left), where $D_4$ could be a 5.6V zener and $D_4$ 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 $S_9$ in position 7 and a 1kΩ 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.
**Constant-current amplifiers**

Typical performance

- Supplies: ±6V
- $A_i$: $\frac{1}{3} \times$ CA3060 (regulator is part of CA3060)
- $R_1$: 53.7kΩ ±1% for i.e. load current changes
- $I_{bias} = 100\mu A$
- $R_2$: 47kΩ for $I_{bias}$

Equivalent source resistance with $I_{bias} = 100\mu A$ is approx. 264kΩ

A value of about 10% for a 1000% increase in $R_L$.

**Component Changes**

- Useful range of supply: ±2.5 to ±7V
- Maximum differential input voltage: ±5V
- Maximum d.c. input voltage: +V to –V

**Circuit modifications**

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 $A_i$ 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 $R_4$ and each of the class-A push-pull transconductance amplifiers are biased independently by a suitable resistor $R_4$. The transconductance of the amplifier is controlled by the bias current i.e. by the value of $R_4$. 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 10kΩ with an equivalent source resistance of about 200kΩ, provided $V_{in}$ is large enough.

Useful frequency range for square wave output current is typically 120kHz.

Circuit description

An amplitude-modulated constant-current source is obtained if the modulating voltage source is connected as a floating source in series with $R_4$ or as a grounded source to the bias terminal through a resistance of the order of 100kΩ. 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 $A_i$, $R_4$ and $R_6$ act as a negative impedance converter. As shown, $V_{bias}$ 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 $R_4$ has been replaced by a voltage source, $V_{in}$ has been set to zero and $R_4$ temporarily removed, for analysis. The output impedance at the load terminals is $Z_o = Z_o/R_4$ where $Z_o = \frac{V}{I}$. For simplicity, let $R_5 = R_6 = R_7 = R = R_4$ then $-A_e = 2(V + e)/R$. Hence $e = -2V/(A + 2)$ and $i = (V + A_e)/R = V/[A + 2]$. Thus $Z_o = V/I = (A + 2)/(2 - A)$ and $Z_o = Z_o(R = R(A + 2)/R_4$. 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, $A = -jA_0 f_0 f$ where $A_0$ and $f_0$ are typically 104 and 10Hz respectively. In this case $Z_o \approx -jA_0 f_0 R/A$ or $Z_o \approx -jA C$ so that $Z_o$ consists of a capacitance $C \approx 2n f_0 A_0 R$. For $R = 10kΩ$, $C \approx 64pF$. Thus, the constant load current will be 3dB down w.r.t. its low frequency value at $f = 1/2\pi CR_4 \approx A_0 f_0 R/4 R_4 \approx 250R/ R_4$ (kHz) for a 741-type operational amplifier.

Cross references Series 6, cards 4 & 6.
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.


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 \( R_G \) 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 1mA. 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.


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.

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
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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 $\bar{P}$, where $\bar{P}$ and $\bar{I}$ are the peak instantaneous values of voltage and current, and as $V = IR_L$ and $V_{rms} = \bar{V}$, alternative expressions are $V_{rms}^2R_L = \bar{V}^2R_L = \bar{P}$; $\bar{V} = \sqrt{2}R_L = \sqrt{2}V_{rms}$. In the absence of signal the current drawn from the supply is $V/2R_L$ giving a supply power $\sqrt{2}V/2R_L$. This shows an efficiency of 0.25. The quiescent power splits 50/50 between transistor and load resistance.

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 over the time, 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 in question 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 both 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 $V/2$. Hence the a.c. power in the load is $(V^2/2)/2R_L = V^2/4R_L$. 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.

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 $(V/2)^2/2$ while supply power is $V^2/2R_L$, 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 $V_{rms}/2$ for the same supply power $V_{rms}^2$, 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 $2V_{e}$ and $2I_{e}$ respectively and the slope of the line is the same as that joining the $V_{e}, I_{e}$ points on the axes. This optimum load is thus confirmed as depending on the quiescent conditions with the slope of the transistor 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 ($I_{e}$) may be calculated from the supply voltage ($V_{e}$) and the intended load resistance. Peak voltage in load is $V_{e}/2$; peak current in load is $2I_{e}$; therefore optimum load resistance is $V_{e}/4I_{e}$. Resulting mean load power is $(V_{e}/2)(2I_{e})/2 = V_{e}I_{e}/2$, 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

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**Fig. 3.** In a resistive-load class A amplifier (a) efficiency is only 25%; with a capacitively-coupled load efficiency could be 12.5% 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%.

**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 \( P_{\text{max}} = \frac{V^2}{8R_L} \). Therefore \( P_{\text{max}} = \frac{V^2}{8R_L} \).

Under these conditions, the mean current drawn from the supply is \( \frac{V^2}{2nR_L} \). 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 \( P_L = \frac{V^2}{2nR_L} \) and the corresponding efficiency is

\[
\eta = \frac{\frac{V^2}{2nR_L}}{\frac{V^2}{8R_L}} = \frac{\pi}{4}
\]

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 \( V = \frac{V_s}{\sqrt{2}} \), 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, where the parasite Q of the circuit must be high. 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 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 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-
Basic power amplifiers

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 $T_1$, 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 $R_1 + R_2 = h_{FE} V'_E / m I_C$. 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 $R_2$ equals the p.d. defined across $R_4$ minus the $V_{BE}$ of $T_1$. 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 $R_2$ 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 $R_4$ to prevent negative feedback within the required frequency range. As $R_4$ 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 $T_1$ conducts during the positive half-cycle and $T_2$ during the negative half cycle. For input voltages close to zero neither transistor conducts as each requires a finite base-emitter voltage for conduction to commence (~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 time. If the output stage is included within

Further reading
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

![Diagram of a servo amplifier circuit]

**Typical performance**

- Supplies: ±15V, 235mA
- Quiescent current: ±1.8mA
- A1: 741
- TR1: BFR81
- TRn, TR4: TIP3055
- TR3: BFR41
- Rn, R4: 15kΩ
- RS1: 47Ω; RS2, RS3: 4.7kΩ

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 Rn/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 R2. The base-emitter junctions of TR4 and TR3 would be forward-biased to a smaller extent unless Rn was greater than R4. However, including D5 and making Rn = R4 produces the desired bias with D5 providing some temperature compensation for the base-emitter voltage of TR1. 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 R1 in push-pull. Transistors TR4 and TR2 form a Darlington pair while TR2 and TR3 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 R4 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 ±18V.

Output power and efficiency fall as supply voltage is reduced: typically Pout is 0.8W and efficiency is 65% with ±6V at 1kHz. With maximum drive, Pout falls as RL increases: for supplies of ±15V, typically, Pout is 12.6W for RL = 6.8Ω and Pout = 3.8W for RL = 25Ω. Total harmonic distortion falls as drive increases: typically 0.45% for V1n = 2.8V and 5.3% for V1n = 150mV (supplies ±15V, RL: 18Ω and f = 1kHz sinewave).

**Circuit modification**

- The TR1-TR3 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 R4 and R3 and/or placing a diode in series with R4 and R3. 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**


**Cross references**

Series 7 cards 1 & 12.
Series 2 card 4.
Series 4 card 8.
**Pulse buffer amplifier**

![Circuit Diagram](image)

**Typical performance**
- $V_t = +14V; V_{in} = +5V$
- $T_{1}: TIS45; T_{2}: TIS50$
- $I_{C1}: I/6 SN7406$
- $R_1: 470; R_2: 100\Omega$
- $R_3: 10\Omega$
- $D_1: P5101; C_1: 680pF$
- Input pulse height: 4V
- Duration: 600ns
- P.R.F.: 50kHz
- Rise time: 20ns

Output pulse: rise time 49 ns; fall time 32ns; pulse height: $= V_1$

(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.

through $R_3$ is significantly greater and could cause excessive current flow in $T_{2}$ 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 $R_4$ which would give the same bias drive current in both directions.

$IC_1$ is a open-collector high-voltage output device which pulls the potential at the bases of $T_{1}$ and $T_{2}$ 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 $T_{1}$ and $T_{2}$ can be replaced by BFR41 & BFR81 or BC125 & BC126 with poorer rise and fall times.
- Typical comparison:
  - Rise time (ns): 12
  - Fall time (ns): 12
- For each capacitor value, overshoot on leading and trailing edges of output pulse is approximately 25% of pulse level.
- Resistive load: 100\Omega, $V_t = 14V, V_{in} = +5V$; output pulse excursion is from 1.6 to 12V.
- Pulse width: 6us. Useful frequency range 3 to 100kHz.
- Corresponding mean current from supply 1.5 to 30mA d.c.
- $IC_1$: SN75451 or SN7407 for greater output voltage levels and faster rise times.

**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 $T_{1}$ driving the capacitor in the positive direction when a positive-going edge is applied at the base connection, while $T_{2}$ 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 $R_3$, as $D_1$ is allowed to conduct. On the negative-going edge, current

**Further reading**
- Texas Instruments Technical Seminar 1972, m.o.s. memory drivers.

**Cross references**
- Series 6, cards 1, 2 & 8.
Push-pull class A power amplifier

![Circuit Diagram](image)

**Typical performance**

For supply of 13V, quiescent current of 950mA, max. output for 3% distortion is 12V pk-pk into 5Ω (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 3Ω @ 5% t.h.d. Distortion: <1.1%, 1W into 3Ω, 100Hz to 10kHz. Voltage gain ~2. Input impedance ~250Ω.

**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 R4 remains constant throughout the a.c. wave form while its d.c. value can be adjusted to set the desired quiescent current. Bootstrapping via C4 ensures that any increase in the collector potential of TR3 is transferred via the emitter follower action of TR2 to reappear at the junction of RE and R5. Hence the change in p.d. across R4 across zero except at very low frequencies where the reactance of C5 becomes significant. As there is no change in RE current, any increase in TR3 current increases the base current of TR4 while reducing the base current of TR5 by substantially the same amount. Accurate current phase-splitting together with matched current gains of TR3, TR5 keep the distortion low. Overall negative feedback via R5 defines the output quiescent voltage as a multiple of the base voltage of TR1 (~1.3V) and the ratio Rs/Re scales this base voltage up to half the supply voltage, i.e. the output transistors operate with equal Vces as well as equal Ie.

**Component changes**

TR5, TR3: Power transistors with closely matched hFE at operating current. Quiescent power (at least twice max. output) determines types and heat sinks.

2N3055 for Ps > 5W. MJ6521 for Ps > 1W.

BFY50, BFR41, etc., for Ps < 1W.

TR1: BFY50, BFR41, 2N3053 for most applications.

C4: Reactance < RE at lowest freq. Typically 200 to 5000μF.

C5: Reactance ≤ RE at lowest freq. Typically 100 to 500μF.

R5, R6: Set output current Vo/2/(R5 + R6) ≈ 2A/hFE. One resistor may be made variable to adjust mean current. Typical range 100Ω to 1kΩ (higher values for low-power circuits).

R5: Sets voltage gain ~ R5/RE and input resistance ~ RE.

R5, R6: Set output voltage (quiescent) to ~ 2V/Re [(R5/RE)+1].

Current in R6, R5 to 5 to 20 times base current of TR1. Typical values R6: 100 to 500Ω. R5: 300Ω to 3kΩ.

**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 Vbe rather than at the supply centre point, i.e. resistors require readjusting for different supply volts. Adding TR3 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 Rs = RE. Voltage gain ≈ (Rs/Re)+1. Reactance of C5 ≤ RE at lowest frequency of interest. Typically Rs, Res: 1 to 10kΩ, RE, Rs: 20 to 200kΩ. (circuit left)

- For higher input impedance, input potential divider may be bootstrapped. Interchanging locations of R6, C4 allows Rs to be bootstrapped, almost doubling input impedance (centre)

- Quiescent current depends on current gains of TR5, TR3.

By monitoring circuit mean current and using result to control drive current TR1, mean current can be made constant, e.g. for TR3 a germanium transistor, DC a silicon diode, mean p.d. across R5 is controlled at 0.4V.(right)

**Further reading**


High-voltage amplifier

\[ V_{CE} = 4V \]
\[ R_1: 10k\Omega \]
\[ R_2: 456\Omega \]
Quiescent voltage: 52V
Input signal: 7V pk-pk
Output voltage: 84V pk-pk
Gain constant up to 20kHz
Variation of output with \( R_1 \) not decoupled shown opposite.
Effective output impedance of transistor configuration 5M\( \Omega \).

Typical performance
Supply: \(+100V\)
\( T_{21}: MJE340 \)

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 \( T_1 \) and the high voltage characteristics by \( T_2 \). The input characteristics aimed at were that the device should behave with a defined gain, so the whole system could be considered equivalent to a valve. Transistor \( T_2 \) is thus a field-effect transistor whose gain is controlled by the quiescent current, which may be set by \( R_2 \). The drain of \( T_2 \) feeds into the emitter of \( T_1 \) whose base is maintained at a constant potential, just high enough to ensure that \( T_2 \) has a quiescent voltage that is above its pinch-off value. The bias voltage should be obtained from a low impedance circuit. Hence \( T_2 \) is operating into a low impedance, while \( T_1 \) is virtually a common-base stage and has thus the highest voltage rating that it could possibly have. The current at the collector of \( T_2 \) 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 feedback between the collector of \( T_1 \) and the gate of \( T_2 \) as the voltage swing at the collector is isolated from the gate of \( T_2 \). The capacitance between \( T_2 \) 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 \( T_1 \).

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 \( V_{BE} \) 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: \(+200V\), \( V_{Q} = 110V\), \( R_2 = 12.2k\Omega\), \( V_{IN} = 7V \) 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, \( R_F \) 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
Class C power amplifier

Typical data
Supply: 12V
$T_1$, $T_2$: BFR41
$R_1$: 100Ω; $R_2$: 50Ω
($carbon$)
$C_1$: 180pF; $C_2$: 360pF
$C_3$: 47pF; $C_4$: 10nF
$C_5$: 500pF; $C_6$: 190pF
$C_7$: 805pF; $L_1$: 2.7μH
$L_2$: 2.16μH; $L_3$: 2.38mH
$L_4$: 230μH; $L_5$: 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 $C_2$, $C_3$, $C_4$, $C_5$, $C_6$, $C_7$ were originally fixed capacitors 'padded' out with variabes. 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 $T_1$ and $T_2$ 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 $R = V_{out}/P_{out}$. 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 $C_6$ to $C_7$. Complete design procedures are given in the first three references.

Further reading
Motorola, application note AN-282: Systemizing r.f. power amplifier design, 1967.

Cross references
Circard series 7, card 1.
Bridge output amplifiers

![Bridge output amplifiers diagram](image)

**Typical data**
- $I_{C_1}$, $I_{C_2}$: 741
- $R_s$: 10kΩ pot
- $R_t$: 10kΩ
- Supplies: ±15V
- $R_l$: 2kΩ
- Output voltage: 15V
  - r.m.s. into 2kΩ (17.5V)
  - r.m.s. o(f) for $k = 0.1$ to 1.0 at 1kHz.

![Output voltage vs. Supply voltage graph](image)

**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. This 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 $R_s = R_t$. 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. Slow-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, 30t, etc.) with appropriate compensation capacitor (reduced compensation possible with increased gain leading to higher slow rate).
- Resistor values non-critical but $R_s = R_t$ gives push-pull output (circuit usable as phase-splitter for succeeding stages). Resistor $R_s$ 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 $R_s$, $R_t$: 1kΩ to 250kΩ. Higher values lead to drift and additional h.f. limitations; lower values absorb too much of the available output current.
- If unity gain is sufficient, $I_{C_1}$ may be replaced by voltage follower, $R_s$ 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**
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.
Class B quasi-complementary output

![Circuit diagram]

**Typical performance**

- Supply: +20V
- $\text{Tr}_1$: BFR81; $\text{Tr}_2$, $\text{Tr}_3$: TIP3055
- $\text{Tr}_4$: BFR41
- $R_1$, $R_4$: 1.5kΩ; $R_5$, $R_6$: 1kΩ
- $R_2$: 470Ω; $R_3$: 330Ω
- $R_4$: 1.8kΩ; $R_5$: 8.2kΩ
- $R_6$, $R_7$: 1kΩ; $R_8$: 8Ω
- $C_1$: 100μF; $C_2$: 22μF; $C_3$: 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%
---

**Circuit description**

This is a circuit of a class B push-pull amplifier in which transistors $\text{Tr}_4$ and $\text{Tr}_5$ complement the pair $\text{Tr}_3$ and $\text{Tr}_1$. To use n-p-n transistors in the output stage for economy, the configurations of the two sections are different, i.e. $\text{Tr}_3$ and $\text{Tr}_4$ are connected as a Darlington pair and $\text{Tr}_1$ and $\text{Tr}_2$ as a complementary pair. They receive essentially the same a.c. drive, but with the bases separated by $\text{Tr}_3$. $\text{Tr}_2$ and $\text{Tr}_4$ conduct for positive-going output signals and $\text{Tr}_5$ supplies base current drive to $\text{Tr}_1$ and $\text{Tr}_3$ for negative-going output signals. Transistor $\text{Tr}_6$ is used in the so-called amplified diode configuration in which the potential difference between the bases of $\text{Tr}_1$ and $\text{Tr}_3$ is set as a multiple of the $V_{be}$ of $\text{Tr}_3$ by the potential divider $R_8$, $R_9$, i.e. $R_8$ can be adjusted to give the desired quiescent current in transistors $\text{Tr}_1$ and $\text{Tr}_3$. 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 $\text{Tr}_7$ is an inverting amplifier with overall negative feedback through $R_9$, the values of $R_4$ and $R_5$ determining the d.c. output potential in conjunction with $R_1$. Because $R_6$ is decoupled, the a.c. properties of the arrangement are determined by the ratio of $R_1$ to the source resistance. Resistors $R_7$ and $R_8$ are centre-tapped and this point is taken to the output via $C_1$, which bootstraps $R_9$ 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 $\text{Tr}_1$ and $\text{Tr}_4$. 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 $\text{Tr}_1$ or $\text{Tr}_4$ to turn on and divert the base current available to $\text{Tr}_3$ or $\text{Tr}_1$, 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 $R_4$ 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.
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.
Wireless World Circard

Series 7: Power amplifiers

Broadband amplifier

![Circuit diagram of a broadband amplifier](image)

Typical performance

- Supply: ±20V, 118mA
- $V_{cc}$ (min) = ±5V, $V_{bb}$ (max) = 140mV r.m.s., supply current is 30mA, and $P_{out}$ ≈ 11mW
- $R_T$ and $R_S$ can both be BFY50 or BFR41.
- $R_T$ can carry a much smaller quiescent current, using for example an ME4103, with increased values of $R_b$, $R_4$ and $R_s$. $R_1$ can be increased or decreased to allow matching to source resistances greater or less than 50Ω respectively.

Component changes

With $V_{cc}$ (min) = ±5V, $V_{bb}$ (max) = 140mV r.m.s., supply current is 30mA, and $P_{out}$ = 11mW, $R_T$ and $R_S$ can both be BFY50 or BFR41.

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 $R_T$ may be drastically reduced. Resistors $R_4$ and $R_S$ would then need to be increased, with a corresponding increase in $R_b$, if this is to be the means of controlling the quiescent operating conditions. The lower $R_T$ current may be chosen to make the natural input resistance of the stage, in the absence of $R_4$, the value required to match the source.

- Input resistance may be defined using shunt-applied feedback, as shown left, where the emitter of $R_T$ is d.c. or a.c. grounded, the feedback is not decoupled and the voltage gain is determined by the ratio $R_3/R_4$. The input resistance is largely that of $R_3$ except at high frequencies where the feedback falls and the impedance at $R_T$ base must be considered.
- Inclusion of $R_4$, as shown right, may be applied to both the previous circuits to allow an output to be taken from the collector of $R_T$. To maximize the signal swing in the collector circuit of $R_T$ the bias network must be readjusted to leave a small voltage at $R_T$ emitter, say by reducing $R_4$ and $R_3$ in the original circuit. The output resistance is approximately $R_4$; this stage is therefore convenient for feeding directly into any other low impedance stage, such as that left, with $R_4$ removed. This mismatch can often be of advantage in extending the bandwidth of the amplifier.

Further reading


Lo, A. W. (and others), Transistor for Electronics, Chapter 9, Prentice-Hall, 1955.


Cross references

Series 7, cards 1, 4, 5 & 10.

Circuit modifications

![Circuit diagram for circuit modifications](image)
Class A op-amp power booster

Typical performance

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 Rg then as the base potential of Tr2 increases the emitter current of Tr3 decreases and with it the collector current. This fall is fed to the base of Tr5 causing it to conduct less, while the fall in emitter current releases more of the constant current in R4 to flow in the base of Tr5. Provided the current gain of Tr1 is reasonably high, the magnitudes of the base current charges in Tr5, Tr6 are equal but the signs are opposite. This represents an approach to ideal current phase-splitting. The constant current in R4 is provided by the bootstrap capacitor C5, such that any change in the potential at the base of Tr5 is coupled via the follower action to the positive end of R5, i.e. with no resulting change of p.d. across R2 in the ideal case. Resistors R3 or R4 require to be variable to set the output current and stability of that current then depends on hFE variation in Tr5, Tr6. The base-emitter p.d.s of Tr5, Tr6 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 Rg, R4 above the supply on positive signal swings. Hence it can drive Tr5 base far enough positive to saturate Tr6 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 R5 may replace R9. (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


Cross references

Series 7, cards 2, 4 & 8.
An a.c. signal can be superimposed at pin 6 via C2 and R4, 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 R4 falls and with it the current through R2. The current in R1 for this voltage exclusion can never exceed R4 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 R4 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 R4 is replaced by the Baxandall constant-current circuit shown left TR1: BFR81, TR2: TIP3055, R4: 18Ω, R6: 3.9kΩ. This permits a much greater input signal level before peak clipping occurs. Resistor R4 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 R4 is 150Ω with the same quiescent current. A comparison of instantaneous currents for the two possible circuits between a and b is tabulated 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 divider 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.

Cross references
Series 3, card 8.
Series 7, card 12.
Class D switching amplifier

Typical performance

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.

V_{bias} = -640mV to set
mean load voltage to
zero with \( V_{in} = 0 \).

Circuit description

Basicall, 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 \( A_1 \) which uses positive feedback
via \( R_2 \) and \( R_4 \). The periodic time of the squarewave fed to
\( R_4 \) depends on the time constant \( RC_1 \); if \( R_4 \) is much greater
than \( R \). 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 \( R \).
Current in \( R_4 \) flows alternately in \( R_1 \) and \( R_4 \) producing p.d.s
across these resistors that are sufficient to switch on \( T_1 \) and
\( T_2 \) respectively. The signal applied to \( R_1 \) 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 \( R_4 \). Diodes \( D_1 \) and \( D_2 \)
protect \( T_1 \) and \( T_2 \) 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
\( T_1 \) and \( T_2 \) collectors and \( R_4 \). 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
4kHz. At signal frequencies less than about 50kHz, the
impedance of this tuned network is inductive, having a maxi-
mum 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 \( T_1 \) and \( T_4 \) on current flows
in the load in one direction and in reversed when \( T_2 \) and \( T_3 \)
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,
Meidt, J. D., Micropower circuits, Wiley, 1969, pp.61, 64 & 65.

Cross references
Series 4, card 8.
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).


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.


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 output 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.

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 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 one state of \( T_{R1} \) saturated, \( T_{R2} \) off, to the state of \( T_{R2} \) saturated, \( T_{R1} \) off. For a given value of \( V_{CC} \), the rate of switching depends on the time constant \( C_1 R_1 \), \( C_2 R_2 \), and on the potential to which the base resistors \( R_1 \) and \( R_2 \) are returned. The more conventional configuration will return \( R_1 \) and \( R_2 \) 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 \( C_1 R_1 \) and \( C_2 R_2 \) 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 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. 1. Series RC produces an exponential waveform across the capacitor.](image1)

![Fig. 2. Basic astable multivibrator. Variation of \( V_{BB} \) changes the period.](image2)

![Fig. 3. Waveforms encountered in the circuit of Fig. 2, where \( C_1 R_1 \) is longer than \( C_2 R_2 \).](image3)

![Fig. 4. A single-capacitor astable multivibrator providing mark-to-space ratio adjustment and anti-phase outputs. Circuits using one capacitor are normally less flexible.](image4)

![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.](image5)

![Fig. 6. Two possible combinations of feedback types using a single differential amplifier.](image6)
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. 7. Four combinations of RC feedback and amplifiers used in the single-ended mode.

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.
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Series 8: Astable circuits

Complementary m.o.s. astable circuit

Typical performance
IC: CD 4007AE
Supply: +10V
Rf: 100kΩ; Rs: 1MΩ
C: 10nF; f: 424Hz
Load resistance: ∞
Supply current: 280μA

Square wave available at
Vout
Output excursion: 0.03
Mark-to-space ratio: 0.93
Rise time: 200ns

Component changes
- With supply of +10V, Rf of 100kΩ, and C of 2.2nF, mark-to-space ratio varies from 0.76 to 0.92:1 for Rs from 0 to 1MΩ.
- Components as listed in typical performance data but with finite load resistance Rf. Output pulse level falls, typically by 10% when Rs = 2-kΩ.
- Minimum value Rf for acceptable waveform: 6.8kΩ. Waveform improved by using third inverter as buffer. With Rs of zero, Rf: 6.8kΩ, C: 39pF, f is 610kHz (supply 10V). With Rs of zero, Rf: 10kΩ, C: 10pF, f is 650kHz (supply 10V).
If supply is increased to 15V, f is 900kHz.

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 Rs, 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 Rf, the voltage across Rs and hence the voltage applied to the gate of the first inverter, falls. When this voltage at the junction of C and Rf 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 Rf and when the voltage at the junction of C, Rf and Rs 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 Rs also improves the frequency stability of the circuit with respect to supply voltage changes, and should be at least twice the value of Rf.

Further reading
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
Wireless World Circard

Series 8: Astable circuits

R.t.l. astable circuit

Typical performance
μL914 package contains
— four 2N706-type
— R_D of 450kΩ; R_C of 640Ω

External components:
R₁, R₃: 10kΩ ± 5%
C₁, C₂: 100nF ± 10%
Supply: 3.6V, 6.5mA
P.r.f. 699Hz (see graph)
Mark to space ratio: 1.06
V_out1 waveform 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 Tr₁ and Tr₃ from the circuit. Transistors Tr₂ and Tr₄ 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 Tr₂ on and Tr₄ off. In this state the circuit levels are: Tr₄ collector: V_CE(sat), Tr₂ base: V_BE(on), Tr₂ collector: +V and Tr₄ base: approx. +V due to the negative-going transition at Tr₃ collector. When switched off to on the charge on C₁ cannot change instantaneously. C₁'s charge will then change with a time constant C₁R₁, as its right-hand plate attempts to change to +V from −V. However, when this potential slightly exceeds 0V, Tr₃'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 Tr₂ base through C₂ switching Tr₂ off. The circuit is now in its other quasi-stable state. This action repeats continuously, producing antiphase square waves at Tr₃ and Tr₄ collectors. The off-times of Tr₃ and Tr₄ are given by t₄ = 0.6931C₁R₁ and t₃ = 0.6931C₂R₂ sec. The p.r.f. of the square waves is thus: f = 1/T, where T = t₃ + t₄. The mark-to-space ratio is adjustable by altering the ratio C₁/C₂ and/or R₁/R₂.

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 C₁ and C₂: 220pF to 660pF (p.r.f. ≈ 1.4Hz).
Mark-to-space ratio: 6:8:1 (C₁:100nF, C₂:22μF) to 1:8:5 (C₁:100nF, C₂:220pF), V_out(max): 1.8V.
Useful range of R₁ and R₂: 2.2kΩ 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 C₁R₁ and C₂R₂ 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 R₁ and R₂ in the form of ganged potentiometers. See circuit left, where R₁, R₂ are 2.2kΩ and R₃, R₄ 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 adding the position of the slider of R₄ (middle circuit) without changing the p.r.f. since f = 1/T and T = t₃ + t₄. Hence, with C₁ = C₂ = C and R₁ = R₃ = R then T ≈ C(R₊R₃) + C[R₊(R₄ − R₃)] ≈ C(2R₊R₄) which is independent of R₄.
- Circuit on right shows modification to use only one capacitor. Useful ranges of C₁ and R₄ are 100pF to 100μF and 470 to 10k respectively. The circuit may be externally synchronized by positive pulses at Tr₁ or Tr₄ 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

Cross references
Series 8, cards 8, 12.
Complementary astable circuit

Circuit description
When the supply is connected, Tr1 base and Tr2 collector are at a potential determined by the ratio $R_4/R_3$, which could be in the form of a potentiometer to set the upper level of $V_{out1}$ and $V_{out2}$. The p.d. across $C_1$ is zero, so the base-emitter junction of Tr1 is reverse-biased and both transistors are cut off. Capacitor $C_1$ begins to charge exponentially with time constant $C_1R_4$, 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 $C_1$ discharges through Tr1 and Tr2 with $R_3$ providing a discharge current-limiting action. Diode $D_1$ 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 $C_1$ to recharge through $R_1$ and $V_{out1}$ returns to its initial value determined by $R_1/R_3$.

During the discharge of the capacitor, a narrow negative-going pulse is obtained at the junction of $R_3$ and $R_4$ due to the conduction of Tr2.

**Component changes**
Useful range of supply: +2 to +18V.
Useful range of $C_1$: 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 $R_2$ and $R_4$ 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 $C_1$ and $R_1$.
- Narrow positive-going pulses in antiphase with those at $V_{out2}$ can be obtained by including a small resistor in series with Tr1 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 $R_1$ into $R_{1a}$ and $R_{1b}$ and bootstrapping their junction with $C_2$, as shown left, where $C_2$ should be of the order of 100μF.

If $D_1$ is required to be a silicon diode, additional silicon diodes $D_2$ and $D_3$ 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 $R_2/R_4$ to set Tr1 base to zero volt in the off-state, negative pulses may be obtained at $V_{out2}$.

**Further reading**

**Cross references**
Series 2, cards 5, 12.
Series 6, card 8.
Series 3, card 6.
Series 8, card 1.
T.t.l. Schmitt astable circuit

Typical performance
- IC: SN7413
- Supply: 5V
- R: 330 ±5%
- C: 220nF ±5%
- f: 10.8kHz
- Mark-to-space ratio: 0.5:1

Component changes
- Useful range of R: 220 to 1000Ω. Astable will not function for R > 1.5kΩ.
- Useful range of C: 2.2nF to 22μF.
- 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 ±3%/V.
- Circuit will supply loads from infinity down to 1kΩ, 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 +Vcc.
- 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 C1/10, where C1 is the smaller of the two multivibrator capacitors (circuit middle).
- A further interconnection is shown right. When Vout is high C2 charges at a faster rate when the first oscillator's output is high than when it is low. Similarly, the discharge rate of C2 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: SN7413 Data Sheet.

Cross references
- Series 2, cards 3, 8.
- Series 8, card 10.
Wireless World Circard

Series 8: Astable circuits

Operational amplifier astable circuit

![Operational amplifier astable circuit diagram]

**Typical performance**
- **IC**: 301
- **Supply**: ±15V
- **C1**: 4.7nF ±5%
- **R1**: 4.7kΩ ±5%
- **R2, R3**: 5kΩ ±5%
- **Output square wave**: 28V pk-pk
- **Slew rate**: 8V/μs
- **Variation of frequency with feedback factor and capacitance C1 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 C1R1 and the feedback factor, determined by the ratio of \( R_2/(R_2 + R_3) \). Assume the output has switched to the positive saturation level; the voltage at the non-inverting input is \( +V_{sat}R_2/(R_2 + R_3) \) and the voltage at the inverting input is negative with respect to this value. However capacitor C1 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. C1 will now charge towards \( -V_{sat} \), but again a rapid transition to the positive saturation state will occur when the voltage across C1 reaches \( -V_{sat}R_2/(R_2 + R_3) \), 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 R1: 6.8k to 2.2kΩ.
Useful range of C1: 10μF to 4.7nF for R1 = 4.7kΩ.
Frequency stability: For C1 = 22nF, R3 = 5kΩ, R1 = 4.7kΩ, supply of ±15V and f = 4470Hz, decreasing supply to ±10V reduces frequency by <1%.
Operation possible down to ±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 R1 = 2.2kΩ, R3/(R2 + R1) = 0.7 and C1 from 10μF to 220nF, frequency in the range 24Hz to 1.8kHz. Slew rate 0.6V/μs.
A comparator such as the 72310 will give an output pulse excursion of 0.5 to 2.8V for supplies of +12V and -6V.
For R2, R3: 5kΩ, useful range of R1 is 1.5k to 6.8kΩ and C1 47pF to 22μF giving frequencies in the range 630kHz to 3Hz.

Circuit modifications

- Interchange C1, R1 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. 741), 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 R1a and R1b are selected by the switching action of diodes D1 and D2, D1 conducting when the output is negative, and D2 when the output is positive.

Further reading

National Semiconductor application note AN4-1.

Cross reference

Series 3, card 5  Series 8, cards 10, 12.
Astable blocking oscillator

Typical performance
Supply: +10V, 860μA, -3V 1.1mA
Tr1: BC125, D: SD2
R1: 6.8kΩ; C1: 4.7μF
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: 1.15μs

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 Vcn to +V. Capacitor C1 should be large enough to ensure that the magnetizing inductance of L2 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; −Vmin: −1V
Useful range of R1: 470Ω to 10kΩ
C1(min): 470nF
Minimum load resistance at Vout: 2.2kΩ.
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 causing D2 to conduct clamping the collector at approximately +V. 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 potential-meter R3 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 of −0.5V, Tr1: BSX29; Tr2: BSY17; D1: EA828; R1: 100kΩ, R2, R3: 50Ω, R4: 2MΩ and C1: 50nF.

Further reading
Tecic, S., Multivibrator with very small mark-to-space ratio, Electronic Engineering, 1967, pp. 671-3.
T.t.l. dual inverter astable circuit

Typical performance
Supply: +5V
IC: 7402
R₁, R₂: 1kΩ ±5%
C: 100pF ±5%
Frequency: 2.78MHz
Stability > ±1% for supply in the range 4.75 to 5.25V for a span of 3kHz to 3MHz

Component changes
Useful range of C: 100pF to 22μF.
Useful range of R₁: 220Ω to 1kΩ.
Useful range of R₂: 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 R₁: 1kΩ, R₂: 330Ω, 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 Cᵣ = C/100. Three separate, harmonically-locked astables can then be produced.

Circuit modifications
• Remove R₂ and connect the capacitor in the feedback loop (circuit left). The third inverter sharpens up the waveform. With supply +5V, C: 22nF, R₁: 1kΩ 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 ±10%/V.
• Tuning resistors in the middle network comprise the integrated circuit resistors. With perhaps resistance variations of ±20% from device to device, and a like tolerance over the temperature range −55 to +25°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 ±5% for device or temperature changes. Typically R should be 1kΩ ±1%.

Further reading
Malmstaa 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.
Coupled logic gates astable circuit

Typical data
- Supply: +5V
- ICS: 1 SN7402N
- $R_1$, $R_2$: 2.2kΩ ±5%
- $C_1$, $C_2$: 0.1µF ±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 0 to 1 logic level of approximately +3V due to the input having reached the threshold value. This output transisiton 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 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 $C_2$.

Circuit changes
Use MC7402F, though note pin numbers different.
Useful range of $R_1$, $R_2$ is restricted to ensure that circuit starts: 2 to 3.3kΩ.
Useful range of $C_1$, $C_2$: 27pF to 10µF with the above resistor values.
With $C_1 = C_2 = 0.1µF$, and a supply of +5V, a variation of supply voltage of ±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 $C_2$.

Further reading
- 2MHz Square Wave Generator uses two TTL gates, p. 110, 400 ideas for design, vol. 2, Hayden.

Cross reference
Series 8, cards 7 & 2.
Emitter-coupled astable circuit

Typical performance
Supplies: +5V, 6.4mA; $-5V, 2.5mA$
$R_{1}, R_{2}$: 2N706
$R_{3}, R_{4}$: 470Ω
$R_{5}$: 3.3kΩ; $R_{6}$: 4.7kΩ

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 $T_{2}$ being off and $T_{1}$ on. The emitter current of $T_{1}$ divides into a component in $R_{5}$ and a charging current in $C_{1}$ and $R_{4}$. At the instant that $T_{1}$ switches on this charging current produces negligible p.d. across $C_{1}$ and a p.d. across $R_{4}$ sufficiently large to ensure that $T_{2}$ is off. As $C_{1}$ charges, its p.d. increases and that across $R_{4}$ falls until the base-emitter junction of $T_{2}$ becomes forward-biased. Transistor $T_{2}$ begins to conduct and the emitter current of $T_{1}$ falls causing the collector potential of $T_{1}$ and the base potential of $T_{2}$ to rise. This action causes $T_{1}$ to conduct more heavily and $T_{2}$ to switch off. This sequence is then repeated with $T_{2}$ emitter providing the current to charge $C_{1}$ through $R_{4}$ until the switching action restores the circuit to its original state of $T_{2}$ off and $T_{1}$ on.

Tr_{1} will be on for a time:

$$t_{1} = \frac{C_{1}R_{1}R_{4}R_{5}R_{6}}{R_{1} + R_{4} + R_{5} + R_{6}}$$

and $T_{2}$ will be on for a time:

$$t_{2} = \frac{-V}{\frac{1}{R_{1}} + \frac{1}{R_{4}}}$$

Component changes
Useful range of $+V$: +2 to +14V.
Useful range of $-V$: -2 to -10V.
Useful range of $R_{1}$, $R_{4}$ and $R_{5}$: 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 $C_{1}$: 180pF to 1000μF.
Frequency stability: -1.2%/V increase in $+V$, -6%/V increase in $-V$.
$T_{1}$ and $T_{2}$: 2N708, HE301, BSY95A.

Circuit modifications
- If $R_{4}$ and $R_{5}$ are replaced by a potentiometer connected across $C_{1}$ 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 $T_{1}$ is independent of the supply voltages, and the on-time of $T_{2}$ depends on the ratio $+V/[-V]$. 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 $+V_{CC}$.
- $R_{5}$ and $R_{6}$ 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 $T_{1}$ and $T_{2}$ equal and controlled by $R_{5}$ allowing the frequency to be varied without affecting the mark-to-space ratio.
- The constant-current sources may be voltage controlled, by $R_{5}$ in the circuit shown right and the currents in $T_{1}$ and $T_{2}$ controlled independently by $R_{6}$ and $R_{6}$ 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.t.-compatible output is obtainable using $+V = +5V$.

Further reading

Cross reference
Series 3, card 2.
**Discrete-component Schmitt astable**

**Typical performance**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>R₁</td>
<td>2.2kΩ</td>
</tr>
<tr>
<td>C₁</td>
<td>10nF</td>
</tr>
<tr>
<td>P.r.f.</td>
<td>9.83kHz</td>
</tr>
<tr>
<td>Mark-to-space ratio</td>
<td>1.13/1</td>
</tr>
<tr>
<td>Rise time</td>
<td>400ms</td>
</tr>
<tr>
<td>Fall time</td>
<td>300ns</td>
</tr>
</tbody>
</table>

**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 +Vcc 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 R₂, R₃ and R₄ and partially through R₃, Tr₂ 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 +Vcc: +4 to +15V.

Useful range of C₁: 100pF to 1000μF.

Frequency stability: -0.46%/V increase in +Vcc.

R₃(min) 47Ω (m-s ratio 1/5), R₃(max) 1MΩ (m-s ratio 48/1).

R₄(min) 680Ω (p.r.f.: 77kHz), R₄(max) 100kΩ (Vout reduced to +5.2V).

R₅(min) 47Ω (m-s ratio 5/1), R₅(max) 2.2kΩ (m-s ratio 1/3, f ≈ 31kHz).

R₆(min) 47Ω (m-s ratio 1/2.8, f ≈ 33kHz), R₆(max) 2.2kΩ (Vout reduced to 5V pk-pk).

Useful range of R₇ ≈ 220Ω to 47kΩ.

Tr₁ and Tr₂: ME4103, BC107, BC109, 2N3904.

Observe VBE(max) as well as VCE(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 R₆, 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 Tr₂, as shown middle, allows the timing of the output square wave to be more nearly controlled by the R₁C₁ time constant. Tr₁ and Tr₂ form a long-tailed pair so that the junction of C₁ and R₄ 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 Vout to swing almost between the levels of the supply rail potentials.

**Further reading**


**Cross references**

Series 2, cards 2, 7.

Series 8, cards 4, 5 & 12.
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.i.l. monostable i.c.s. 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 R3 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 R4, 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 R4, 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.c.s without any loading effects on the interconnection circuitry.

Component changes
With C1, C2: 10μF and minimum setting of R1, frequency is 1.6kHz. 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
Astable circuit with f.e.t.

Typical performance
Supply: +9V, 760μA
Tr1: 2N5457
Tr2: BC126
R1: 100kΩ; R4: 10MΩ

- \( R_2: 3.3kΩ \)
- \( R_4: 6.8kΩ \)
- \( C_1: 1nF \)
- P.R.f. 31.2Hz

Mark-to-space ratio: 1:17:1

Circuit description
When the supply is connected, the base of transistor Tr1 is set to the voltage \( V_{CC} \). The collector potential of Tr1 jumps sharply to a level approaching \( +V_{CC} \) and the source of Tr1 jumps towards \( V_{CC} \) of \( R_2 \) and \( R_4 \). With \( C_1 \) initially charged, the positive input to 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 \( R_2 \) into ground via \( R_4 \). The initial charging current of \( C_1 \) is thus larger than it would have been if \( C_1 \) charged simply through \( R_2 \). The p.d. across \( R_4 \) falls rapidly as \( C_1 \) charges through \( R_4 \) causing the f.e.t. junction to be reverse biased and the charging time constant of \( C_1 \) to charge to the much larger value of \( C_1R_2 \). Capacitor \( C_1 \) continues to charge until the gate potential of Tr1 falls below its source potential by an amount that approaches the pinch-off voltage causing Tr2 to switch off due to the reduction of base current. The output switches back to virtually 0V as \( C_1 \) discharges through \( R_3 \), \( R_4 \) and \( R_6 \), where \( R_3 \gg R_4 + R_6 \), 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 \( C_1 \) 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 \( C_1: 10pF \) to 100μF, low-leakage type.
- \( R_{1(min)}: 4.7kΩ \)
- \( R_{max}: 1 \) to 200MΩ.
- Useful range of \( R_2: 2.2 \) to 33kΩ.
- Useful range of \( R_4: 330Ω \) 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 \( C_1 \) charges through \( R_1 \) until the gate potential of the n-channel f.e.t. reaches the threshold voltage of the unijunction transistor \( Tr_2 \), \( -V_{GS(off)} \) of \( Tr_1 \). On reaching this gate voltage, \( +V_{CC}(R_1/R_4 + R_6) - V_{omin} \), the gate-source junction of \( Tr_1 \) becomes forward biased and \( C_1 \) discharges through it to ground. The output pulse across \( R_4 \) and \( R_6 \) providing an output pulse across \( R_3 \). Transistor \( Tr_2 \) then cuts off and the charge-discharge cycle of \( C_1 \) 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 \( Tr_1 \) of a Schmitt astable has been changed to an n-channel f.e.t. but for a given pinch-off voltage and \( R_1 \)-value there may be insufficient drain-source p.d. to allow the switching action to take place, except by critically adjusting the ratio \( R_1/R_2 \). This situation is improved by inserting a zener diode between the drain of \( Tr_1 \) and the base of \( Tr_2 \).
- Circuit right shows a cross-coupled astable where n-p-n transistors have been replaced with n-channel f.e.t.s. This circuit will not switch unless \( R_3 \) and \( R_4 \) 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.

Cross references
Series 8, cards 2, 5 & 10.
Astables

1. This circuit is a high performance complementary astable. The elements determining the switching functions are the complementary pairs $T_{R_a}$ and $T_{R_b}$. These act as complementary Schmitt triggers, channeling the current in $R_a$ via the long-tailed pair formed by $T_{R_a}$, into one or other of the load resistors $R_b$, $R_c$. Transistors $T_{R_a}$, $T_{R_b}$ form part of two distinct circuit functions viz the Schmitt and long-tailed pair switches. The l-to-p transition ensures excellent switching performance since the total current remains constant while the voltage changes required at the bases of $T_{R_a}$, $T_{R_b}$ to achieve this switching remain small. Timing capacitor $C_1$ is charged alternately by $T_{R_a}$, $T_{R_b}$ the other transistor of the pair having its collector current absorbed by whichever of pair $T_{R_a}$, $T_{R_b}$ is conducting at that time. Transistors $T_{R_a}$, $T_{R_b}$ 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.

![Diagram 1](image1)

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.

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, i.e. diodes 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.d.s 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—by 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 divided 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 emitters 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 i.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-crystalline materials (CdS photoconductive cells). Light falling on a p-n junction 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 single 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

\[ R_{opt} = \frac{R_T}{1 + \frac{R_T}{R_S}}. \]

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.](image)

![Fig. 4. Lamp intensity is a linear function of current, except at higher temperatures (broken line).](image)
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 voltae 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.

Fig. 9. Because the spectral response of cadmium sulphide cells matches eye response they are widely used in photographic applications.

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 phototyristors 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 phototyristor 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 is 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.
Null, level and overload l.e.d. indicators

Typical performance
Supply: -9V
IC: 301
Diode: TIL209
R₁, R₂: 10kΩ
R₃: 10kΩ variable
C: 1µF
LED lights for Rₓ < R₃
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 \( R₁ = R₂ \) bridge balance is achieved when \( R₃ \) is adjusted to equal \( Rₓ \). The lower limit of \( Rₓ \to 0 \) (though with zero offset problems in op-amp) while the upper limit is set by \( R₃ \) and the minimum input current of amplifier.

As shown, the l.e.d. is off for \( Rₓ > R₃ \) which makes bridge convenient for portable use with unknown contacts normally open. Quiescent current is standby current of op-amp plus current in \( R₃ \) 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
R₁: 100Ω to 1MΩ. At lower values current drain is excessive. At high end amplifier input current unbalances bridge.
R₃: As above. Accuracy of matching of \( R₁, R₂ \) or a suitable ratio determines overall accuracy.
Rₓ: Range chosen to suit \( Rₓ \). A linear pot is easier to calibrate.
C: Not critical; used to minimize effects of hum pickup.
IC₁: Any op-amp if bridge balance is achieved at potentials near midpoint of supply. For \( Rₓ, R₃ \) small, amplifier common-mode range must extend near to most positive supply potential. 301, 307.
D₁, R₄: Not critical. \( R₄ \) not required if op-amp has internal current limiting. D₁: HP 5082-4440, Monsanto MV5094, etc.

Circuit modifications
- While op-amp may be used as null indicator for ground-referred voltages this may require ± supplies. By using a Darlington pair and a GaAs l.e.d., the transistors will conduct for input voltages close to zero (i.e.d. p.d. ≤ 1.4V at 5mA).
- For negative inputs transistors non-conducting, all current flows in l.e.d.; for positive inputs all currents diverted into transistors. If input > +500mV input base-collector becomes forward biased.
- The l.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 l.e.d.s can replace diodes in voltage limiting circuits while indicating overload. Examples include voltmeter protection with a series limiting resistor to protect l.e.d. For a high resistance voltmeter of say 1V f.s.d. a resistor of 1kΩ could limit current to l.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 l.e.d.s illuminated).

Further reading
Steen, T. J., Three-l.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.d.s: 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 \( T_{R1} \) is driven into conduction by input logic signals, \( T_{R2} \) conducts also. The low saturation voltage of \( T_{R1} \), together with junction p.d.s of \( T_{R1} \) and \( D_1 \) prevent \( T_{R2} \) from being forward biased. The I.e.d. is effectively short-circuited, speeding fall-time of light pulse. With \( T_{R1} \) off, \( T_{R2} \) receives no bias but \( T_{R2} \) 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, \( T_{R3} \) 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 interent 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 \( R = (V_s - V_{in,i})/I_{e.d.} \).

Alternative I.e.d.s: 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 \( I_F = 40\text{mA and } C_{\text{CE}} = \frac{C}{200} \)
\( V_F > 200\text{mV} \)
\( R_T > 1.25\times \text{k}\Omega \text{ for } V=10\text{V} \)
\( R_L > 2.5\times \text{k}\Omega \text{ for } V=20\text{V} \)
\( r_{\text{tot}} = \frac{V_{\text{CE}}}{I_{\text{C}} \text{ with } I_P = 0} > 5\times \text{M}\Omega \)
\( r_{\text{on}} = \frac{V_{\text{CE}}}{I_{\text{C}} \text{ with } I_P = 40} \text{ (see graph at bottom)} \)

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 \( I_{\text{F}} \) is applied. This is ensured if \( V/R_{\text{T}} \) is less than the current at the knee of the characteristic for the particular \( I_{\text{F}} \) used. Graph shows that the greater the value of \( R_{\text{T}} \), the smaller is the value of the voltage drop over the transistor. \( R_{\text{T}} \), however, cannot be increased indefinitely—

if it becomes comparable to \( r_{\text{tot}} \) then voltage will appear over \( R_L \) when \( I_{\text{F}} \) is zero. The characteristics show that the voltdrop over the transistor is reduced the larger the value of \( I_{\text{F}} \). For example, a 50% reduction in \( V_{\text{CE}} \) was observed when \( I_{\text{F}} \) was increased from 15 to 60mA, \( V \text{ and } R_L \) being fixed at 20V and 10k\Omega respectively. In addition, for a given \( V \), increasing \( I_{\text{F}} \) allows lower values of \( R_L \) to be used. For this device \( V_3 \) 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: OPI002, OPI062, 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, \( T_{\text{R}} \) is off and current passes through \( D_1 \) and \( D_2 \) (a general-purpose diode). Phototransistor \( PT_1 \) then conducts and \( V_4 \) is applied to the load, \( R_L \). When \( T_{\text{R}} \) conducts, however, the voltage across \( T_{\text{R}} \) and \( D_1 \) falls below that necessary to make \( D_1 \) and \( D_2 \) conduct because the voltage across \( T_{\text{R}} \) when conducting is less than that necessary to make \( D_1 \) conduct. \( PT_1 \) is then switched off, \( PT_2 \) is switched on and \( V_4 \) 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 10k\Omega, \( R_2 \) 100\Omega, and \( R_4 \) 10k\Omega, minimum \( V_4 \) is 4V and maximum \( V_4 \) is 4V. Range for \( V_1 \) and \( V_2 \): up to ±30V. 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

Cross references
Series 9, cards 4, 7, 8, 9 & 12.
Integrated-circuit optoelectronic switch

at distance of 2-5cm with optimum alignment. With I.e.d. and IPL15 in contact (negligible leakage) switching initiated by \(< 10\mu A\) in I.e.d.

Low sensitivity:
- R: 100k\(\Omega\)
- C: 3.3nF

Switching at room ambient levels.

Load voltage on-condition: 12V
Load voltage off-condition: \(\approx 0V\)

**Typical performance**
- High sensitivity:
  - R: 680k\(\Omega\), R\(_L\): 1k\(\Omega\)
  - C: 220nF
- Supply: 15V
- Ambient light excluded
- Output switches for 150\(\mu\)A in I.e.d. (TTL209)

**Component changes**
- R: 47k\(\Omega\) to 1.5M\(\Omega\)
- C: 2nF to 2\(\mu\)F
- R\(_L\): 330\(\Omega\) to \(\infty\)
- 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.

**Further reading**

**Cross references**
Series 9, cards 3 & 8.

**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 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 forth 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.

![Diagram](image-url)
Characteristics and applications of I.e.d.s

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.d.s 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 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 (≈ -2mV/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 device 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 ≈ 10mA 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 VBE 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 voltage 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).

Further reading
Yen, T. T., LED doubles as sensor, Electronics, 4 Dec. 1972, p.113.

Cross references
Series 9, cards 1, 2 & 11.
Op-amp/comparator driving of 1.e.d.s

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 1.e.d.s 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 1.e.d.s. The amplifiers used have internal current limiting that prevents the 1.e.d. current exceeding a safe value. For low current 1.e.d.s, 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 1.e.d.s 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 1.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 1.e.d. will be significantly illuminated.

Lamp control by op-amp difference mode
Supplies: ±15V, CM: ±12V

<table>
<thead>
<tr>
<th>DM</th>
<th>$R_s$</th>
<th>$D_1$</th>
<th>$D_2$</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>+</td>
<td>0</td>
<td>on</td>
<td>off</td>
<td>L.e.d. current limited by op-amp s/c current limit</td>
</tr>
<tr>
<td>-</td>
<td>0</td>
<td>off</td>
<td>on</td>
<td>Current may be reduced by increasing $R_s$</td>
</tr>
<tr>
<td>+</td>
<td>$-V_s$</td>
<td>on</td>
<td>off</td>
<td>$D_1$, $D_2$ may glow for $R_1$, $R_2$ &gt; 150Ω</td>
</tr>
<tr>
<td>-</td>
<td>$+V_s$</td>
<td>off</td>
<td>on</td>
<td></td>
</tr>
<tr>
<td>+</td>
<td>$+V_s$</td>
<td>off</td>
<td>on</td>
<td></td>
</tr>
</tbody>
</table>

For CM voltages within 1.5V from most negative potential on i.e. 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 500Ω). Some comparators only have single polarity outputs allowing on/off applications only.
- Diodes $D_1$, $D_2$ may be gallium arsenide, gallium arsenide-phosphide 1.e.d.s with $p$.d.s of 1.5 to 3V and currents of 530mA. TIL209, HP 5062-4484, Monsanto MV5094, RS LED2.
- Resistors $R_1$, $R_2$ chosen to prevent illumination by quiescent currents. Typically 150Ω to 1kΩ.

Circuit modifications
- The 1.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 1.e.d.s 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 1.e.d.s but with a very high input impedance (above, right).

Cross references
Series 9, cards 2 & 11.
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 4.7kΩ) 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 4.7kΩ for r.t.l. and from 10kΩ 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 <300Ω, or an additional negative bias time be available. The choice of R is a compromise between worsening noise level margins when R → 300Ω and requiring excessive current from the phototransistor as R → 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 Vbe, worsening the noise margin. The second circuit interchanges the positions of Tr1 and R1 while adding a logic inversion by using Tr2 in common emitter mode. This retains overall function required (phototransistor illuminated gives logical '1' at input) and retains good noise margin as Tr2 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 = 1MΩ requires a photocurrent of <10μA in diode for Vx 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 R1/R2. The output transition is also speeded up and allows operation of succeeding counters from slowly-changing light intensities.

Further reading

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 \( I_E \). 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

\[
\begin{align*}
V_{oc} & \text{ measured with instrument having } 1M\Omega \text{ input resistance} \\
I_{se} & \text{ measured with instrument having terminal volt drop of less than } 12mV.
\end{align*}
\]

Graph right shows that, over a wide range of \( I_F \) (corresponding to \( I_F > 10mA \)), \( V_{oc} \propto \log I_F \)

Graph left shows \( I_{se} \propto I_F \).

Further reading


Cross references

Series 9, cards 3, 4, 7 & 10.
Optically-coupled isolator: pulse characteristics

Measurement data
Phototransistor TIL112. V\textsubscript{o} monitored on c.r.o. with 1M\Omega, 28pF input impedance. I\textsubscript{F} generated from pulse source having 10ns rise and fall times when feeding 50-\Omega load.

Experiment description
Graph top right was obtained by varying I\textsubscript{F} to obtain particular values of I\textsubscript{C}, and by deduction V\textsubscript{CE}, and then measuring t\textsubscript{r} and t\textsubscript{o}. Values of t\textsubscript{r} were identical, within experimental error, to the values for t\textsubscript{o}. Clearly t\textsubscript{r} and t\textsubscript{o} are proportional, though not linearly, to R\textsubscript{L} and are only slightly affected by I\textsubscript{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\textsubscript{F} constant at a value such that I\textsubscript{C} when supplying a short circuit was 10mA. In this case R\textsubscript{L} was varied sufficiently to obtain readings when transistor was saturated. Graph above shows a typical pulse shape for low R\textsubscript{L} and graph right for high R\textsubscript{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\textsubscript{r} and t\textsubscript{o} 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\textsubscript{L} can be increased until V\textsubscript{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

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 \( R_{opt} = V_o/I_{sc} \) 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 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 is 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 \( R_x \) 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 \( R_x \) and \( R_y \) 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

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 here 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 used 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.\( \)s 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.
Cross references
Series 9, cards 1, 2, 5, 6, 7 & 10.
Choppers and rectifiers

Typical performance

Opto-isolators TIL112
R1, R2: 100Ω; R3: 10kΩ
C1: 10pF
Drive frequency: 1kHz
Drive pulse height 5V
V: 10mV from 10–Ω source
V0 = 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 I.e.d. drive varying
from 3 to 6V, meter change < 1%. With drive
at 5V and \( C_1 \) varying
from 10pF to 0.01μF,
meter change < 2%. With drive at 5V and \( R_3 \)
varying from 2 to 33kΩ,
meter change < 4%.

Circuit description

Circuit 1 shows a chopper circuit using a pair of opto-isolators
with phototransistor output stages. Diodes D1 and D2 are
light-emitting diodes driving the phototransistors PT1 and
PT2, respectively. When one of the phototransistors is illu-
minated, 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 I.e.d.s are driven in anti-
phase then only one phototransistor is on at a time and either
the signal voltage is applied to the resistor \( R_3 \) through PT1
with PT2 non-conducting or PT2 blocks the signal while PT1
absorbs any current which leaks through PT2. 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 PT1 is not essential if the d.c. signal \( V \) is large
but it does have the advantage that any stray or load capaci-
tance 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 I.e.d.s 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
phototransistors 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

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 i.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.e.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.


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 30μs, 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 i.e.d. bringing the thyristor into conduction rapidly discharging the capacitors.


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 near-perfect triangular wave.

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 \( I_c \propto V_{be} \) 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 \( V_{ce} \) is very much less than the \( V_{be} \) values above, though it must be observed that \( V_{ce} \) often rises with temperature while \( V_{be} \) always falls (typically the temperature coefficient of \( V_{be} \) is \(-2mV/K\)). 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 negligible for circuits coming under the micropower heading.

The use of germanium transistors to minimize voltage requirements is well 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 lower 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 (b).

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).
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 1μA 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.c.s. 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.](image)
Low-voltage a.c. amplifier

Typical performance
R₁ 100kΩ, R₃ 22kΩ
R₂ 4.7kΩ, R₄, R₅ 470kΩ
C₁, C₂ 10µF
TR₁, TR₂ BC125
TR₂ BC126
Supply +1.5V
Voltage gain 2100
Output impedance 3kΩ

Input impedance 150kΩ
(all at 1kHz)
With R₁ = pR₂, R₂ = pR₃ etc.
and with equal hᵦₑ
Voltage gain = \frac{20hᵦₑ}{hᵦₑ + 20p}
ignoring hᵦₑ, hᵦₑ 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 Vᵦₑ 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 TR₁ is determined by the p.d. across R₁ which is in turn fixed by the Vᵦₑ requirements of TR₁. The Vᵦₑ of a silicon transistor has a typical variation of <10% for normal ambient temperature, supply and tolerance effects and so TR₁ current is well-defined. The same argument applies to TR₃ current in terms of TR₁ Vᵦₑ. 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 Vᵦₑ characteristics, rather than the widely varying hᵦₑ. As hᵦₑ 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 hᵦₑ.

Component changes
TR₁, TR₂ can be n-p-n small-signal transistor, particularly those with high current gain at low currents e.g. 2N2484, BC109, etc.
TR₂ : p-n-p complement to above.
R₁, R₄, R₅: see circuit description for ratios. Resistor R₂ determines output current (≈ Vᵦₑ/(2R₂)), and R₁ indirectly the input current (≈ Vᵦₑ/hᵦₑR₁). R₁: 4.7 to 470kΩ, R₃: 4.7 to 100kΩ; R₅: 1 to 47kΩ.
R₄, R₅ should be <10R₁ if TR₂ base current is not to develop too large a p.d. across R₃, R₅ pushing output voltage towards +V.
C₁, C₂ determine lower frequency cut-off. Capacitor C₂ decouples negative feedback, gives rise to inductive term in input impedance at low frequencies. C₁, C₂: 0.1 to 500µF.
Supply: circuit operates down to 1.0V with reasonable gain, to 0.85V with reduced gain and output swing. Upper limit set by transistor Vᵦₑ ratings—increase R₁ to limit TR₃ 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 R₂ \gg R₁ when n.f.b. may be too small to be of value. Feedback can be increased if dominant CR lag (R₃, C₂) cuts gain before phase shift reaches 180°. Resistor \approx 0.1R₁ to 0.01R₁, C₂ R₃ 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 R₅ by constant-current load provided by current mirror. (middle)
- Direct-coupling possible to sources with near-zero direct voltage if R₅, R₁₀ chosen in conjunction with R₃ (typically R₅ + R₁₀ \approx R₁ for supply of 1.0V). Decoupling required if supply impedance is significant. (right)

Further reading
Low-voltage astable circuit

Typical performance

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_1$, $C_2$</td>
<td>10nF</td>
</tr>
<tr>
<td>Supply</td>
<td>1.2V</td>
</tr>
<tr>
<td>$f$</td>
<td>14.1kHz</td>
</tr>
<tr>
<td>$R_1$ to $R_4$</td>
<td>1kΩ</td>
</tr>
<tr>
<td>$R_L$ (A-B)</td>
<td>470Ω</td>
</tr>
</tbody>
</table>

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 converters 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 $C_1$, $C_2$ 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

$T_{1a}$, $T_{1b}$: Any p-n-p silicon transistor for given supply voltage range. For higher currents use BFR81.

$T_{2a}$, $T_{2b}$: n-p-n, otherwise as above (BFR41). For lower supply voltages substitute germanium complementary pairs 2N1302/03 or 2N1304/05.

$R_1$ to $R_4$: 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.

$C_1$, $C_2$: 100pF to 1000pF depending on frequency required. At high frequencies charge storage effects increase no-load current.

Circuit modifications

- By adding $T_{3a}$ 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-to-dc converter 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 $\approx 1.2V$.

Further reading


Cross references

Series 10, cards 4 & 8.
Low-voltage regulators

Typical performance

<table>
<thead>
<tr>
<th>R_1</th>
<th>R_2</th>
<th>R_3</th>
<th>BC126</th>
</tr>
</thead>
<tbody>
<tr>
<td>1kΩ</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>R_4</td>
<td>12kΩ</td>
<td></td>
<td></td>
</tr>
<tr>
<td>R_5</td>
<td>3.3kΩ</td>
<td></td>
<td></td>
</tr>
<tr>
<td>R_6</td>
<td>R_7</td>
<td>1kΩ</td>
<td></td>
</tr>
<tr>
<td>C_1</td>
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</tr>
<tr>
<td>T_1</td>
<td>T_2</td>
<td>2N1302</td>
<td></td>
</tr>
<tr>
<td>T_3</td>
<td>2N1303</td>
<td></td>
<td></td>
</tr>
<tr>
<td>T_4</td>
<td>T_5</td>
<td>T_6</td>
<td>B2125</td>
</tr>
</tbody>
</table>

For \( V_s = 400\text{mV} \), \( V_s = 4\text{V} \), \( \Delta V_s/\Delta R_s \pm 20\% \), \( \Delta V_s/R_s \pm 20\% \), \( \Delta V_s/V_s \approx \pm 0.5\% \).

Circuit description

This circuit is relatively complex because the problems of voltage regulation at very low voltages are severe. Transistors T_1, T_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 T_3, with a bias voltage for T_4. This latter provides a constant current for the long-tailed pair T_5, T_6, which compare the reference voltage with the regulator output. Transistors T_7, T_8, T_9 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_{sat}(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, T_10, T_11 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 is capable of handling relatively large currents but may be used at currents down to \(<1\text{mA}\) if required.

Component changes

- R_1, R_2: Not critical but should be comparable in value and depend on output current required. Typical range 330Ω to 2.2kΩ.
- R_4, R_5: Ratio used to adjust temperature coefficient of reference voltage. Keep ratio \( R_4/R_5 \) low to minimize supply voltage e.g. \( R_4 = 0 \), \( R_5 \infty \) leading to temperature drift of \( \pm 0.1\% \) (deg C).
- T_1, T_2: Must be germanium for these low voltages. Any general-purpose complementary germanium transistors with T_4, T_5 at equal temperature for low drift.
- T_3, T_5, T_6: General-purpose silicon e.g. 2N2926, ME4103 for n-p-n; ME0413, 2N3702 for p-n-p.
- T_7: High current n-p-n for low saturation voltage; voltage rating unimportant.
- C_1: > 10μF.
- D_1: General-purpose silicon diode.

Circuit modifications

- For output voltages greater than the reference voltage, the reference is compared with a fraction of the output i.e. \( V_1 = (R_1/R_{11} + 1)V_{ref} \). 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.

Further reading

Optical link with low standby power

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 transistor with \( D_3 \) holding the circuit in the non-conducting state and with \( C_3 \) storing charge. On receipt of a positive-going transition \( T_{R3} \) begins to conduct, pulling the base potential of \( T_{R1} \) below the emitter potential as stored on \( C_3 \). The cumulative increase in current in \( T_{R1} \) and \( T_{R2} \) discharges \( C_3 \) (with \( D_3 \) reverse-biased) through the transistors and the i.e.d. The short duration pulse may be at a relatively high current ensuring high intensity from the i.e.d., without excessive mean dissipation. After \( C_3 \) has discharged to the level at which \( T_{R1}, T_{R3} \) come out of conduc-

Component changes
\( R_1 \): This limits rate of recharge of \( C_3 \) and hence recovery time. Typically 39 to 150Ω with higher value better at low repetition rates.

\( R_2 \): Limits peak current in i.e.d. may be lowered to 3Ω for i.e.d. capable of >1A peak current. If high intensity not required may be increased to >100Ω.

\( D_1, D_2 \): Not critical, any general-purpose Si diode, though high-speed preferable for short pulses.

\( T_{R1}, T_{R2} \): Complementary n-p-n/p-n-p silicon transistors. Parameters not critical but peak current rating must be sufficient for required i.e.d. peak current.

Supply: \( +5 \) to \( +20 \) V. At low voltages, high-intensity light pulses difficult to achieve.

I.e.d.: Any low-cost i.e.d. may be used, but optical coupling between i.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 i.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 \( R_{1, 2}, R_3 \) with inefficient coupling.

- Alternatively a phototransistor may replace \( T_{R1} \) and the photodiode \( D_3 \) (middle) This applies equally if the i.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 \( R_4 \) a Schmitt trigger action is obtained as on Circards series 2 card 3. Sensitivity is set by \( R_3 \), hysteresis by the ratio of \( R_3/R_1 \) and the output has good rise- and fall-times for operating following circuits.

Further reading

Cross references
Series 9, cards 3, 4 & 11.
Wireless World Circard

Series 10: Micropower circuits

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 < 1 kΩ 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 when negative resistance is required at micropower levels. At low currents, the total p.d. is of the order of the device threshold voltages (≈ 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, $T_n$ 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 $D_n$ and $D_p$ prevent conduction via the collector-base paths of $T_n$ and $T_p$ 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 $C_1$, $D_n$, $D_p$ and smoothed by $C_2$. 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

Cross references
Series 10, cards 3, 7 & 10.
Micropower crystal oscillator

Typical performance
IC 1/3 × CD4007AE
R₁ 15MΩ
R₂, R₃ 100kΩ

C₁ 15pF, C₄ 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 C₄, 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 C₄ may be omitted with oscillation being dependent on the presence of stray/circuit capacitance. The data is based on an industrial grade i.e., 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 < 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 100kΩ controls amplitude of oscillation at expense of loading π-network.

C₁, C₄: 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 <100kHz to >2.5MHz. 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 ~20pF, the current consumption at 250kHz is ~60µA and rise and fall-times ~0.3µs. 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.
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, ±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; TR6, TR7, and TR8 constitute a current mirror system that define the currents in the long-tailed pair TR3, TR4 and the output stage TR8 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.c.s such as CA3045, CA3046. Alternatively, resistors inserted in emitters of TR1, TR2, TR5, dropping 50 to 500mV, equalize and define currents in sequence 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 \( \frac{1}{4} \times \) 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.c.s 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.c.s mentioned above. As TR1 and TR4 carry the same current, then provided TR3 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 define the gain or the hysteresis if used as a level sensing circuit (Circards series 2). (right)

Further reading


Cross references

Series 10, cards 1, 3 & 9.
Wireless World Circard Series 10: MicroPower Circuits

RC oscillator for low voltages

Typical performance

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_1 )</td>
<td>4.7kΩ</td>
</tr>
<tr>
<td>( R_2, R_3 )</td>
<td>10kΩ</td>
</tr>
<tr>
<td>( C_1 )</td>
<td>22nF, ( C_3 = 10nF )</td>
</tr>
<tr>
<td>( C_2 )</td>
<td>100µF (optional)</td>
</tr>
<tr>
<td>( T_1, T_2 ) matched p-n-p pair from CA3084</td>
<td></td>
</tr>
</tbody>
</table>

\( T_{1b}, T_{4a} \) matched n-p-n pair from CA3046

\( f = 1.77kHz \)

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 \(-4mV/\text{deg}\). 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 \( T_{1b}, T_{4a} \) and \( T_{1a}, T_{2b} \) 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 \( f = \frac{1}{2\pi R_4 C_4 R_5 C_5} \). 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

\( R_1, R_2 \to 100kΩ, R_4 = R_5/2 \).
\( C_1, C_2 \to 1nF, C_3 = 10µF, C_4 = 2C_5 \).

Frequency of oscillation that can be sustained falls with falling current. Setting of R controls the onset of oscillation. \( R \approx R_4 \).

\( T_1 \) to \( T_4 \) 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 \( C_1 > 2C_5 \) and \( R_1 < R_5/2 \). Fixing \( C_2 \) at 0.01µF, \( C_3 \) at 0.022µF for example allows \( R_4/2 \) 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


Cross references

Series 10, cards 1, 2 & 9.
Class B low-voltage amplifier

Typical performance
- $R_1$, $R_2$, $R_3$: 100kΩ
- $R_4$: 4.7kΩ, $R_5$: 560Ω
- $C_1$: 1μF, $C_2$: 3.6nF
- $C_3$: 100pF
- $Tr_1$: BC125, $Tr_2$: BC126
- $Tr_3$: BFR81, $Tr_4$: BFR41

Supply: 1.2V, 2.1mA
- $R_L$: 100Ω
- $V_o$: 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 $V_{EE}$(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 voltagee 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 $h_oe$ at low currents.
- $Tr_3$: ME4103, BC109, 2N2926, 2N930, 2N3707, etc.
- $Tr_4$: ME0413, BC179, 2N2904 etc.
- $Tr_5$: Low saturation voltage p-n-p: BFR80, BFR81 etc.
- $Tr_6$: Low saturation voltage n-p-n: BFR40, BFY50 etc.
- $R_1$, $R_2$: Set gain and input impedance. Increasing $R_3/R_1$ increases gain, reduces feedback upsets output quiescent if taken too far. $R_2$, $R_3$: 10 to 47kΩ.
- $R_4$: Typically $R_4/5$ to $R_4/50$.
- $C_1$, $C_2$: Standard input/output coupling. $C_1$: 0.1 to 10μF, $C_2$: 10 to 500μF.
- $R_5$, $C_3$: 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. $Tr_3$ 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

Cross references
- Series 10, cards 1 & 8.
Wireless World Circard

Series 10: Micropower circuits 10

Low-voltage d.c. converter

Typical performance
- R₁: 470kΩ
- R₂: 1MΩ
- C₁, C₂, C₃: 22µF
- IC: 1/3 × CD4009AE
- D₁ to D₅: Bridge rectifier
- R₄: 1.6µA

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 R₁ and R₂ 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 C₄ is charged via D₄ to a p.d. less than the supply voltage by one diode forward voltage. When the B returns to zero, the output of C₅ tries to swing negative bringing D₃ into conduction and transferring charge from C₄ to C₅. At low repetition rates the continuous loss of charge to R₂ prevents a significant accumulation of charge in C₅. As the repetition rate increases (or C₅ increases) the p.d. across C₅ approaches a magnitude of \[ V_r - 2V_f \] where \( V_r \) 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 \( ±1V \) 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)
- R₁: 470kΩ upwards
- D₁: Bridge rectifier or 4 × 1N914, etc.
- R₃: 10kΩ to 10MΩ.
- R₅: 22kΩ to 22MΩ.
- C₁, C₂: 100nF to 100µF.
- C₅: 1 to 100µF.
- 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 R₁, the on-resistances of the m.o.s. devices within c.m.o.s. inverters limit output. Choose R₁, R₃ 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 output voltages, 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.

Cross references
Series 10, cards 2, 5 & 12.
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 \( \approx -2 \text{mV/deg. } C \). 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.

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 is 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.

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 \( \approx -2 \text{mV/deg. } C \), increasing in magnitude at lower current densities.

Assymetric 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°C. Voltage stability even at currents below those for which the devices are specified is generally adequate for biasing transistors.

Further reading

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 (µA722, 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 ±1V 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 $\approx +18mV$). The temperature drift in $V_{be}$ is around $-2mV/\deg C$ 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 $\approx 430mV$ 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 $1/2R_1C_4$ with $R_4C_4 = 2.2R_1C_1$. Via pins 3 and 9, $C_4$ swings between ground and $V_{REF}$, while $C_1$ 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 $V_{C1}$ exceeds the trough of $V_{C2}$ 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 in $V_{REF}$ can then control the pulse width.


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


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 Vbe changes. The article also describes a low-voltage complementary astable which can be modulated by the amplifier output.
