Sets 25 & 26: RC oscillators

One might have expected the introductory article to deal with the point of theory concerning nullors. They are instead introduced in a concise and easily assimilble way on page 67, making any explanation here superfluous. Those unfamiliar with nullors will see from page 00 the usefulness of the concept in rearranging and regrouping circuits to generate new circuits (even though the method gives no phase information). See also Circuit Designs 1, pp.122/3. Circuits covered are obvious from the titles, the first ten or 11 covering Wien circuits, single-component frequency control and amplitude control methods. Set 26 deals mainly with phase shift and T-circuits.

RC oscillators

Amplifiers oscillate; oscillators may not. These guiding principles have been developed and confirmed over many years of patient experimenting, not least during the preparation of Circards. Early versions of operational amplifiers were particularly critical of the source/load/supply impedances and were prone to oscillate at high frequencies unless carefully used. Early transistors had low values of current gain and cut-off frequencies making it difficult to produce controlled oscillations.

These properties point to a dividing line between oscillators and amplifiers with feedback viz that they are of the same kind, differing only in the quantity and nature of the feedback. The point can be illustrated by Fig. 1 in which an amplifier of gain A has a portion of its output volage β subtracted from the signal at the input. The gain of the amplifier with feedback can be greater or less than A, and the output will in general differ in phase. For well-controlled characteristics, the phase-frequency response has to be such that the feedback does not become regenerative until the magnitude of the βA term is below unity.

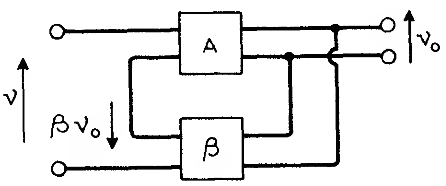
Feedback theory is formally expressed in many different ways, but one graphical approach that is helpful is to consider the root locus (Fig. 2). The graph plots the locus of the system transfer function as the frequency varies. Points on the horizontal axis correspond to phase shifts of zero (to the right of the origin) and 180° (to the left). Points on the vertical axis represent phase shifts of +90° and −90°. The distance of a point from the origin represents the magnitude of the transfer function. Thus in many amplifiers the region of the locus near the horizontal axis would represent a very wide range of frequencies since the gain remains constant and the phase-shift is zero or 180° over this range.

An important point on this graph is the point $1\angle 0°$. A general criterion, due to Barkhausen, suggests that if the locus of the system response does not enclose this point then the loop may be safely closed and the feedback will not cause the amplifier to become unstable. An exceptional state of conditional stability can result where the amplifier/feedback network has multiple reactive elements producing a complex locus which would enclose the point in the event of a fall in the magnitude of the gain.

When the locus passes through the point we have $\beta A = 1\angle 0°$ commonly called positive feedback and this constitutes an oscillator of constant but undefined amplitude, i.e. the signal feedback is just sufficient to sustain the output unchanged and without the need for an input signal. Alternatively we may view it as an amplifier of infinite gain, the denominator of the expression, $1-\beta A$, having gone to zero.

The inevitable small variations in β and A caused by temperature, supply or load conditions as well as by long term drift in component values, cause the amplitude either to decay away ($\beta A>1$) or to increase ($\beta A<1$). The limit is set by non-linearities in the system either inherent to the amplifier or deliberately added externally in the feedback network(s). These reduce βA and the oscillations settle down to a stable situation in which the mean value of βA over the cycle is unity.

For good frequency stability a number of precautions have to be observed (1) the amplifier should have negligible or very closely controlled phase-shift at the frequency of oscillation. (2) Amplitude of oscillation should be controlled to minimize distortion, since harmonics are fed back to the



$$(V + \beta V_o)A = V_o$$
$$V_o(1-\beta A) = A_V$$
$$T_V = \frac{V_o}{V} = \frac{A}{1-\beta A}$$

Fig. 1. A fraction of the output is added to the source at the input in deriving a standard form of the basic feedback equation. Positive feedback occurs when βA is positive.

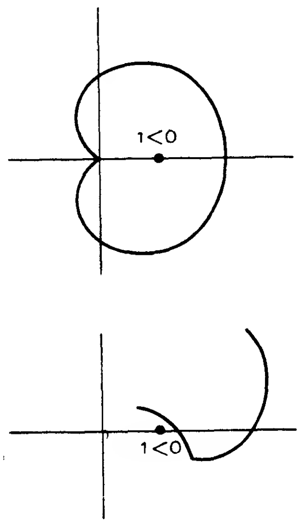
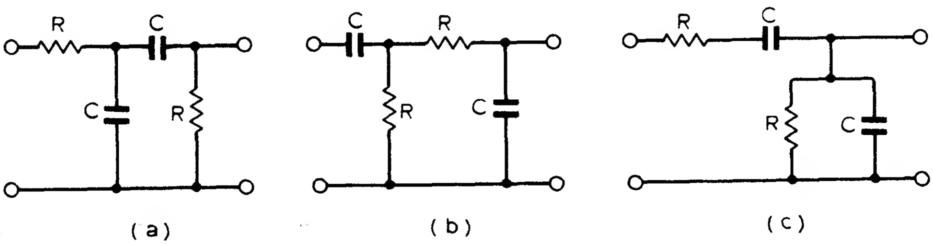


Fig. 2. If the in-phase and quadrature components of the overall loop gain are used as axes, the locus as the frequency is varied indicates the stability of the system.

Fig 3. These three networks have an identical transfer function and can be used interchangeably in oscillators.



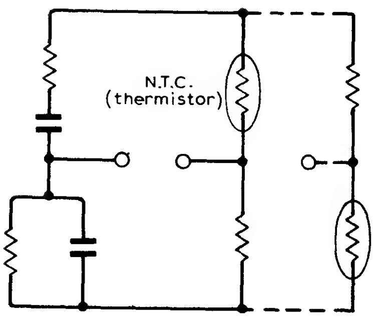


Fig. 4. Some oscillators use temperature-dependent resistors heated by the amplitude of the oscillation.

input and the resulting intermodulation reduces the frequency of oscillation below that predicted from the simple theory. (3) Input and output impedances of the amplifier must not load the RC networks significantly.

A major class of oscillators which includes the Wien bridge circuits, uses networks as in Fig. 3. Using equal values of R, C throughout, the transfer function of each of these circuits is the same, with the output reaching a maximum of one third of the input when the phase shift is zero. Frequency is $1/2\pi RC$. Each can be used with an amplifier of gain $+3$ to produce sustained oscillation. Many other combinations of these networks and amplifiers can be devised, by using current, transconductance and transresistance amplifiers.

Amplitude control may be via a gain-controlled amplifier whose gain is reduced as the output exceeds a given value, usually via a peak- or mean-rectifier and f.e.t. or similar controlled resistor. The classical solution is to use the RC network as part of a bridge configuration with a high-gain amplifier monitoring the bridge unbalance. One of the bridge resistors is made amplitude sensitive, e.g. a filament lamp or thermistor arranged so that increasing amplitude of oscillation increases the amount of negative feedback thus stabilizing the oscillation amplitude Fig. 4.

These oscillators are controlled in frequency over a very wide range commonly by switching in pairs of capacitors as the coarse control or range-setting, with ganged resistors for fine control. The reverse is possible with high input-impedance amplifiers where high resistances allow the use of ganged tuning capacitors. Single-element control has obvious advantages of simplicity and economy, as well as the possibility of remote control via light dependent resistors and the like. Most solutions to this problem require a larger number of amplifiers to provide separate feedback paths by splitting the passive network in some way, and in addition there is an effective loss in Q of the system that results in increased distortion. One example out of many that have been designed is shown in Fig. 6. Frequency ranges of up to three decades have been reported, while the amplitude control mechanisms are similar to those above.

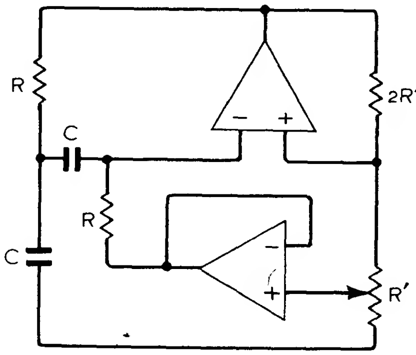
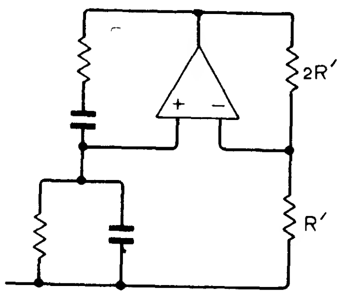
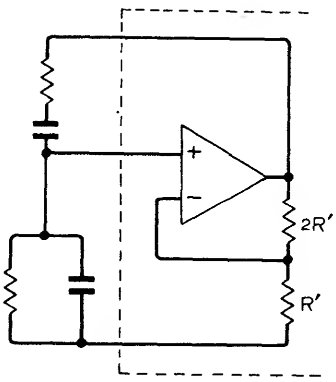


Fig. 6. Adding another amplifier at appropriate points in various oscillators allows a single control to change the frequency without varying loop gain.





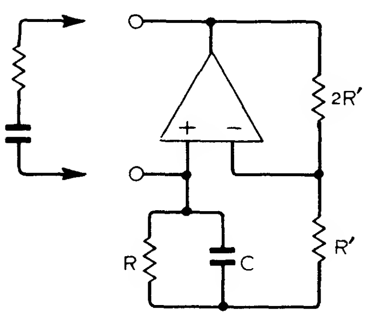


Fig 5 (a). A standard configuration uses an operational amplifier to activate the bridge. (b) Circuit partitioned to show an amplifier of gain +3 with the frequency-dependent network providing positive feedback. (c) Removing one section of the network, the impedance across the port is found to be negative, leading to Negative Impedance Converter interpretations.

T, phase-shift, and two integrators

Since both inverting and non-inverting amplifiers are obtainable, circuits can be designed in which the phase-shifts in the external networks are 180° and zero (or 360°) respectively. An example of the former is the classical three section phase-shift circuit shown in Fig. 7. Using equal values of resistors and capacitors the network attenuation is rather large, the output being 1/29th of the input at the frequency where the overall phase-shift is 180°. It is usually preferred to the alternative form using interchanged Rs and Cs, because the increased attenuation at high frequencies reduces the harmonic distortion and with it the corresponding shift in frequency.

If the RC values are scaled, then with n large each section can be analysed separately since the loading effects of the following section can be ignored. The phase shift of each section is then close to 60° at the critical frequency with a halving of the signal level. The amplifier then needs a voltage gain of -8 but the demands on input and output impedances are more severe (the current that can be drawn from the network without loading it becomes very small while the current needed to supply it increases). Alternative methods are to separate the phase-shift networks, using one amplifier of gain -2 between each section. Fig. 8 shows a related circuit that combines the gain and phase-shifting sections. Variants such as this are convenient for three-phase oscillators particularly as gain required from each stage is minimal.

A separate class of oscillators is based on null/notch/band-stop RC networks in which the signal transfer function tends to zero or a low value at a particular frequency (Fig. 9). These can give improved sharpness of tuning with lowered distortion, but interaction between the impedances can make them less tolerant of component drift. More important is the difficulty of tuning such circuits since several components need to be changed simultaneously. Separating the paths through these networks and driving them with individual amplifiers can allow control of the frequency without change in the amplitude condition.

There is a very close relationship between active filters and oscillators. They share common passive networks and in many cases one can be converted simply into the other by adjusting the damping factor (sharpness of tuning). A very important configuration which has wide application in both fields is the two-integrator loop (Fig. 10). Well-known in analogue computing, it and its near relatives appear under a number of names including 'bi-quad,' 'triple,' 'state-space,' 'gyrator' etc. For ideal amplifiers the circuit Q is infinite without the need for positive feedback and it is particularly suited to the design of high-Q active filters by the addition of a small amount of negative feedback. In practice the net feedback will depend on internal phase-shifts as well as finite-amplifier gains, and both positive and negative feedback may be used to produce controlled oscillations.

If a single resistor or capacitor is varied then with ideal amplifiers, the circuit is still on the edge of oscillation, but the frequency at which oscillation can be sustained is varied. Single-element control of frequency is of considerable advantage in simplifying the construction of oscillators, since dual-gang controls are difficult to keep in a well-matched condition over a wide range. The feedback needed remains small under these conditions, being sufficient only to overcome amplifier imperfections and the finite Q of the capacitors.

Because the amount of feedback required is small it can be introduced via a clipping network that comes into action sharply at a particular amplitude without bringing in significant distortion. This gives instantaneous control of amplitude without the time delay due to heating effects with thermal control. In addition there are three separate outputs with 90° phase differences and the addition of another inverting amplifier gives the fourth phase if required. Again there are a number of combinations of amplifiers and network which share these desirable properties as in Fig. 11. In all of them there is a tendency to instability at high frequencies where the slew-rate limiting of the amplifiers produces a jump phenomenon that locks the oscillator into an output oscillation of higher frequency and uncontrolled amplitude.

Some of these networks are more usually interpreted as forms of impedance inverters/converters, in particular the gyrator, *viz*, a circuit that with a capacitor across one port synthesizes a purely inductive reactance across a second port. If that port has a second capacitor placed across it, a resonant circuit is established which sustains oscillation if a small amount of positive feedback is introduced. It is instructive to draw out the passive networks in such circuits since this clarifies the interrelationships between the various forms of oscillator and filter (Fig. 12).

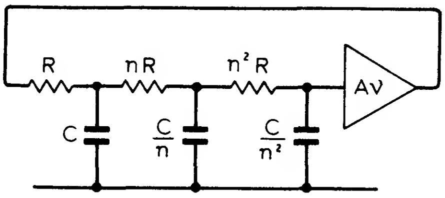


Fig. 7. If the impedances are graded to minimize loading of each section on the preceding one, each contributes 60° to the overall phase-shift at the frequency of oscillation.

$$\left.\begin{array}{l} n=1 :- A_v \longrightarrow -29 \\ n \longrightarrow \infty :- A_v \longrightarrow -8 \end{array}\right\}$$

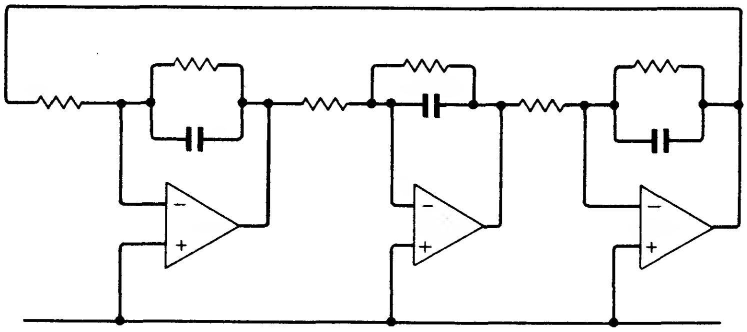


Fig. 8. Three amplifier stages with defined gain/phase characteristics comprise a three-phase oscillator.

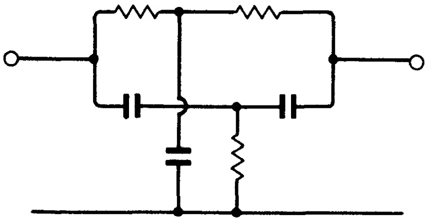
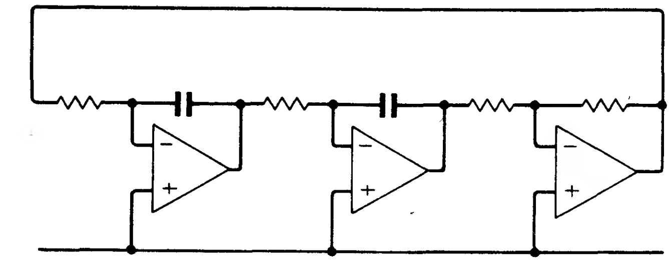


Fig. 9. T-networks can have zero-transmission at a particular frequency. Oscillators utilize positive feedback with the T-network in a negative feedback path.

Fig. 10. Two integrators plus an inverter form the nucleus of a number of oscillators and filters.



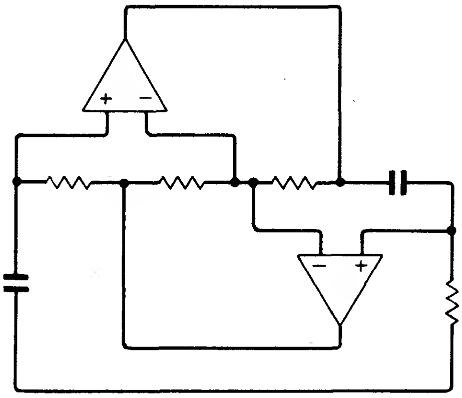
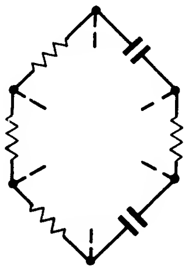


Fig. 11. Gyrators are a class of circuits that synthesize an inductive reactance from a capacitor. An oscillator results from resonating the reactance with a second capacitor.

Fig. 12. The previous two oscillators share a common passive network and can be shown to be functionally identical.



Nullors, networks and n.i.cs

Two circuit elements were devised to complement the short-circuit (zero p.d. at any current) and the open-circuit (zero current at any p.d.). They are the nullator (zero V, I) and norator (arbitrary V, I). Neither has a separate real existence but a high-gain amplifier embedded in a feedback network approximates to the combination. Such an amplifier can be replaced in circuits by a nullor, the name given to the combination, and this can simplify drawings by allowing the separation of input and output networks.

An op-amp has the constraint that one end of the norator is grounded, a transistor that of a common point between nullator and norator. An f.e.t. has a lower gain and requires an additional resistor to simulate it. With multi-amplifier systems if several points are equipotential because of nullator action there are multiple ways of achieving this effect.

To illustrate this consider the differentiator circuit. Feedback theory indicates the alternative of placing an integrator in the feedback path of an amplifier, and this solution can be convenient in analogue computing.

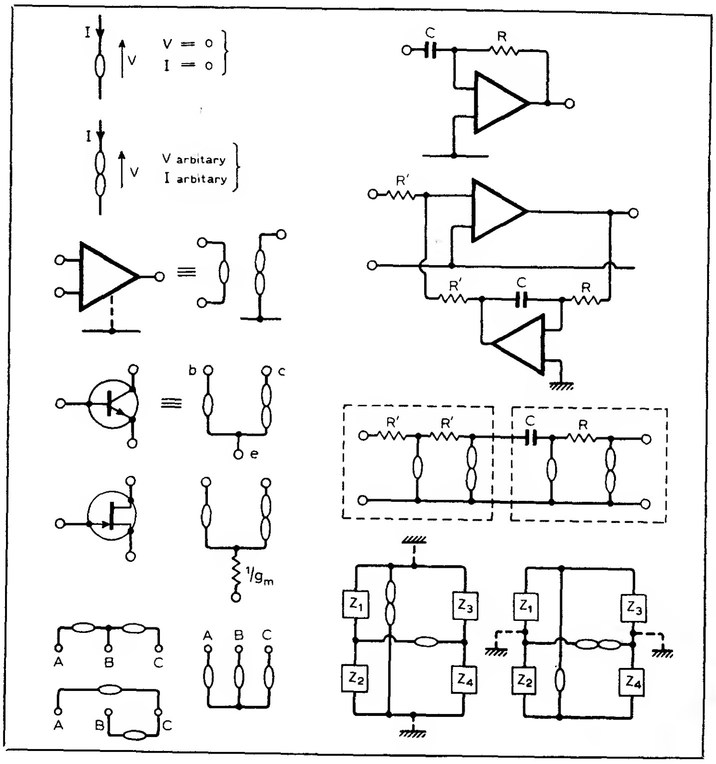
Re-drawing in nullor form, and combining the nullators and norators, shows that the system is equivalent to an amplifier followed by a differentiator. It suggests that drawing in nullor form and then re-pairing is another method of generating new circuits. Yet another method is that of interchanging the positions of a nullator and a norator in changing the source and load in bridge measurements. This lead to four forms of oscillator derived from a common passive network.

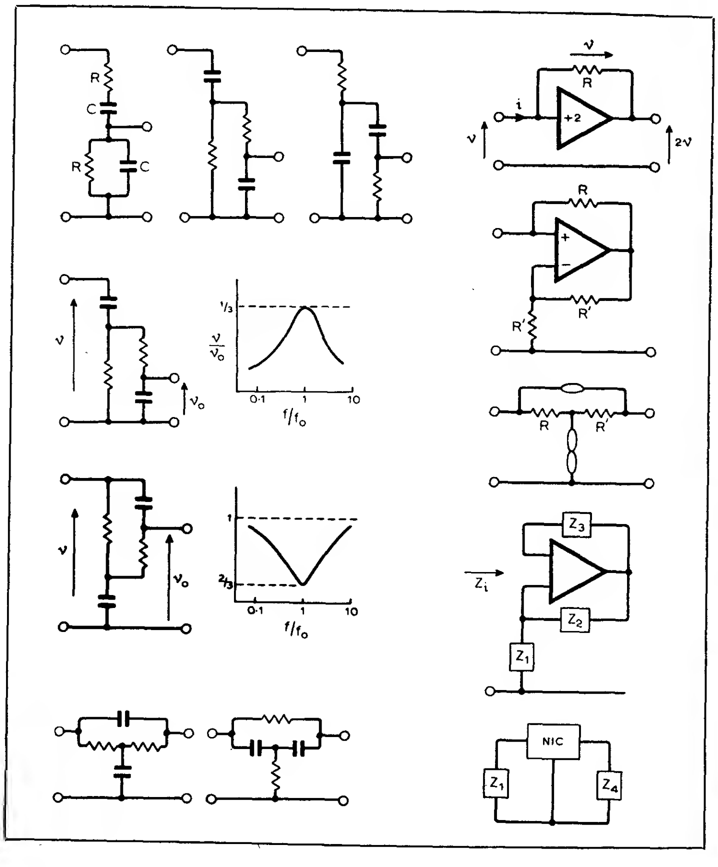
A standard network in RC oscillator design is that due to Wien, having a transfer function

$$v/v_o = T_v = \frac{1}{3+j(f/f_o - f_o/f)}$$

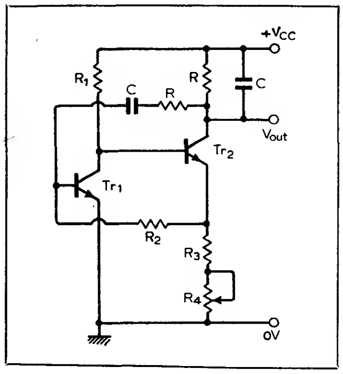
where $f_o = 1/2\pi CR$. Two other networks of cascaded lead and lag sections have identical transfer functions for identical R, C values. The response is a maximum of 1/3 with zero phase shift at $f = f_o$, falling at both high and low frequencies. If the network is inverted (taking the lead-lag as an example) the response has a minimum value at $f = f_o$. The former network can be used as a feedback path, balanced against resistive negative feedback to inhibit oscillation except at the peak of the response; the inverted network is balanced against positive feedback, the latter dominating only at the trough in the network's response.

Re-drawing this inverted network gives the bridged-T. The nullor gives no information on the phasing of input-output to be used in practical amplifiers that implement it. A class of circuits called negative impedance convertors (n.i.cs) show the advantages and snags. Resistive feedback across an amplifier with positive voltage gain leads to a negative input resistance $R_i = v/i = v/-(v/R) = -R$. If the input phasing is reversed the input impedance is still found to be $-R$, but is now stable only for source resistances $>R$ instead of $<R$. Extracting the amplifier and feedback resistances leads to a form of n.i.c. which can be shown, generalized to the form shown, where $Z_i = -Z_1Z_3/Z_2$. This allows negative resistances, capacitances and inductances to be synthesized using only Rs and Cs. If a n.i.c. is combined with parallel and series RC networks which cancel at a single frequency, oscillations result.

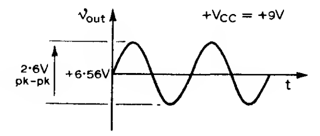


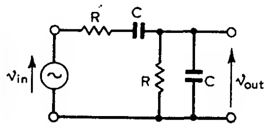


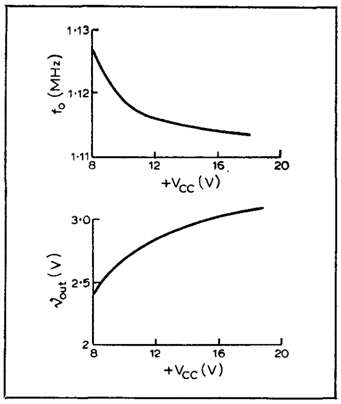
Current-driven Wien oscillator

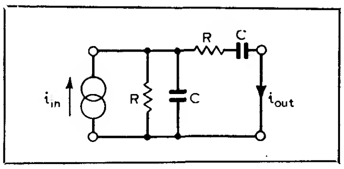


Typical performance
+Vcc +9V, 2.7mA
R 2.2kΩ, C 56pF
R_1 4.7kΩ, R_2 1.5kΩ
R_3 470Ω, R_4 470Ω
Tr_1, Tr_2 1/5×CA3086
(Note: Tr_1 emitter is pin 13)

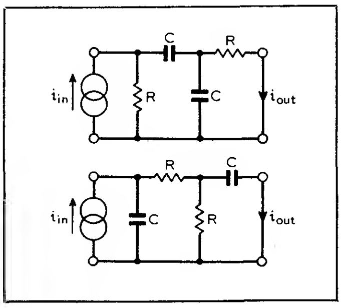




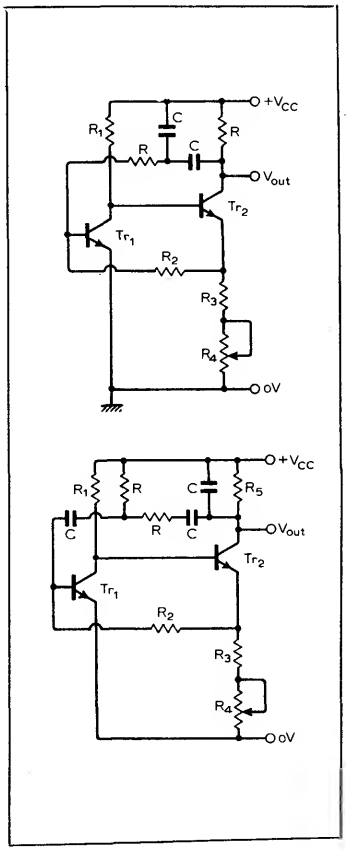




Circuit description
Many RC oscillators use a Wien network as the frequency-determining part of the closed loop and very often the Wien network is used where it may be driven from a low-impedance source and loaded by a high-impedance load as shown right.
However, in many applications it is more convenient to use the current dual of the above network which is driven from a current source and ideally is loaded by a short circuit. The reciprocity principle requires that this network has the same transfer function as the voltage-driven form, which is $V_{out}=V_{in}Z_2/(Z_1+Z_2)$ where Z_1 is the impedance of the series-connected RC pair and Z_2 is the impedance of the shunt-connected RC pair. The resulting dual network is then as shown top right.
The transfer function of this network is given by $i_{out}=i_{in}Z_2/(Z_1+Z_2)$.
The circuit shown above left is an RC oscillator employing the current-dual network to close the loop of a bipolar transistor amplifier which is in the form of a d.c. feedback pair. The Wien network is supplied from the collector of Tr_2 which serves as a reasonably good current source due to the increase in output impedance provided by the feedback network. The output of the Wien network is loaded by the d.c. feedback network in shunt with the input impedance at the base Tr_1. Whilst not an ideal load negative feedback through R_2 provides a sufficient approximation to the desired low value.
In order to provide sustained oscillations the closed-loop phase-shift must be zero and the amplifier must provide sufficient gain to overcome the losses in the Wien network. For the usual case of equal Rs and equal Cs in the series and shunt parts of the Wien network the frequency of zero phase shift occurs at $f_0=1/2\pi CR$ and at this frequency $i_{out}=i_{in}/3$. Hence, the minimum gain required from the amplifier is 3. In the d.c. feedback pair shown this current gain is given approximately by $A_i=1+R_2/(R_3+R_4)$. Resistor R_4 is provided in the form of a variable to allow convenient adjustment of the loop gain to provide a reasonably-sinusoidal output waveform. The more the gain exceeds the critical value the more distorted will the output waveform become and the lower will be the frequency of oscillation. (See graphs for effect of increasing supply voltage, for example, after setting the critical condition at a lower value of +Vcc.)



Component changes
Useful range of Vcc +4 to +20V. Scale RC values for lower-frequency operation. For higher oscillation frequencies, f_0 will become less-accurately predictable due to presence of internal transistor capacitance and any shunt capacitance of load. R_4 will need re-adjustment if operating frequency is changed.
Separate transistors may be used or the whole circuit excluding capacitors and R_4 integrated on a single monolithic chip.

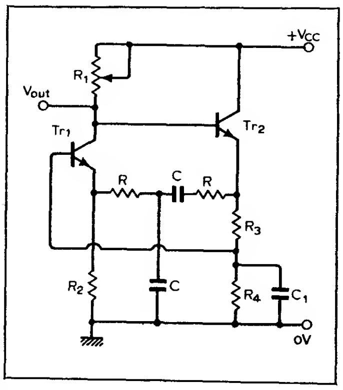


Circuit modifications
- Any other frequency-determining network which operates with an optimum source impedance tending to infinity (a current source) and an optimum load impedance tending to zero can be used in place of that shown.
Two examples are given.
- D.c. feedback pair forms of these networks are shown below.
C, R and R_1 to R_4 same values as original circuit. X_{C1} tends to zero at f_0 and R_5 is large, say 10kΩ.

Further reading

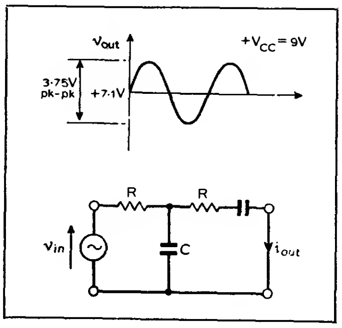
Williams, P. Wien Oscillators, *Wireless World*, November 1971, pp. 541-6.
Stott, C. Transistor RC Oscillator, *Wireless World*, February 1962, pp. 91-4.

RC oscillators



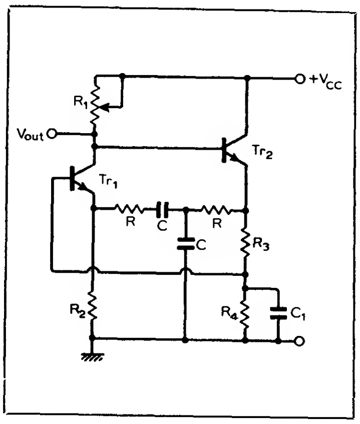
Circuit description
The above oscillator circuit uses a Wien-network which is fed from an emitter-follower Tr_2 and is loaded by a common-base Tr_1. This realization is an example of an RC oscillator which has the Wien-network ideally fed from a voltage source and loaded by a short-circuit as shown below. At the frequency of oscillation $f_0=1/2\pi CR$ Hz, the loop phase shift is zero and the trans-admittance of this frequency-determining network is $(1/3R)$ siemens. Hence, to sustain impedance of the common-base stage only an approximation to a short-circuit load on the network. However, as R_2 is much greater than the common-base stage input impedance, virtually all of the output current from the RC network enters the emitter of Tr_1. If Tr_1 has a reasonably large current gain the base current be neglected to a first approximation, so that the collector current in R_1 is virtually equal to the output current from the RC network. Thus, assuming that the voltage gain of the emitter follower is only slightly less than unity, the amplifier gain $A_z=V_{out}/i_{in}$ is essentially equal to R_1. The circuit will oscillate, theoretically, provided that R_1 is made three times the value of R with C chosen to determine the frequency of oscillation.

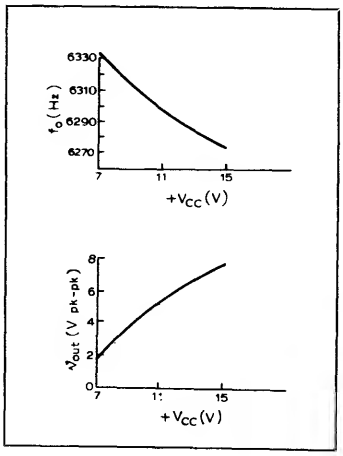
Typical performance
$+V_{cc}$ +9V
R 4.7kΩ, C 4.7nF
R_1 50kΩ var. (typically 15kΩ)
R_2 22kΩ, R_4, R_3 2.7kΩ
C_1 2.2μF
Tr_1, Tr_2 1/5×CA3086
(Note Tr_1 emitter is pin 13)

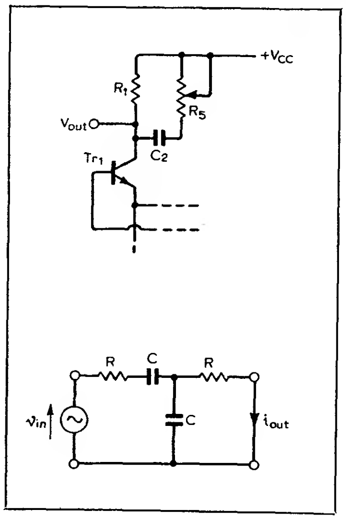


In the above circuit, for example, where R was selected oscillations the amplifier must provide a current-to-voltage gain (A_z) of $3R$ ohms to make the closed-loop gain unity i.e. $\beta_y.A_z=1$.
In the circuit arrangement the emitter follower is only an approximation to the ideal voltage source to feed the RC network and the input as 4.7kΩ, R_1 should be $3\times4.7k\Omega=14.1k\Omega$ and in practice, using 5% tolerance resistors, the circuit just oscillated with $R_1\approx15k\Omega$.

Component changes
The useful range of supply voltage is approximately +4 to +20V, but note that changing the supply will change the operating currents and hence







the overall gain of the amplifier. Hence, large changes in supply voltage will change the loop gain sufficiently to either prevent oscillation or to severely distort the output waveform. For example, in the above circuit the oscillations cease at V_{cc} of +6V (having adjusted R_1 with V_{cc} of +9V) and the output waveform becomes clipped on positive peaks when V_{cc} exceeds about +11V. Thus, change of V_{cc} will normally require a readjustment of R_1 for a reasonable sinewave output waveform.
Scale RC values for different oscillation frequencies but note that C scaling only allows R_1: R ratio to be maintained without changing R_1 significantly.

Circuit modifications
- If it is desired to change the frequency of oscillation by changing both the R and C values, but without changing the d.c. operating conditions, then R_1 can be fixed and the a.c. closed-loop gain adjusted by the arrangement shown.

From an a.c. viewpoint R_5 is in shunt with R_1 provided $X_{C2}<<R_5$ at the frequency of oscillation.
- Another network having the same transfer function as that shown already at the frequency of oscillation is shown on the left.

Again the frequency of oscillation is given by $f_0=1/2\pi CR$ Hz and the required amplifier gain is again $A_z=R_1$ in the ideal case. A practical realization is shown bottom again using an emitter follower and a common-base stage.
The same component values may be used for the same frequency of oscillation.

Further reading
Williams, P. Wien oscillators, *Wireless World*, November 1971, pp. 541-6.
Industrial Circuit Handbook, SGS-Fairchild 1967, pp. 42/3.

Op-amp Wien oscillator

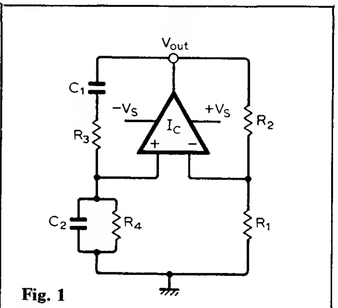
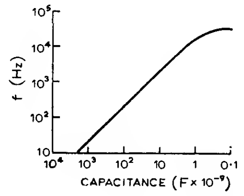
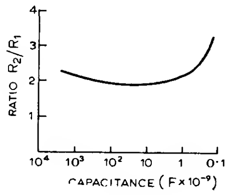


Fig. 1

Typical data
±Vs ±15V
IC 741
R3, R4 4.7kΩ ±5%
C1, C2 0.1μF
Frequency 355Hz
R1 3.4kΩ, R2 6.6kΩ
Harmonic distortion 0.4%
Output 26V pk-pk up to 14kHz
Drops to 16V pk-pk at 27kHz





Circuit description
This is a single-frequency oscillator circuit which may be envisaged as a bridge network, where oscillation will occur when the bridge is balanced and the differential input to the amplifier is near zero. Positive feedback is applied via two reactive arms, and negative feedback via the resistor potential divider R_1, R_2. The gain of this loop is $(R_2+R_1)/R_1$. The gain of the positive feedback loop is real at a positive loop frequency that makes $A_+=1+R_3/R_4+C_2/C_1$ and if $R_3=R_4$, $C_2=C_1$, then $A_+=3$. Oscillation then occurs when $A_-=3$, or when $R_2\approx 2R_1$. This oscillatory condition will require a minimum distortion to maintain a stable output. However, if the A_+ gain falls below this value, oscillation will stop due to the negative feedback being more than the positive feedback, and if the gain increases, the output amplitude will increase until it is limited by non-linear distortion. This could be the power supply limitations or additional network limiting. In general the frequency of oscillation is given by $f=1/2\pi\sqrt{C_1C_2R_3R_4}$ Hz for C in farads, and R in ohms.

Component changes
- Varying $C_1(=C_2)$ over the range of 4.7μF to 220pF provides frequency range above. (Slew rate-limiting of the op-amp causes fall-off at lower C values.)

- Simpler technique maintains C constant, but demands ganged potentiometer for R_3 and R_4 for frequency adjustment with single control, i.e. ratio R_3: R_4 is maintained constant.

- Amplitude limiting is available using either of the additions shown in Fig. 2.

- Nominally R_5 should be much greater than R_2, and R_5 in parallel with R_2 should be less than $2R_1$ when the diodes are conducting (see Fig. 2).
R_1 3.3kΩ, R_2 8.62kΩ
R_5 22kΩ, frequency 354±1Hz
Vs ±12V
Amplitude is stable then for power supply increases up to ±18V. With diode or zener diode limiting, the frequency is more independent of the power supply, but is more dependent on the break level of the limiting network. Another advantage is that instantaneous oscillation is available at low frequencies. With no limiting, time for build up of amplitude is frequency dependent.
- Note that at high frequencies, since gain of amplifier is finite, it is more noticeable that a small differential input must exist, which demands an adjustment of the R_2: R_1 ratio (see second graph).

Circuit modifications
Circuits above, Figs. 3 to 5, provide exactly the same performance as the basic circuit. Notice that the output and ground terminals are shifted depending on op-amp connection (see Nullors, card 1).
With limiting network, typical performance given below.
Diode limiting for circuits 1, 3, 4, 5.
Outputs: 16.8, 16.8, 16, 15.6V pk-pk, harmonic distortion: 0.95%, 0.95%, 0.9%, 0.89%, respectively.
R_2 6.39kΩ, R_1 3.61kΩ for R_5 22kΩ.
Zener limiting for circuits 1, 3, 4, 5.
Outputs: 21.5, 26.5, 26.5, 19V pk-pk, frequency: 350, 352, 352, 352Hz, harmonic distortion 5, 5.9, 5.9, 5.1%, respectively.
R_2 6.9kΩ, R_1
In general, a trade can be made between distortion and amplitude stability.

Further reading
Williams, P. Wien Oscillators, *Wireless World*, Nov. 1971, pp. 541-6.

Cross references
Set 25, cards 6, 5.
Set 21, card 4.

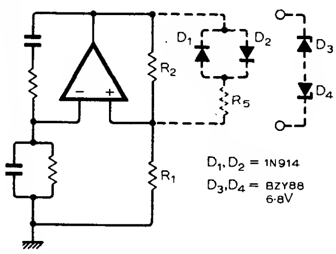


Fig. 2

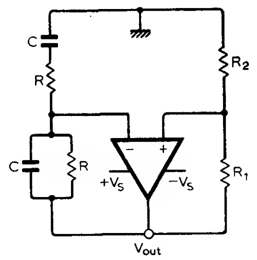


Fig. 3

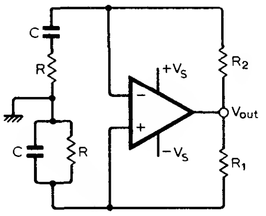


Fig. 4

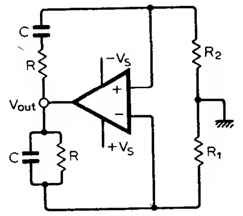
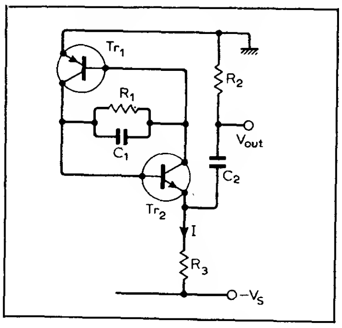
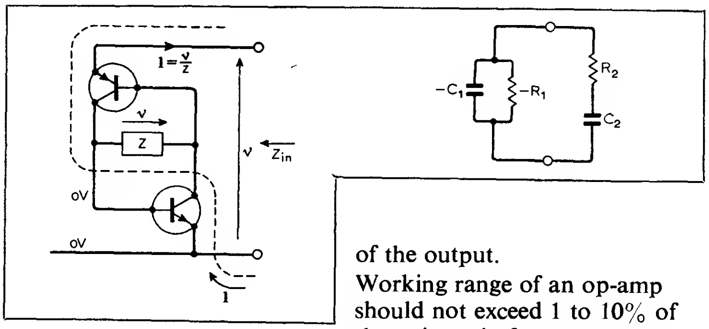


Fig. 5

Micropower oscillator



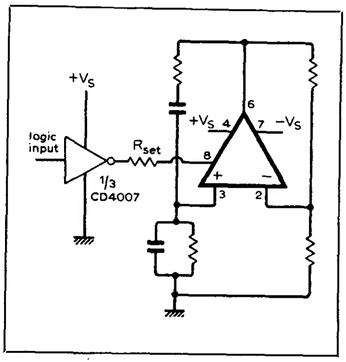
Typical data
$-V_S$ -20V, R_3 220k
This combination approximates a constant current source
I 86μA
R_1 3.85kΩ, R_2 1.5k $\pm 5\%$
C_1 100nF, C_2 220nF
V_{out}: 54mV pk-pk
frequency 436Hz
Tr_1 BC126, Tr_2 BC125



Circuit description
This is one form of a negative impedance converter circuit. It may be considered as a two terminal device, and be inserted in a constant current path that may already exist in a circuit. The circuit will provide a low voltage amplitude sinusoidal oscillation (but not necessary low distortion) with a minimum of components, no biasing being necessary but the choice of components is critical. Provided the appropriate gain and phase shift are possible, transient analysis shows that oscillations will build up exponentially in such a network after shock excitation, if the circuit losses are negligible. This is achieved by presenting a negative resistance across an existing circuit resistance. Oscillations are sustained when the average value of the negative resistance provides the correct ratio for the frequency chosen.
In the above circuit, V_{be} of the transistor is neglected, and hence the direction of current through the impedance Z must be as shown: through the collectors. The impedance presented at the input terminals, then is $Z_{in}=v/(-v/Z)=-Z$. The loop impedance for the given network is:

$$R_2+\frac{1}{j\omega C_2}+\frac{-R_1/j\omega C_1}{-R_1-1/j\omega C_1}$$

The numerator reduces to $1-\omega^2C_1R_1C_2R_2+j(\omega C_1R_1+\omega C_2R_2-\omega C_2R_1)$. For oscillation, the loop impedance should be onset of visible distortion. Circuit will operate at lower currents but with lower output —see graph.
Typical data for absolute minimum current of I 5.8μA, R_1 13kΩ, R_2 8kΩ, C_1 10nF, C_2 22nF, R_3 3.3MΩ, V_S -20V. To obtain low currents demands scaling of



zero giving $C_1/C_2+R_2/R_1=1$ or $R_1=2R_2$, $C_1=C_2/2$ and $f=1/2\pi\sqrt{C_1R_1C_2R_2}$.

Component changes
Minimum value of R_1 2.73kΩ
Frequency 320Hz
Variation of output with R_1 shown over.
R_1 maximum 2.9kΩ before resistors for optimum performance.

Circuit modification
- Micropower operation with the classical Wien network (this series, card 4) is possible with the LM4250 programmable operational amplifier. For a fixed dual polarity supply, one external resistor determines the quiescent current and consequently the slew rate and gain-bandwidth product. Voltage range is ±1 to ±18V. Care is necessary in the choice of frequency in relation to the programmed set-current because of the slew-rate and gain-bandwidth dependence. For a cisoidal waveform

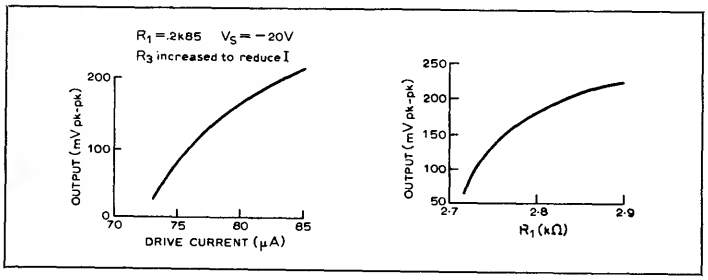
$$\text{frequency (Hz)}=\frac{\text{slew rate}}{2\pi V_{max}}\text{(V/s)}$$

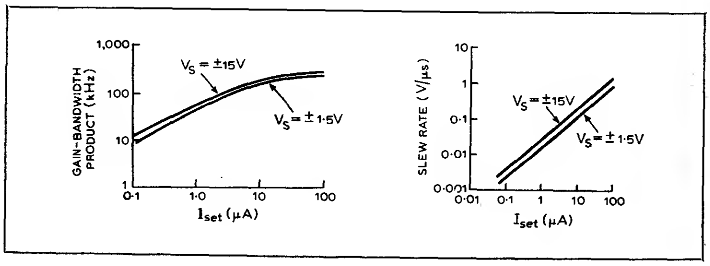
where V_{max} is the peak value of the output.
Working range of an op-amp should not exceed 1 to 10% of the unity gain frequency to avoid severe unbalance at the bridge input, i.e. at $I_{set}=1\mu$A, gain-bandwidth product is 70kHz. If oscillator frequency is 700Hz, amplifier gain is 100, and hence input required is 1% of output. If set current is reduced to 0.1μA, then for same frequency gain is only 10. Hence for the same output greater unbalance required at the input.
- Normal diode limiting possible at much lower currents.
- High values of resistance usable to minimize power drain, however stray-capacitance effects may then be significant for designed frequency. High input impedance of op-amp will not load bridge arms.

Oscillator may be gated using a c.m.o.s. inverter to sink the programmed set-current, shown left.

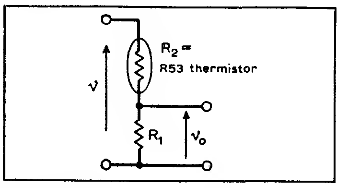
Further reading
Braun, J. Equivalent n.i.c. networks, nullators and norators, *IEEE Trans.* vol. CT-14, 1967.
Linear Integrated Circuits, LM4250, National Semiconductor.

Cross references
Set 10, card 8.
Set 25, cards 8, 1, 4.

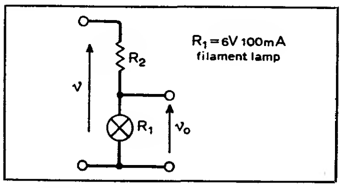




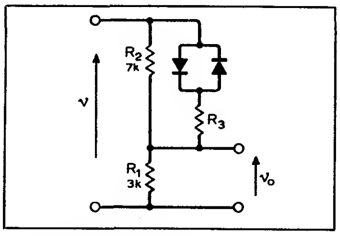
Amplitude-control methods



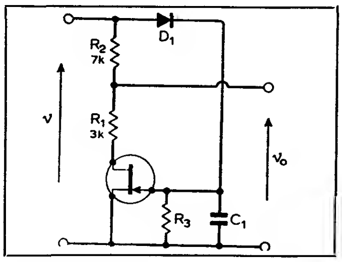
Thermistor
A bulk semiconducting resistor with negative temperature coefficient. Placed in series with a resistor, and with an increasing voltage applied, the thermistor is heated by the resulting flow of current; its resistance falls and the attenuation of the network decreases. There will be a single amplitude at which a desired attenuation is achieved. If the network is incorporated into an oscillator, where the frequency-determining network has zero phase shift together with the same attenuation (at a particular frequency) then oscillations will be sustained.



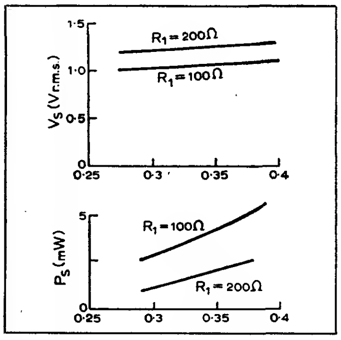
Lamp
Place in series with a resistor which is low enough to permit sufficient current to heat the lamp. As the voltage increases, the lamp resistance rises and again the attenuation decreases. The lamp can operate at a higher temperature than the thermistor and is less sensitive to ambient temperature changes. Its power consumption is considerably higher and the sensitivity to amplitude changes markedly less (dissipation for significant temperature rise >50mW for most lamps—as little as 3mW for thermistors designed for this application).



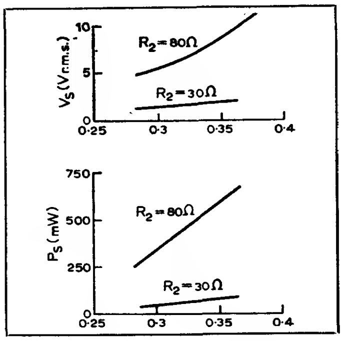
Non-linear network
Self-sustaining oscillation can also be achieved by passing the feedback through a non-linear network. At low amplitudes the diodes are non-conducting and the attenuation is determined by R_1 and R_2. For larger peaks, the diodes conduct and R_3 appears in parallel with R_3 increasing the attenuation. There will be a single amplitude of input at which the average value of the output meets the condition. There is a compromise between distortion and sensitivity of the amplitude to small changes in the resistor values.



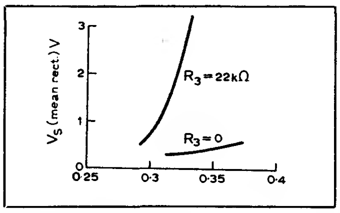
Peak rectifier
A field-effect transistor has an output slope resistance that is moderately linear and is controlled by the gate-source voltage. If the amplitude of the input increases, the peak rectifier has an increased output. It is applied to the gate of a p-channel junction f.e.t. reverse biasing it and increasing its resistance. This decreases the a.c. attenuation of the network. The range is small as the f.e.t. cannot accept a large drain-source voltage in this mode without increasing the distortion.



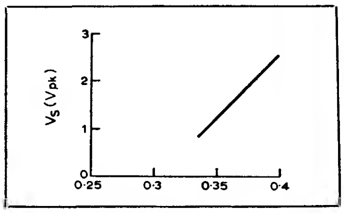
Thermistor
Maximum device dissipation of 3mW for the quoted device changes its resistance by two orders of magnitude. The rise in temperature is relatively small and the final amplitude when used in an oscillator is slightly temperature dependent. A 25% change in the transfer function of the oscillator (due to either passive components or fall in amplifier gain at high frequencies) would be accommodated by an amplitude change of as little as 2% when oscillations re-stabilize.



Lamp
For the same change in the oscillator network (25%) the amplitude using a lamp might only be stabilized to within 20-40%. The power consumption is 10-100 times greater making it more difficult to choose a suitable amplifier. The lamp costs less and, operating at a higher temperature, is less affected by ambient changes. To extend the lamp life it is advisable to limit the power input which brings its performance somewhat nearer to that of a thermistor.



Non-linear network
The ratio R_2: R_1 is set to give an attenuation less than that needed to inhibit oscillation. The diodes begin to conduct when the peak voltage across R_2 exceeds 0.6V. At some higher voltage the attenuation exceeds the critical value. With $R_3=0$ this state is rapidly reached and the amplitude is controlled within reasonable limits— 50% change for a change in the oscillator network of 15%. With R_3 of 22kΩ the amplitude has to change by a much larger amount to cope with a corresponding change in the network—in this example a greater than 5: 1 range. In return the resulting distortion is much lower.



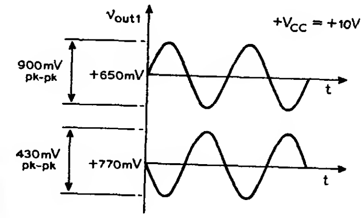
Peak-rectifier/f.e.t.
The network consisting of D_1, C_1, R_3 feeds the gate of the f.e.t. with a voltage equal to the peak input less the diode forward drop. For input amplitudes above 0.6V peak the f.e.t. becomes reverse-biased with a drain-source resistance increasing from about 500Ω to many times that value. At high values of reverse bias, the range of drain-source voltages for low distortion is reduced. This conflicts with the increased p.ds across the network that cause the bias and limits the range attenuations that can be achieved—a range of about 12% in this case which is sufficient to cope with the usual spreads.

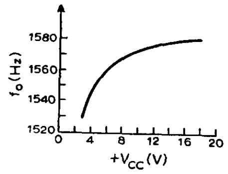
Baxandall RC oscillator

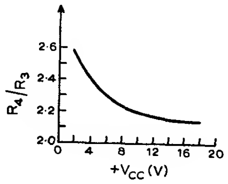
Circuit description
The circuit below is a bipolar transistor version of a Wien network oscillator due to Baxandall. In comparison with its more common operational amplifier form it has the merit of simplicity. The usual negative supply rail for the operational amplifiers may be dispensed with, and only two resistors (R_1 and R_2) are required in addition to the Wien bridge resistors (R_3 and R_4) and frequency-determining components (R and C). Also, the Wien network resistors are arranged so that the circuit is self-biasing. This circuit simplicity is achieved at the expense of departure of the frequency of oscillation from the ideal value of $f_0=1/2\pi RC$ Hz.
In the ideal case, the circuit would just oscillate when $R_4/R_3=2$ but the finite gains in the transistors cause the critical condition to be achieved with an R_4/R_3 ratio somewhat in excess of this value. Also, as the supply voltage is varied the operating conditions of the transistors, and hence their gains, change which will require readjustment of the R_4: R_3 ratio either to sustain oscillation or to restore the output waveforms to reasonably undistorted sinusoids. Although, with the component values shown, the output waveforms from Tr_1 and Tr_2 collectors are of different magnitudes, they have the useful feature of being in antiphase.

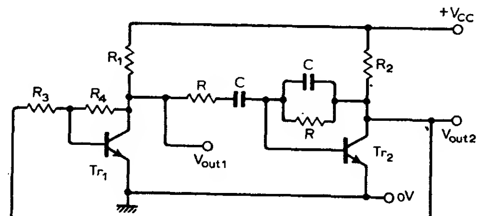
Component changes
Useful supply range +3 to +20V
Useful range of R 100 to 100kΩ
Useful range of C 1n to 10μF
With given R values the frequency of oscillation can be changed without changing the d.c. conditions by changing the capacitor values only.

Circuit modifications
The output waveforms from Tr_1 and Tr_2 collectors can be made of equal amplitude as well as being in antiphase, without changing the frequency of oscillation, by scaling the frequency-determining RC components of the Wien network as shown below. With this arrangement the impedances of the series and parallel RC components are the same (R_2) at the frequency of oscillation so that ideally, the gain from Tr_1 collector to Tr_2 collector is −1. Hence, in the ideal case $R_4/R_5=1$ for the critical oscillation condition; i.e. the gain from Tr_2 collector to Tr_1 collector is also −1 giving a closed-loop gain of +1. In practice, with the circuit values R 10kΩ, C 10nF the frequency of oscillation was virtually unchanged, and R_4/R_3 was typically 1.1 to sustain oscillations with V_{out1} and V_{out2} typically 1 volt pk-pk and 980mV pk-pk respectively. A closer approach to the ideal state can be obtained by using two inverting operational amplifiers in place of Tr_1 and Tr_2 as shown below.
R, C, R_3 and R_4 are direct replacements from the bipolar transistor version and A1 and A2, which could be 741s, replace Tr_1 and Tr_2 respectively. The same scaling of the R and C values to provide equal-amplitude outputs may be used as before. These outputs are now available from low output impedance points.
As with other oscillator circuits, a bandpass filter may be obtained by setting the gain (by adjusting the ratio R_4: R_3) just below the value needed to produce oscillation and injecting the signal through a reasonably high resistance at the inverting input to A_1. This can also be achieved with the bipolar transistor version by injecting the signal at Tr_1 base, where the need for a high impedance source is more stringent since this point is an imperfect virtual earth.





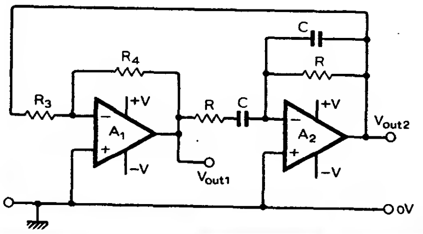


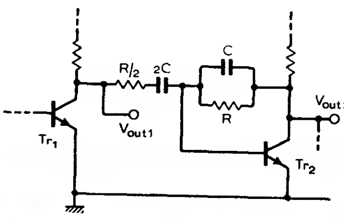


Typical performance
+V_{cc}+10V, 2mA
R_1, R_2, R 10kΩ, C 10nF
R_3+R_4 10kΩ (realized with 10kΩ potentiometer to adjust loop gain)
Tr_1, Tr_2 1/5×CA3086
(Note Tr_1 or Tr_2 emitter is pin 13)

Further reading
Williams, P. Wien Oscillators, *Wireless World*, Nov. 1971, pp. 541-7.

Cross references
Set 1, card 3.
Set 10, card 8.





wireless world circard

N.I.C. oscillators

Circuit description

A negative impedance converter is a two-port active device which if loaded at one port with an impedance Z, can provide an impedance $-nZ$ at the other port. If the impedances are frequency dependent, then a frequency constraint exists, for at only one single frequency the serial and parallel RC networks have equivalent magnitudes and the appropriate phase angle to the make the loop phase shift zero. The Fig. 1 circuit is of similar form to Fig. 2 for which an approximate analysis is given. Assume the base-emitter voltages are negligible. Then the signal v_{in} will appear as v_{out}. The input signal current i_{in} comprises v_{in}/R_1 and $i_{in}-v_{in}/R_1$ up through Tr_1. The voltage across R_1 is then $i_{in}R_1-v_{in}$ which must be that across R_2. Therefore the emitter, thus collector, current of Tr_2 upwards is $i_{in}R_1/R_2 \; -v_{in}/R_2$. Thus i_{out} is $-(v_{in}/R_2+i_{in}R_1/R_2 \; -v_{in}/R_2)=-i_{in}R_1/R_2$. If an impedance Z is across the V_{out} terminals, then

$$Z_{in}=v_{in}/i_{in}=v_{out}/i_{in}$$

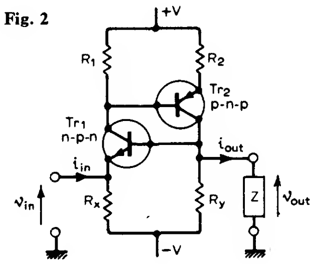
$$=-\frac{v_{out}}{i_{out}R_2/R_1}=-\frac{R_1Z}{R_2}$$

If $R_1=R_2$, $Z_{in}=-Z$.

Typical data

Tr_1 BC126, Tr_2 BC125
R_1, R_2 10kΩ, R_3 5kΩ
R_4 4.66kΩ R_5, R_6 4.7kΩ
C_1, C_2 0.1μF
V_s ±5V
Ratio required of R_2: R_4 slightly greater than two to allow for component tolerances
Frequency 373Hz (maximum 20kHz)
v_{out} 1.55V pk-pk, 0.3V offset

Fig. 2



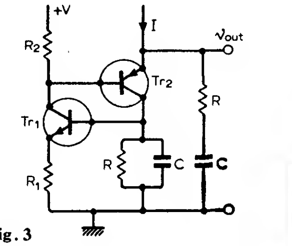
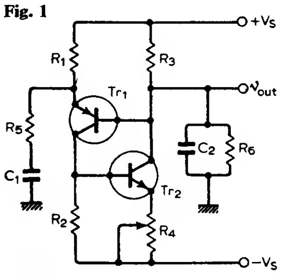
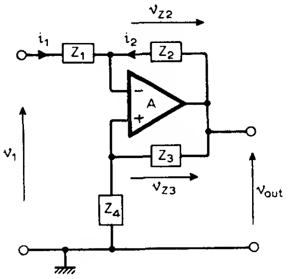
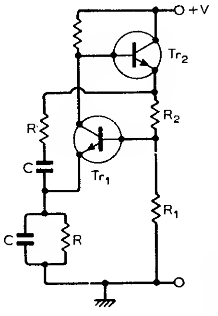


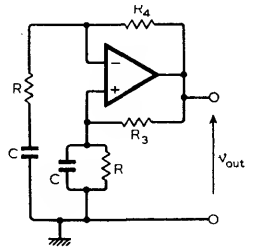
Fig. 3

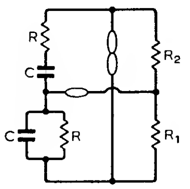
Fig. 1

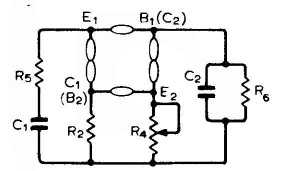


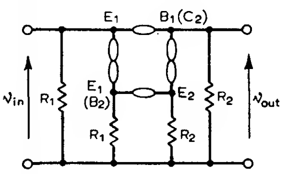


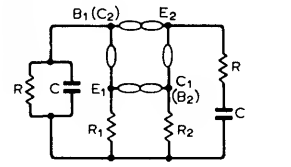












Component changes

Another oscillator using a different n.i.c. is shown above. Separate voltage and current supplies are necessary, but low level operation is possible. Typically, for V of 1V, I of 70μA, circuit will oscillate ($f=1/2\pi RC$) to provide a peak output of about 1.2V. An advantage of these circuits is that the frequency-tuning networks have a common ground point, and a minimum of components is used. When these circuits are analysed using nullor concepts (card 7), the similarity is more obvious. The equivalent representations are given in the circuits showing nullator, norator interchange. Bridge oscillators using operational amplifiers (card 4) can also be considered as a form of n.i.c. oscillator. For the above general case

$v_{Z3}=v_oZ_3/(Z_3+Z_4)=v_{Z2}$

$i_2=v_{Z2}/Z_2=v_oZ_3/(Z_3+Z_1)Z_2-i_1$

From Millman's theorem,

$v_-= v_1y_1+v_oy_2/(y_1+y_2)$

$v_+= v_oy_3/(y_3+y_3)$

For high A, $v_-=v_+$ and

$$\frac{v_1}{v_o}=\left(\frac{y_1+y_2}{y_3+y_4}\right)\frac{y_3}{y_1}-\frac{y_2}{y_1}$$

Also $Z_{in}=v_1/i_1=$

$$-\frac{v}{v_o}\left(\frac{Z_3+Z_4}{Z_3}\right)Z_2$$

If $Y_2=Y_3$, then $Z_{in}=Z_1-Z_4$

If $Z_1=0$ $Z_{in}=-Z_4$

A parallel tuned circuit of dynamic impedance L/Cr connected across the input terminals would be the basis for an oscillator, if the magnitude of Z_4 is made equal to r. Specifically, $Z_{in}=-Z_4.Z_2/Z_3$. Hence if Z_4 is a parallel RC network, then a series RC network across the input, with an appropriate ratio of R_4: R_3 provides one form of Wien bridge oscillator. The nullor concept allows certain nullor/norator interchanges which provide an alternative Wien shown below, which again can be analysed from a n.i.c. concept. A discrete version is shown above, where if the transistors gains are high, the ratio R_2:R_1 approaches two.

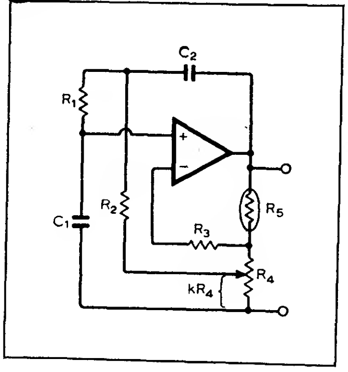
Further reading

Pasupathy, S. Transistor RC oscillator using negative impedances, *Electronic Engineering*, December 1966.
Newcomb, R. W. Active integrated circuit analysis, Prentice-Hall, 1968.
Pasupathy, S. Equivalence of LC and RC oscillators, *Int. J. Electronics*, vol. 34, no. 6, pp. 855-7.
William, P. Wien oscillators, *Wireless World*, Nov. 1971, pp. 541-6.

Cross references

Set 25, cards 1, 5.

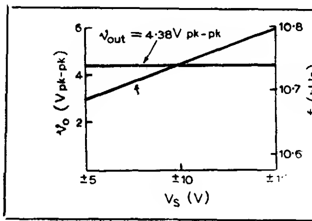
Single-element-control oscillators—1

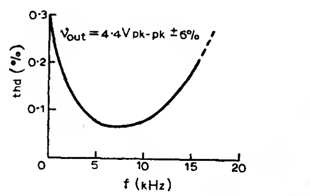


Typical performance
IC 741
Supplies ±15V
R_1, R_2 (=R) 10kΩ
R_3 15kΩ, R_4 1kΩ log
R_5 ITT thermistor R24
C_1, C_2 (=C) 1nF
f 1.5kHz to 20kHz

$$f=\frac{\sqrt{1-k}}{2\pi CR}$$

provided $R_2 \gg R_4$





Circuit description
As described earlier the lead/lag network, C_2R_2 followed by C_1R_1, is an alternative to the Wien network. If the network is driven by two amplifiers (see ref.) then it is possible to vary the frequency of oscillation by changing the gain of one amplifier, without changing the condition for sustaining the oscillations. This simplifies the amplitude control circuitry but still requires two amplifiers, one with variable gain. If instead the impedance level of the frequency dependent network is much greater than the resistance of the amplitude-controlling network, one amplifier can be eliminated. Resistor R_2 is tapped onto R_4. As k is reduced to zero the frequency of oscillation reverts to that of the basic Wien bridge/lead-lag oscillator viz. $1/2\pi RC$. As $k \to 1$ the frequency $\to 0$. This leads to a practical range of frequencies in excess of 10: 1 on a single control without any serious increase in distortion. The component count is comparable to that with conventional oscillators, but the need for a twin-gang control which normally requires a good match is avoided. A further advantage of this circuit is that low frequencies are obtained without using large values of capacitance. Provided the thermistor has a long enough thermal time constant, frequencies down to 10Hz are possible with capacitances of 0.1μF.

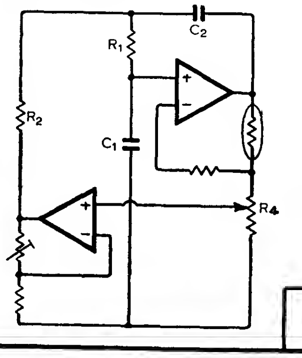
Component changes
IC: any general-purpose compensated op-amp. For lower frequencies, f.e.t. input op-amps allow the use of larger resistors. This also minimizes the loading on the potentiometer and widens the range that can be covered by variation of k alone.
Supplies: not critical. Should be appreciably greater than required peak-peak output if rapid thermal stability of thermistor to be achieved. Typically ±6 to ±15V.
C_1C_2: to suit frequency range, but 220pF to 1μF possible. Normally $C_1=C_2=C$.
R_1, R_2: 1k to 1MΩ. High values only possible if f.e.t. op-amp available, e.g. CA3130. Allows very low frequency with suitable amplitude control mechanism. Normally $R_1=R_2=R$.
R_4: selected to suit thermistor. $R_4 \ll R$, typically 100 to 470Ω.
R_3: not critical. Minimizes offset due to unbalanced input currents. Excessive offset disturbs amplitude control by permanently heating thermistor.

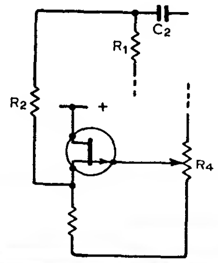
Circuit modifications
- Adding a unity gain buffer amplifier between R_4 and R_2 removes interaction permitting wider range for fixed Cs. End-resistance effects on R_4 prevent $k \to 1$ and a voltage gain >1 in the buffer stage corrects for this. Some versions of this circuit can achieve a range in excess of 100: 1 on a single potentiometer.
- A simpler buffer may suffice in some applications where only the impedance levels are critical. A source follower removes the loading on R_4 though the maximum buffer gain is unlikely to exceed 0.9, i.e. range of frequencies is restricted.
- The original circuit again has a related form in which output and ground on the bridge are interchanged as are the inverting and non-inverting inputs. This is in line with the results on the basic Wien bridge oscillator (cards 2, 3, 4). Performance is basically similar to the original. The advantages of alternative configurations are that different components are grounded which can simplify frequency and amplitude control.
The nullor form is shown. It gives no information on phasing can make it easier to generate new versions. An alternative viewpoint is that of shifting the groundpoint in the system.
- Further oscillators (not shown) interchange the locations of the Cs and Rs to make lag-lead oscillators. These can be used with inverting buffer amplifiers where it is required to reduce the frequency of oscillation below the basic value $1/2\pi CR$.

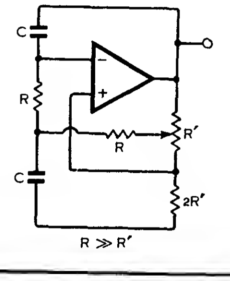
Further reading
Sun, Y. Generation of sinusoidal voltage (current) controlled oscillators for integrated circuits, *IEEE Trans.* 1972, CT-19, pp. 322-8.

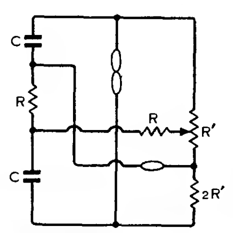
Cross reference
Set 25, card 10.

Circuit modifications

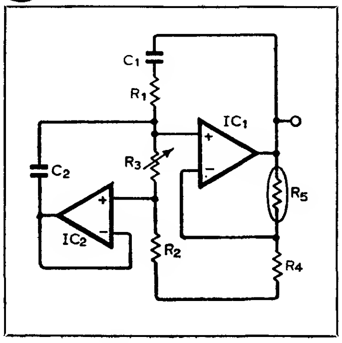






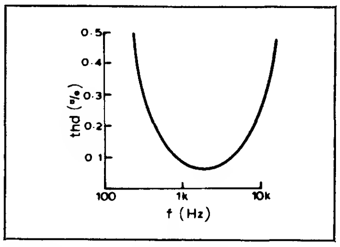


Single-element-control oscillators—2



Typical performance
$IC_{1,2}$ 741
Supplies ±15V
R_1, R_2 2.2kΩ
C_1, C_2 0.047μF
R_3 2MΩ log
R_4 470Ω
R_5 ITT thermistor R54
At f 1kHz, v_o 6.4V pk-pk
T.h.d. 0.1%
f_{max} >18kHz as $R_3 \to 0$
f_{min} <120Hz as $R_3 \to 2M\Omega$

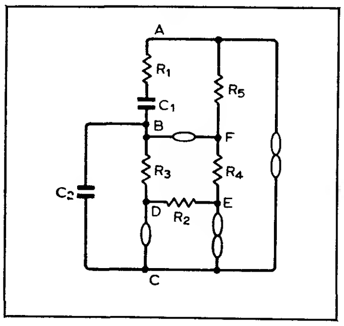
Circuit description
A circuit given by Brokaw (see ref.) used a modified form of Wien bridge oscillator. In it, one of the frequency determining resistors is varied, while a second amplifier has a variable gain controlled by that same resistor. The form of circuit used (see over) gave a wide range of frequencies on a single control with no change in the amplitude control condition. By drawing the nullor equivalent circuit the alternative form was found in which one of the amplifiers was used as a voltage follower. This allows the substitution of specially optimized voltage followers such as the LM310 to minimize errors in this stage. For even simpler circuits less demanding of performance



emitter or source-followers can be used. The frequency of oscillation is $1/2\pi\sqrt{R_1R_3C_1C_2}$ provided that $R_2=R_1$ and the amplitude-maintaining condition $R_5=R_4$ is maintained.

N.B. R_3 can be replaced by any other element or device that behaves as a linear resistor. If a photo-conductive cell is used the frequency becomes light sensitive and a wide range of light intensities can be covered. Oscillator amplitude stability ±1% from 120Hz to 15kHz, V_s ±10 to ±15V.

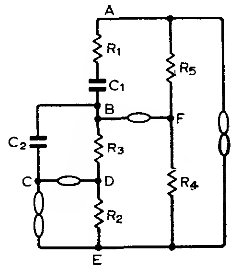
Component changes
IC_1: any compensated op-amp. For low frequency operations, the increase in R_3 results in greater errors due to op-amp input currents, and f.e.t. input stages would be better. As indicated above, 100 : 1 range is readily obtained with general purpose units.
IC_2 can be replaced by a voltage follower, source follower, Darlington-connected emitter-follower etc. A useful feature is that d.c. offset in this stage has no effect.
Supplies: Not critical. Usual range 5V to ±15V for op-amps. Restriction mainly placed by choice of amplitude control devices/circuits.
C_1, C_2 to suit frequency range but typically 1nF to 1μF. $C_1=C_2$.
R_1 1k to 100kΩ.
R_3 Range should be wide if frequency range is to be great since $f \propto 1/\sqrt{R_3}$. Typically R_3 may range from 100Ω to >1MΩ.
R_2 For $C_1=C_2$, $R_4=R_5$ for amplitude control and $R_2=R_1$ is the remaining condition.
R_4 chosen to suit the particular thermistor used—see manufacturer's data.

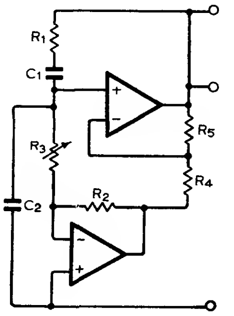


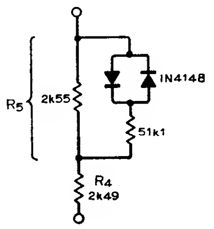
Circuit modifications
• The input of each amplifier is replaced by a nullator, the output-ground port by a norator and the circuit is redrawn in this nullor form in Fig. 1. Points C, E and A may be at arbitrary relative potentials because of these norators. One alternative configuration having the same property is Fig. 2. This can also be interpreted as a shift of ground point from E to C. Re-pairing of these nullators and norators leads to the practical circuit of Fig. 3 given in a recent reference. It is a high performance oscillator with a frequency range 200: 1 but again uses f.e.t. input amplifiers. A number of other oscillators can be similarly developed by moving the norators to change their common point, and then cross-pairing the nullators and norators in different ways. The disadvantage of the nullor approach is that it gives no information as to the phasing of the amplifiers. This has to be deduced once the format of the circuit has been established by considering the feedback paths. The advantage is that by generating fresh circuits, particular units will have the merit of having anti-phase outputs, grounding of more convenient components or will suggest simplifications not apparent in the original.
• A suitable amplitude-control network for the circuit of Fig. 3 is suggested in ref. 1 and is shown below As the amplitude increases the diodes conduct and the average value of R_5 is reduced until it equals R_4—the condition for stable oscillations.

Further reading
Brokaw, P. FET op-amp adds new twist to an old circuit, *EDN*, June 3, 1974, pp. 75-7.

Cross reference
Set 25, card 9.

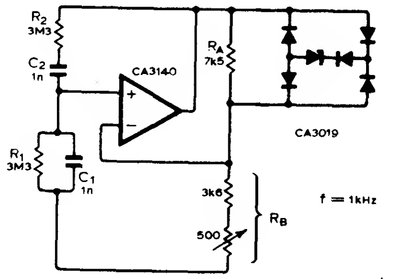






Circuit modifications

The reference quoted brings out two points worth noting. Amplitude control by non-linear feedback is a known method of rapidly stabilizing an oscillator. The disadvantage is that it can be difficult to combine stable amplitude and low distortion, the degree of non-linearity for the first inhibiting the second. One cause of distortion is unequal clipping because of unmatched zeners. By placing a zener inside a full-wave bridge of matched diodes the clipping is identical for both polarities minimizing distortion. The diodes are part of an array and the temperature coefficients of zener and diodes are said to achieve good temperature compensation. The amplifier has p-m.o.s. input devices allowing large resistors and hence small capacitances for low-frequency oscillations. The high slew-rate allows full-output ($\sim$16V pk-pk, quoted) up to 180kHz. Better use of the high gain-bandwidth follows from scaling the series impedances down and reducing the minimum gain for oscillation. The modified relationships are $f = 1/2\pi\sqrt{R_1R_2C_1C_2}$ and $\frac{R_A}{R_B} = \frac{C_1}{C_2} + \frac{R_2}{R_1}$.



References

Bailey, M. Op-amp Wien bridge oscillator, *Wireless World*, vol. 83, Jan. 1977, p.77. Application Note CA3140, RCA.

Voltage control can be applied to lead-lag and Wien oscillators in various ways. One approach is via the Blumlein-Miller effect of Fig. 1 where the input resistance is reduced to very small values as $A > -\infty$. This requires grounded resistances and to modify the passive network to permit this the sustaining amplifier requires a controlled gain and differential inputs (Fig. 2). The resistors R′ then represent a pair of circuits as in Fig. 1 with voltage-controlled amplifiers providing the variable gain to set R′. Such an amplifier (Fig. 3), similar to Set 22, card 8, can also be used as the sustaining amplifier of Fig. 2 with the control voltage pre-set or driven from a peak or mean-sensing circuit for amplitude control. The number of amplifiers can be reduced to two in a circuit related to that of card 9. The low-gain needed for sustaining oscillations ($\sim$3) is defined by the resistor R_F between the emitters while the variable gain amplifier setting the frequency can be as in Fig. 4. By extension the Blumlein-Miller effect can be applied to LC oscillators to replace a capacitor that normally appears in a feedback path by either a grounded capacitor or one connected across a separate amplifier. If A_1 is fixed to sustain oscillations, then A_2 varies the effective capacitance and hence the frequency without disturbing the amplitude condition.

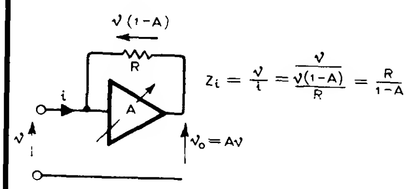


Fig.1

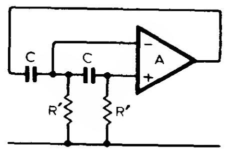


Fig.2

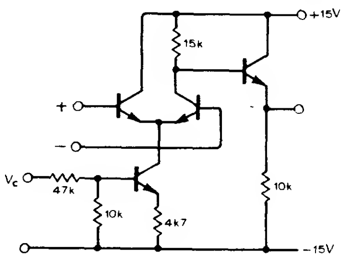


Fig.3

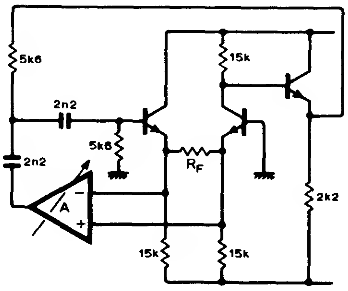
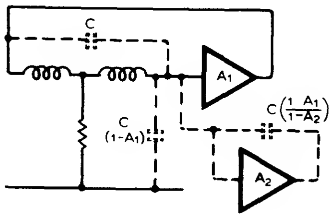


Fig.4



Saha, S. K. Electronically tunable RC sinusoidal oscillator, *IEEE Trans. Instrum. & Meas.*, vol. 24, June 1975, pp. 156/9.

Sun, Y. Generation of sinusoidal voltage (current) — controlled oscillators for integrated circuits, *IEEE Trans. Circuit Theory*, vol. CT-19, March 1972, pp. 137-41.

Passive and active networks

T-networks. The parallel or twin-T network is widely used to obtain a sharp null at a given frequency. By combining this with a small amount of positive feedback oscillations may be sustained at the null frequency if the network is in a negative feedback path i.e. inhibiting oscillation at all other frequencies. For other ratios of parallel and series arm impedances, an inverted output of reduced magnitude is obtained, and this can be applied directly to the inverting input of an amplifier. An apparently new oscillator follows from a change in ground point on the T-network or by redrawing the circuit in nullor form. The output is still applied between A and C and the amplifier is driven by the resulting p.d. between B and C. The network now has an in phase output slightly greater than its input. If the original transfer function T_1 is negative then the new transfer function becomes $T_2=1-T_1$ i.e. $> +1$. A second network has a minimum response at a given frequency and can also be used as negative feedback combined with resistive positive feedback to initiate oscillations. The network is the Bridged T, Fig. 2. Inverting it yields a network with a peak in its response if the output is taken between B and A, and this is the previously described lag-lead network. The response is identical to that of corresponding lead-lag and Wien networks.

The oscillators may be interpreted as (i) frequency dependent n.f.b. with a minimum value at a given frequency just insufficient to neutralize the fixed positive feedback. (ii) frequency-dependent positive feedback just large enough at the same frequency to overcome the fixed negative feedback. Alternatively they may be recognized as particular forms of the bridge oscillators described in the previous set of Circards.

Phase-shift networks. Cascaded RC networks produce a lagging phase-shift while attenuating the signal. A minimum of three sections is needed if the phase-shift is to reach 180° at a finite frequency. With three identical sections the attenuation for this condition is large requiring an amplifier with a gain of -29 to sustain oscillation. The amplifier may be an ideal voltage-amplifier (low output-impedance high input-impedance). Thus networks I or II (Fig. 4) may be used with an amplifier of voltage gain A_v with signal flow as in Fig. 6(a) or with an amplifier of current-gain A_i with signal flow as in Fig. 6(b). Network II can be adapted to provide a current output into a short-circuit, having the same phase relationship to the input, as does the voltage output in II. Thus network III should be fed from a voltage source, should feed into a low impedance input and requires the amplifier to have a defined transimpedance. For the example shown this could be achieved using a standard shunt-feedback amplifier where $A_z=v/i=-R_f$. If $R_f=29R$ the condition for sustained oscillation is met. Where this level of voltage-current gain cannot be obtained, the impedance levels can be graded reducing the interaction between the sections. This reduces the voltage gain requirements to around -8 if n is large (Fig. 5).

Cross references
Set 25, cards 1, 4
Set 26, cards 2, 3, 4
Set 17, card 1

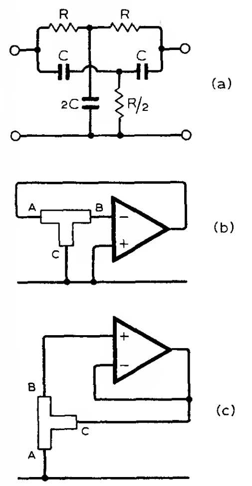


Fig. 1

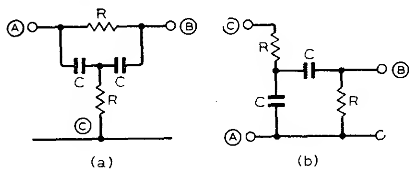


Fig. 2

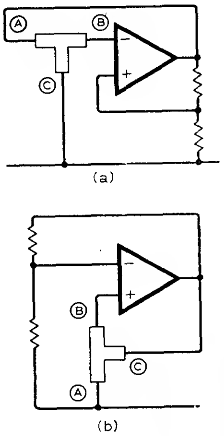


Fig. 3

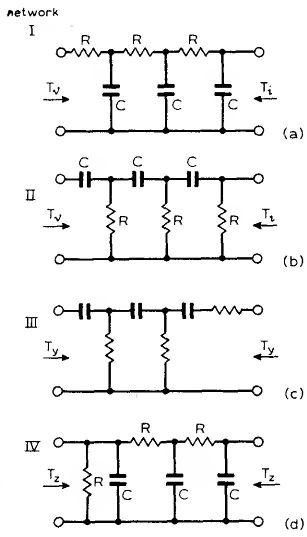


Fig. 4

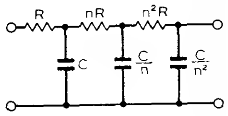


Fig. 5

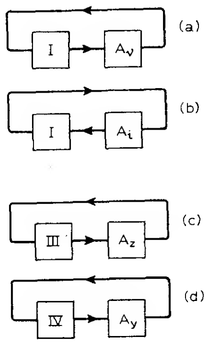


Fig. 6

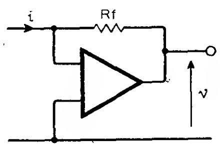


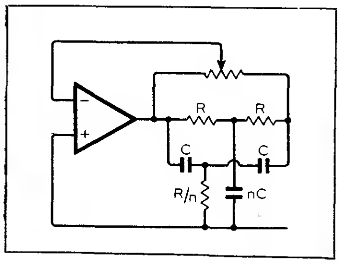
Fig. 7

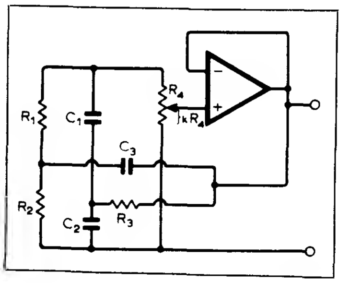
Parallel-T oscillators

Circuit description
It is often assumed that RC networks must attenuate any voltage signal applied to them, and that the voltage gain of the associated amplifier must exceed unity to sustain oscillations. Certain networks have a voltage output that slightly exceeds the input at a particular frequency (see Card 1). They can be derived from networks having phase inversion, and the parallel-T network is one such. The op-amp has a voltage gain very close to unity and with the components chosen there is a "voltage gain" due to the passive network of 1.09. This is sufficient to produce overdriven output with clipping. Adding R_4 to attenuate feedback allows the level of oscillation to be set for minimum distortion. The amplitude control methods shown on Set 25, card 6 are applicable, provided the value of R_4 is kept very much higher than R_1, R_2.
Let $R_1=R_2=R$
$C_1=C_2=C$
$R_3=nR$
$C_3=C/n$
The passive network transfer function is then given by

$$T_v=\frac{(f_0/f-f/f_0)+j(2n-1)}{(f_0/f-f/f_0)+j(2n+1+1/n)}$$

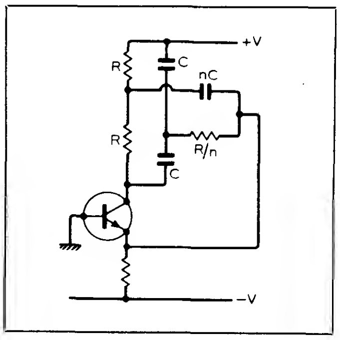
where $f_0=1/2\pi RC$. For $n=1/5$ the response for zero phase-shift when $f=f_0$ is $T_v(f=f_0)\approx 1.094$.
Similarly for $n=1/3$, $T_v(f=f_0)\approx 1.073$.
This suggests that the amplifier gain must not fall more than 5% or so below unity if oscillations are to be maintained.





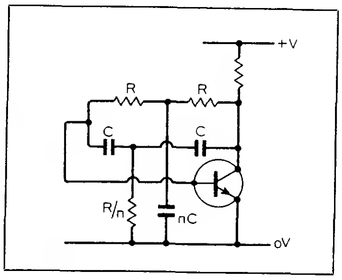
Component changes
IC not critical. Any op-amp capable of accepting 100% negative feedback, or an i.c. voltage follower.
Supply voltage: Normal range of op-amp supplies e.g. ±5 to ±15V.
R_1, R_2 1k to 100kΩ.
R_3. Can range from $R_1/2$ downward. As $R_3 \rightarrow R_1/2$ the response at zero phase-shift→null. When $R_3 \ll R_1$, R_2 loading of the output or by the input becomes more critical.
$R_4 \gg R_1$, R_2. If the ratio is not 10× or more, significant departure from predicted frequency occurs.
C_1, C_2 1n to 1μF, select from frequency equation, when resistors have been chosen from loading requirements.
C_3. C_1/n.
n typically 1/3 to 1/8.
Frequency change in graphed results because increasing feedback brings increased distortion.

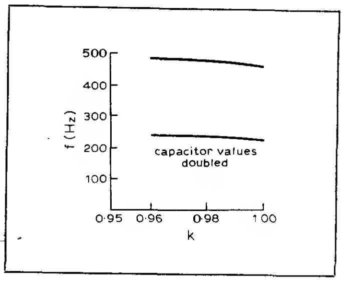
Circuit modifications
● The first circuit can be derived from that overleaf either by shifting the ground-point on the passive network and determining the sign of the amplifier gain required or by drawing the nullor circuit and shifting the ground point on



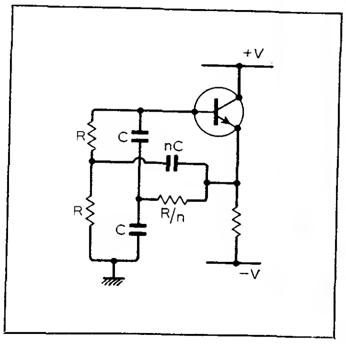
Typical performance
IC_1 741
Supplies ±15V
R_1, R_2 12kΩ
R_3 2.2kΩ
R_4 100kΩ
C_1, C_2 33nF
C_3 150nF
f 475Hz
Oscillation commencing with R_4 set to ≈ 0.965 of maximum

that. Re-drawing in op-amp form then leaves the gain sign to be deduced. Again the loop gain is greater than is needed to sustain oscillation, and the feedback is attenuated to produce minimum distortion. The same constraints on impedance levels apply, and the performance is very similar to the original.
● A transistor has sufficient gain to be used with this network and in one form sometimes referred to as current-driven, the resistors of the T-network also provide the d.c. collector load. The circuit may use either a centre-tapped supply or the base may be fed from a decoupled potential divider across the single-supply. If the transistor is replaced by a nullor and/or the bias components omitted, with the supplies replaced by short-circuits, other forms of the circuit can be visualized. There are three in total, corresponding to which of the three device electrodes is grounded. Note that it is not correct to speak of "common-emitter" etc. since an oscillator has no input and there can be no "common" point.
● If the emitter is grounded then the resistors of the T-network provide d.c. negative feedback to the base. The collector load resistance





loads the output circuit and can with advantage be replaced by a constant-current stage. The transistor needs to be operated at a low current to raise its input impedance so that the passive network is loaded as little as possible.
● The third transistor configuration corresponds to the op-amp circuit on the front of the card viz it is an emitter follower with a voltage gain slightly less than unity. Though not approaching the ideal as closely as an op-amp the gain in each of these configurations is sufficient for oscillation if biased carefully. The frequency response of a transistor can be so much higher than for a compensated op-amp that oscillation up to the MHz region could be possible.



Phase-shift oscillators

Circuit description
The conventional phase shift oscillator uses a cascade of three RC or CR networks giving an output with 180° of phase shift at a particular frequency. It is fed into an inverting amplifier with sufficient gain to overcome the network losses, and oscillation is achieved. An op-amp has an excess of voltage gain and could use a separate passive feedback network to control the gain to the precise level required. Alternatively the final resistor in the chain could be removed from ground and used to drive the virtual ground point of the amplifier as shown, with no change in loading on the passive network. The gain is then defined by R_4/R_3.
The nullor form of the circuit is shown and indicates another viewpoint—summing two signals, one in phase and one inverted, both derived from the output. When these are equal in magnitude oscillations are maintained at the frequency of zero phase-shift.
By changing the ground point an apparently new circuit is obtained as in Fig. 2 where the amplifier is used as a voltage follower. The frequency of oscillation and the gain condition are not identical for the two circuits since neither can be an exact realization of the nullor version. They will differ amongst other reasons because of input common-mode effects with circuit 2 not present in circuit 1. A disadvantage of the CR-sections is that harmonics are progressively less attenuated. When oscillations are vigorous, and distortion ensues, these harmonics are fed back introducing intermodulation distortion and shifting the frequency of oscillation away from the 180° phase-shift frequency of the network.

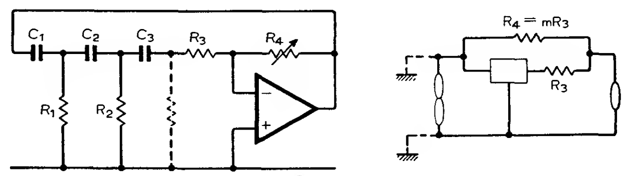


Fig. 1

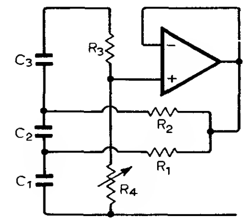


Fig. 2

Component changes
IC: general purpose op-amp—high input impedance advantageous.
R_1, R_2, R_3 1k to 33kΩ
C_1, C_2, C_2 1n to 1μF
$R_4 \approx 29R_1$ if $R_1=R_2=R_3$.
Larger values needed to overcome losses in most circuits.
If the network is graded with the impedances progressively increasing, then the voltage attenuation is reduced e.g. if succeeding resistors are increased by a factor of 2 and capacitances reduced by the same factor the gain required is changed from −29 to −16. In the limit as $n \rightarrow \infty$ the gain requirement is relaxed to −8. This raises the network output impedance to such a level as to place excessive demands on the input characteristics of the op-amp and ratios from 3 to 5 are more realistic.

Circuit modifications
- The principle of the voltage follower circuit is illustrated above. The gain requirement of the amplifier ranges from 0.97 for the original network down to 0.90 for one with graded components as described above.
- In the original circuit the action could be viewed as summing in phase and inverted currents, derived from the input and output voltages of the phase-shift network. A dual form can be constructed in which the input and output currents generate voltages which are summed to zero at the amplifier input. It is shown in summary form in Fig. 4. Re-drawing in nullor form and shifting the ground point to alternate sides of the norator leads to different practical versions.
- The first employs a voltage follower, and the resistor R/n is used to set the condition for oscillation. For equal Rs and Cs in the remainder of the circuit $n \rightarrow 29$ is the appropriate condition. For R 10kΩ, C 33nF, and $R/n = 330$, oscillations were sustained at a frequency of ≈ 1.18kHz.
- The second form has an op-amp with non-inverting input grounded and uses comparable component values to achieve the same frequency. Increasing the capacitance values to 0.1μF reduced the frequency to 375Hz. The components can again be graded to change the voltage- and current-gain requirements of the amplifier.

Typical performance
IC 741
Supplies ±15V
R_1, R_2, R_3 12kΩ
C_1, C_2, C_3 33nF
R_4 360kΩ
f 161Hz for circuit 1
156Hz for circuit 2
N.B. Normal attenuation of equal-valued 3-section phase-shift network is 1/29, indicating $R_4 \approx 29 \times 12$kΩ.
For the voltage follower version, R_4 had to be increased to >500kΩ suggesting loading effect of op-amp input impedance.

Theory. The transfer function of the CR network shown leads to a frequency of 180° phase-shift given by

$$\omega_0{}^2 = \frac{1}{C^2R^2} \cdot \frac{1}{(3+2/n+1/n^2)}$$

and a gain condition.
$A_0 = 1-(3+2/n+1/n^2)(3+2/n)$.
For $n=1$, $\omega_0 = 1/CR\sqrt{6}$,
$A_0 = -29$
For $n \rightarrow \infty$, $\omega_0 \rightarrow 1/CR\sqrt{3}$,
$A_0 \rightarrow -8$.

Cross references
Set 26, cards 1, 2, 4

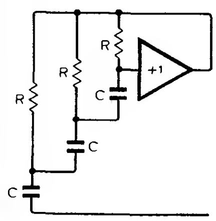
Fig. 3


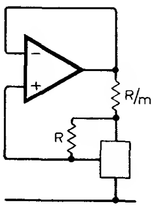
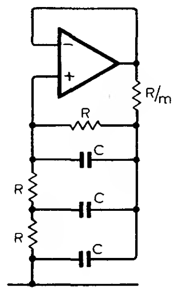
Fig. 4


Fig. 5


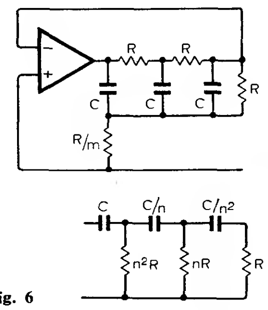
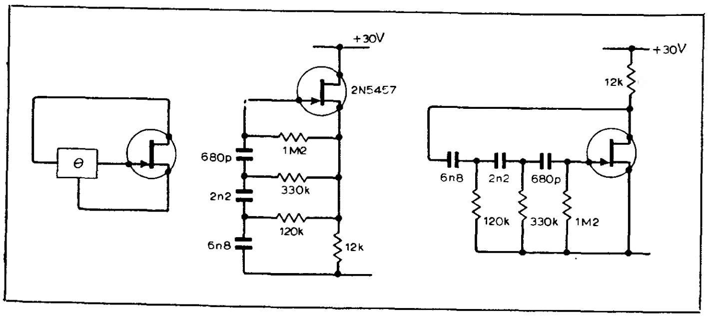
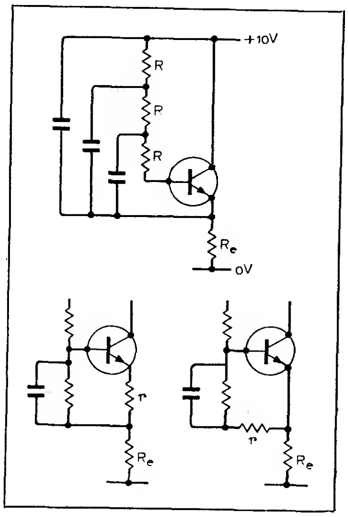


Fig. 6

F.e.t. phase shift oscillators





Circuit description

The basic principle is that of a passive network which has a frequency dependent transfer function such that in one configuration the output is 180° out of phase with the input at a particular frequency. The network contains a cascade of RC or CR sections and by grading the component values, the loading effects of successive sections is reduced. In the limit each section attenuates by $\frac{1}{2}$ with a 60° lag leading to a gain requirement of −8. Practical ratios raise the magnitude of this figure to 10 or 12, within the range of a f.e.t., while the high f.e.t. input impedance prevents loading problems from being significant. The f.e.t., if a junction device, needs either a zero gate/source p.d. or a reverse-biased gate for correct operation.

- This is achieved in the first circuit by allowing the final resistor of the phase-shift network to be connected directly between gate and source. The f.e.t. then has zero gate-source p.d. and operates at its maximum g_m. This is not the condition for maximum voltage gain unless the load resistance can be replaced by a constant current load.

This circuit corresponds directly to that of card 3, circuit 2, with the reduced gain of the f.e.t. just sustaining oscillation.

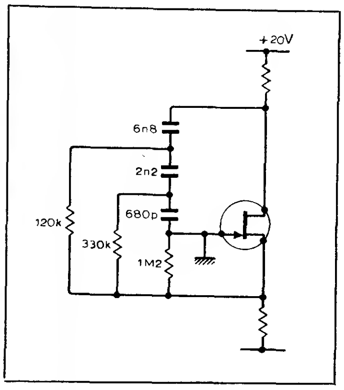
- Corresponding to card 3, circuit 1, the phase-shift network can be interposed between drain and gate with source grounded. Again the restricted voltage gain of the f.e.t. means that in the absence of a constant-current load, the supply voltage has to be raised to maximize the load resistance for a given operating current and g_m. In general a junction f.e.t. has a greater voltage gain for a given supply voltage when operated at a lower current. This is because as the current falls, the g_m falls more slowly while the load resistance increases directly in proportion to the reduction in current. Hence the voltage gain $(-g_m R_L)$ increases in magnitude. The limit is set when the passive network exerts significant loading on the output.

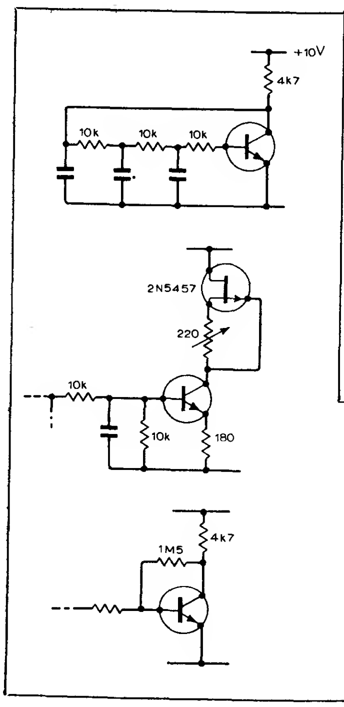
- The third form (over) of the oscillator is that of grounded gate in which both source and drain loads have to be present. One method of increasing the voltage gain in each of these circuits is to include a decoupled resistor in the source lead. This lowers the current allowing the load resistors to be increased as indicated above.

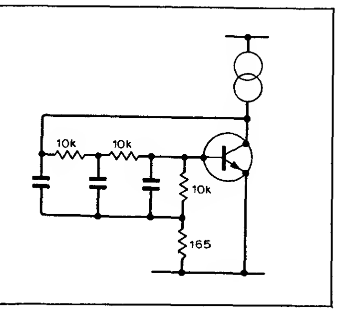
- With the component values as indicated, the frequency of oscillation was ≈ 75Hz. Increasing the resistors to 330k, 1.2M and 3.3MΩ respectively reduced that frequency to ≈ 30Hz.

- A bipolar transistor requires a forward bias on its base-emitter junction. A resistive feedback path between collector and base provides this, and this leads to the use of the RC phase-shift network as opposed to the CR network with the junction f.e.t. (there is another advantage of this network, in that harmonics are attenuated reducing the shift in frequency due to intermodulation effects when the output distorts). The network is used in its reversed mode ideally requiring current drive and a low impedance load. By using a f.e.t. or other constant current load, the output impedance approaches the ideal while series feedback can increase the input impedance. The final network resistor is then grounded and the p.d. across it used as the feedback, more nearly satisfying the impedance conditions. Shunt feedback would improve the input matching but would drive the network more nearly from a voltage source. Any mismatch shifts the frequency of oscillation. A better solution is to combine the high output impedance arrangement of the amplifier with the network arrangement of card 3, circuit 6.

- Grounded-collector and grounded-base versions of the oscillator follow in the same way as for the f.e.t. oscillators and, feedback arrangements to control the amplitude of oscillation are indicated.







Cross references

Set 1, card 9
Set 26, cards 1, 3
Set 25, card 1

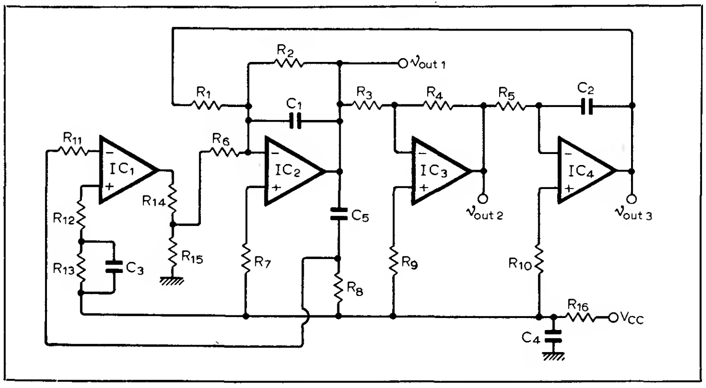
C.d.a. oscillator

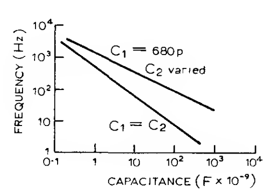
Circuit description
This circuit* takes advantage of the quad current-differencing amplifier package and is in the form of a two-integrator loop (IC_4, IC_2) plus inverter (IC_3), which does not contribute to the loop gain, but provides a 180° phase shift between v_{out1} and v_{out2}. For $R_3 = R_4$, frequency of oscillation is given by $\omega = 1/C_1R_1C_2R_5$. Resistor R_2 acts as a damping resistor IC_1 acts as a comparator and supplies the input square wave fed back to IC_2 to keep "ringing" the circuit. IC_2, IC_3, IC_4 is a bandpass filter with a Q defined by R_2C_1. To initiate oscillation with the loop closed, the differential input to IC_1, (C_4 charge initially zero) causes IC_1 to rise initially to positive saturation, but rapidly changes to 0V when C_4 charges. This shock-excites the loop (overall phase shift zero), amplitude of oscillation builds up until a balanced condition is reached, where the output across R_{15} is just sufficient to maintain a steady oscillation, approximately 220mV square waves.

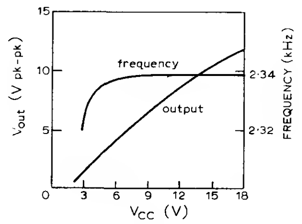
Typical data
IC_1 to IC_4 $\frac{1}{4}\times$ LM3900N or MC3401
R_1, R_3, R_4, R_5, R_9 100kΩ
R_2 6.8MΩ, R_6 330kΩ
R_7, R_{10}, R_{12} 220kΩ, R_8 470kΩ
R_{11} 22kΩ, R_{13} 270kΩ
R_{14} 47kΩ, R_{15} 1kΩ, R_{16} 4.7kΩ
C_1, C_2 680pF, C_3 2.2μF
C_4 10μF

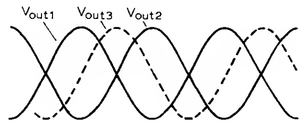
V_{CC} +10V
V_{out1} V_{out2} V_{out3} 6.9V pk-pk

Oscillation frequency 2338Hz
Phase difference between outputs 90° as shown in waveform diagram.
Total harmonic distortion: 0.55%.





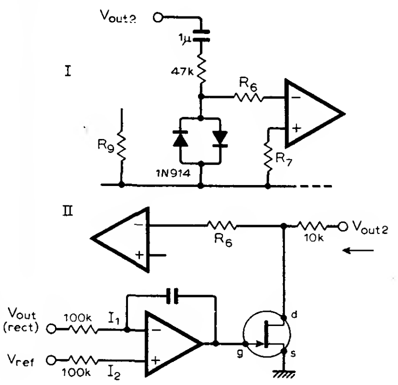


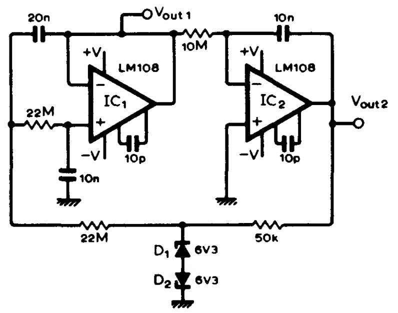


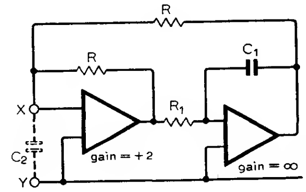
Component changes
R_{15} 680Ω, V_{out} 4.2V pk-pk, t.h.d. 0.35%. For 1500Ω, V_{out} 8.6V pk-pk, t.h.d. 0.5%. R_{14} 39kΩ, V_{out} 8.6V pk-pk, t.h.d. 0.4%. For 56kΩ, V_{out} 5.6V pk-pk, t.h.d. 0.25%. R_6 220kΩ, V_{out} 8.8V pk-pk, t.h.d. 1.2%. For 470kΩ, V_{out} 4.4V pk-pk, t.h.d. 0.3%. R_2 7.8MΩ V_{out} 6.8V pk-pk, t.h.d. 0.6%. For 4.7MΩ, V_{out} 4.1V pk-pk, t.h.d. 0.2%. For all above alterations, frequency does not change more than 0.2%. Variation with supply C_1, C_2 shown on graphs.

Circuit modifications
- Comparator can be replaced by diode limiter of I above. Control is less precise because square wave is not available across diode.
- More precise limiting achieved from II. V_{out2} is half-wave rectified, capacitor current is $(I_1 - I_2)$. If output increases, the p-type junction f.e.t. gate is driven negative to increase f.e.t. conduction, and v_{out2} is attenuated to decrease overall positive feedback and thus reduce output.
- Alternative integrator-based oscillators. IC_1 is connected as a low pass filter and IC_2 as an integrator. Diode limiters D_1, D_2 introduce distortion. The symmetrical clipping minimizes even harmonics, and the predominant third harmonic is attenuated by about 40dB by low-pass filter action. Temperature compensation to stabilize amplitude is obtained if n-p-n transistors connected as diodes (base-emitter breakdown 6.3V) are used as limiters. Circuit between X and Y is a simulated inductor. C_2 induces oscillation if amplifier gain is +2.

Cross references
Set 26, cards 6, 7 Set 17, card 2
*Rossiter, T. J. M. Sine oscillator uses c.d.a. *Wireless World*, April 1975.







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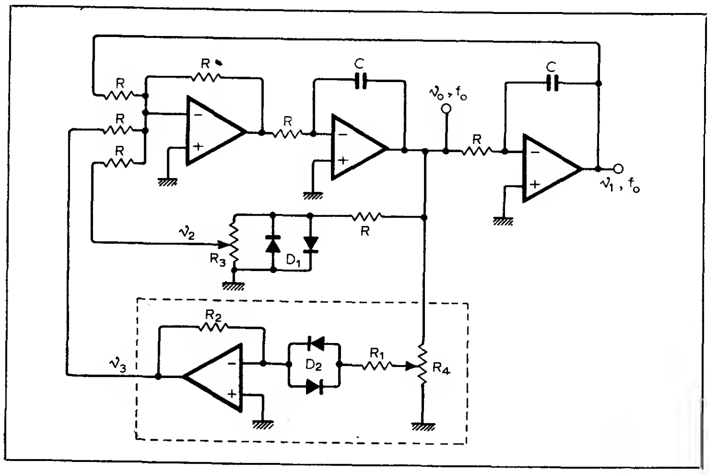
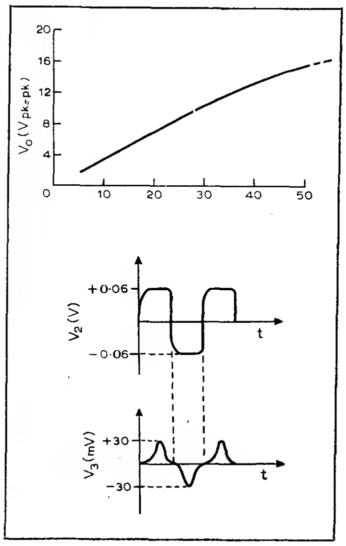


Two-integrator loop oscillator

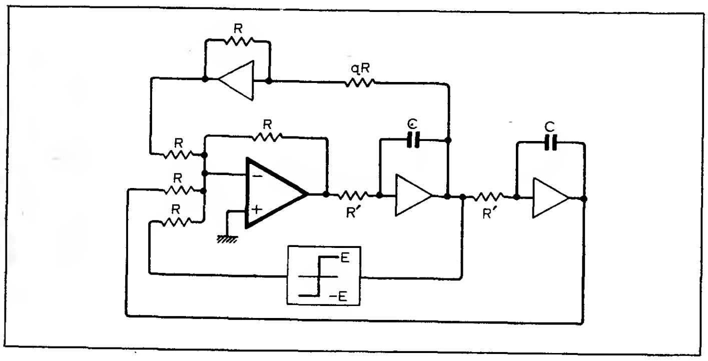
Components
Supply ±15V
IC 741 op-amps
R 10k $\pm5\%$
R_1 22k $\pm5\%$
R_2 220k $\pm5\%$
R_3 100kΩ, R_4 1kΩ



C 0.1μF $\pm5\%$ polyester
D_1, D_2 1N914

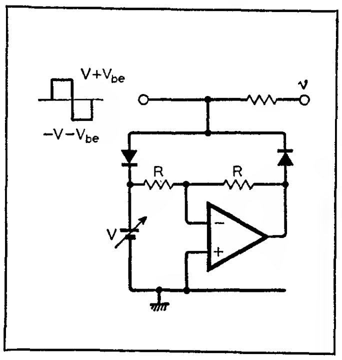
Performance
With R_3 set at 0.1, a value which caused the circuit to oscillate with slight clipping when R_4 was at zero, variation of v_o plotted against the inverse of R_4 setting is shown. Total harmonic distortion throughout this range lay between 0.96% and 0.86% and the frequency f_o was 159.35 ±10Hz. With perfect op-amps, identical R and lossless capacitors, the theoretical f_o is $1/2\pi CR$. Typical traces for v_2 and v_3 are shown.

Circuit description
The circuit is a straightforward two-integrator oscillator[1] with v_2 providing sufficient positive feedback to overcome the damping inherent in the imperfect capacitors and op-amps thus ensuring oscillation. This positive feedback is of a constant nature in that once oscillation has reached a level to ensure conduction of the diodes D_1 then v_2 is approximately a square wave whose magnitude does not depend on v_o. Potentiometer R_3 can provide a measure of amplitude control[1] with the positive feedback always balancing out the inherent damping. However, although the circuit is invariably used as a low frequency oscillator (< 1kHz) problems of amplitude control do arise at high frequencies when phase shifts in the imperfect components reduce the inherent damping[2]. Under these conditions the basic oscillator tends to go into oscillation limited only by saturation of the op-amps. To prevent such oscillation the dead zone limiter (the dotted section) is added. This section produces an output only when v_o is sufficient to cause the diodes D_2 to conduct; when this happens a large amount of negative feedback is applied, thereby damping out any tendency to oscillate with too large a magnitude. One therefore has a section comprising D_1 etc forcing oscillation and another section comprising the dead zone limiter holding down the oscillation, giving good overall control[2].
Distortion content in the output v_o (or v_1 which is 90° out of phase with v_o) depends on circuit Q. With good quality passive components Q is approximately $K/2$ where K is the op-amp open-loop gain. should not be so low as to cause damage to the diodes D_2.



Component changes
Polycarbonate capacitors up to 10μF and higher values of R can reduce the operating frequency to a fraction of one hertz. Amplifiers of greater gain than that of the 741 would be necessary to increase the operating frequency much beyond 1kHz (for which Wien type oscillators are available, Set 25). The ratio R_2:R_1 should be kept large to give heavy negative feedback when necessary. At the same time R_1 should not be so low as to cause damage to the diodes D_2.

Circuit modifications
- Because the negative feedback section prevents oscillations from growing without bound there is no need to include the limiter in the positive feedback path. The positive feedback can therefore be directly from the output of first integrator back to the summing inverter, and might be expected to produce lower distortion figures.
- Amplitude control with



sinusoidal negative feedback and limited positive feedback can be obtained as shown above. It can be shown[1] that for this circuit the output is given by $2qE\sqrt{2}/\pi$ volts r.m.s. A possible voltage-controlled limiter is shown above

References

1. Girling, F. E. J. and Good, E. F. Active filters—8. *Wireless World*, March 1970.
2. Foord, A. Two-phase low-frequency oscillator, *Electronic Engineering*, Dec. 1974.

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Set 26: RC oscillators II—7

Four phase oscillator

Components

R_1 to R_7, R_{12}, R_{22} 10kΩ
R_8 to R_{11} 47kΩ
R_{13} 220kΩ, R_{14} 330kΩ
R_{15}, R_{16} 100Ω
R_{17} 100kΩ, R_{18} 2.7kΩ
R_{19}, R_{20} 2.2MΩ, R_{21} 22kΩ
R_{23}, R_{24} 47kΩ twin gang
R_{25} 10kΩ, R_{26} 50kΩ
R_{27} 220kΩ
C_1, C_2 22nF
D_1 to D_4 1N914
D_5 6V zener, F.e.t. BF244
IC_1 to IC_6 741 op-amps with ±15V supplies

Performance

This circuit provides four outputs, at points A, B, C and D. If A is taken as the reference then B, C and D are 180°, −90° and +90° out of phase with A respectively. The output signal magnitude was set at 15V peak-to-peak by suitable setting of R_{25} and the graphs shown indicate the total harmonic distortion at various frequencies, obtained by varying R_{23} and R_{24}. The scales chosen are not ideal but the graphs all show the same general shape as that for B which is shown in full. In relation to one another they are as one might expect with the exception that one would anticipate B to be better than C since harmonics are reduced by an integrator. The difference is slight and could be accounted for by imperfections in the op-amps.

Description

This oscillator is closely related to the two-integrator loop oscillator[1].
It has one extra inverter (IC_4) which enables one to produce four outputs all 90° apart.
This allows the use of a four-phase rectifier (D_1 to D_4) which requires little smoothing to produce a direct voltage proportional to the output magnitude.
The oscillator works on the basis of fixed negative feedback (from D via R_{13} to IC_1) being balanced by a variable amount of positive feedback. The amount of positive feedback is dependent on the discrepancy between the desired output voltage (set by D_5 and R_{25}) and the actual output as sensed by D_1 to D_4. The positive feedback path is from the output of IC_2 via IC_6 and R_{14} to IC_1. IC_2 and its associated circuitry provide a path of positive gain, the magnitude of which gain is set by the resistance of the f.e.t. (n-channel). The output of IC_2 is

$$v_p \frac{R_{12}}{R_{12}+R_{27}} \cdot \frac{R_{21}+R_F}{R_F}$$

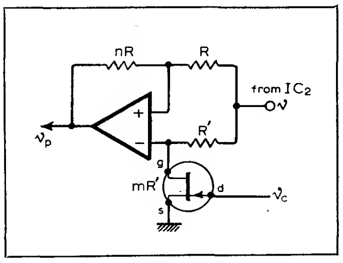
where R_F is the f.e.t. resistance. R_{20} and R_{19} serve to linearize the f.e.t. resistance and it is possible to omit R_{20} and replace R_{19} by a short for the purpose of explanation.
Bearing in mind the f.e.t. characteristics note that a positive increment in v_c (corresponding to a reduction in output voltage) gives a lower value of R_F and a larger positive feedback voltage, v_p, which will increase the magnitude of the oscillation.
A similar balancing effect occurs if the oscillations increase in magnitude. Note that the d.c. value of v_c must be slightly negative for correct f.e.t. operation and that this is ensured by the integrator action of IC_5 and its associated circuitry.
A four-phase oscillator allows rapid amplitude stabilization since any change in amplitude is quickly sensed; this is of importance in low frequency oscillators[2].

Circuit modifications

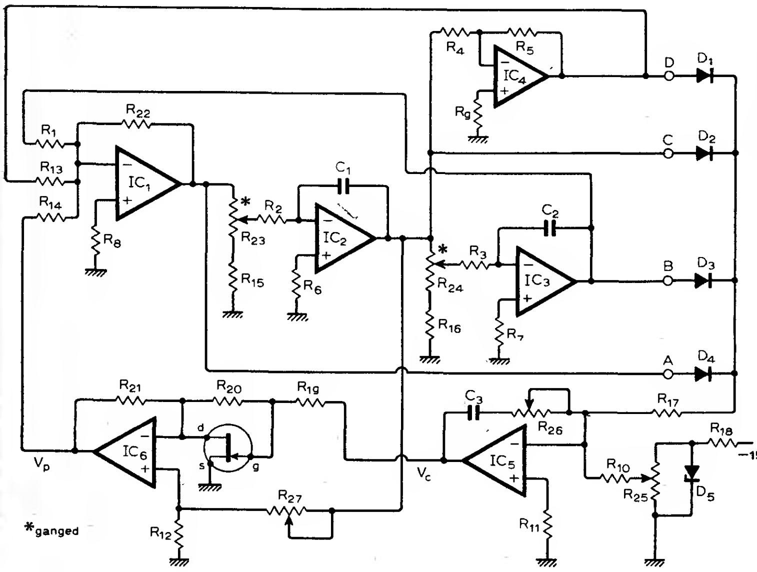
- Use of fixed positive feedback and variable negative feedback requires care with polarities of most of the items in the amplitude control loop.
- It is possible to remove the positive feedback link altogether and obtain either positive or negative feedback, depending on whether the output is low or high, by using the circuit above right to generate the feedback signal. For this circuit

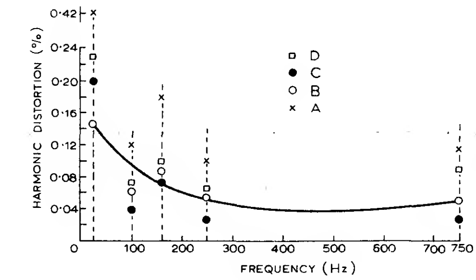
$$v_p = v(1+n)\left(\frac{m}{m+1} - \frac{n}{n+1}\right)$$

Since m can be controlled by v_c, positive and negative values of v_p can be obtained giving positive and negative feedback respectively.



- An alternative, though more expensive scheme, to provide the same function of positive and negative feedback control on the same path is to use a multiplier, one input of which is the output of IC_2 and the other is v_c. The magnitude and sign of v_c control the amount and nature of the feedback[2].

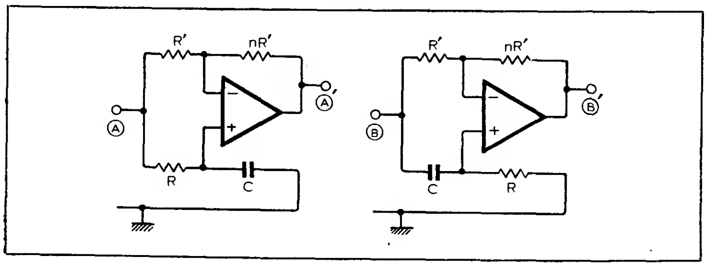




References

1. Circards this set, card 6.
2. Fast amplitude stabilisation of an R-C oscillator, Vannerson and Smith, *IEEE Journal of Solid State Circuits*, vol. SC9, no. 4, Aug. 1974.
3. Circards, this set, cards 9, 10.

All-pass network oscillator



Typical performance
IC_1, IC_2 741
Supplies $\pm 15V$
R, R′ 10kΩ
C 15nF
$n \approx 1$
f 1065Hz
Connect A′ to B, B′ to A.
Oscillation commences for loop gain > 1 at zero phase-shift. Set either value of $n > 1$.

Circuit description
Most oscillators have a peak in their loop amplitude response at the frequency of zero loop-phase-shift. This stems from the band-pass nature of their transfer function. The Barkhausen criterion demands only that the loop-gain exceeds unity at this frequency and it is not necessary for the amplitude response to be frequency dependent—though it may assist in reducing the effect of distortion on frequency. If circuits having frequency dependent phase-shifts are combined such that the overall phase-shift is zero at a single frequency then oscillations can occur at that frequency. These circuits are called all-pass (non-minimal phase-shift circuits) and can be made with one op-amp three resistors and one capacitor. The gain can be trimmed about unity. The outputs are 90° out of phase and the frequency of oscillation is the same as for Wien, two-integrator and gyrator oscillators using the same components.
Transfer function of A to A′. T_{VA} given by equating potentials at inputs

$$\frac{v_O/n+v}{n+1}=\frac{v}{1+sCR}$$

$$\therefore T_{VA}=\frac{v_O}{v}=\frac{n-sCR}{1+sCR}$$

For $n=1$, magnitude of transfer function is unity at all frequencies with the phase shift varying from zero as $f\to 0$ to $-\pi$ as $f\to\infty$. The phase shift is $\pi/2$ lagging at $\omega=1/CR$.

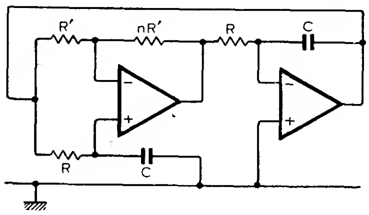
Similarly $T_{VB}=\dfrac{sCR-n}{1+sCR}$
with unity gain for $n=1$ and a $\pi/2$ phase-lead at $\omega=1/CR$.

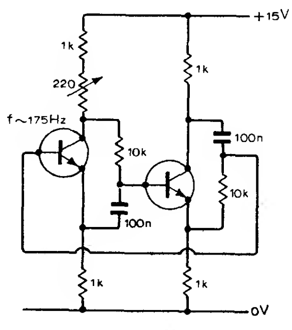
Component changes
IC_1, IC_2: compensated op-amps not critical except where high frequency operation required.
R′: 1k to 100kΩ
nR′: Replace one of the resistors by a thermistor or other amplitude sensitive network for amplitude control. This may necessitate reducing R′ to suit the operating resistance of the thermistor.
C: 1n to 1μF
R: The capacitors can be switched to change ranges as in Wien etc. oscillators and the resistors R replaced by a twin-gang potentiometer for continuous frequency control.
N.B. In the majority of oscillators of this kind the frequency of oscillation is of the form $f=1/2\pi RC$ where all Rs and Cs are equal.

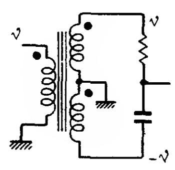
Circuit modifications
The transfer function of an integrator is given by $T_V=-1/_sCR$Combining this with the first all-pass circuit overleaf gives an overall transfer function.

$$T_V=\left(\frac{n-sCR}{1+sCR}\right)\left(\frac{-1}{sCR}\right)$$

For $n=1$ this reduces to unity when $(sCR)^2=-1$, i.e. at $f=1/2\pi CR$ the same frequency as in the previous circuit. The outputs differ in phase by 90° and amplitude control can be achieved by replacing nR' by a low-power thermistor, and R' by a low value resistor to match e.g. R54 (ITT) and a 1kΩ resistor.
Many other oscillators have been designed using various combinations of lead, lag, all-pass and integrator networks and some examples are quoted in the references. In all cases some non-linearity is required to control and limit the amplitude of oscillation—these include diode clipping/limiting saturating high-gain amplifiers, thermistors lamps etc.
A very simple form of oscillator based on the all-pass network idea can be implemented using only two transistors. The all-pass networks load the transistor outputs causing departures from the theoretical performance, but this can be allowed for by increasing one of the collector resistors as shown. The voltage swing is limited because the self-biasing action of the circuit leaves a relatively small collector-emitter p.d. for each transistor. The outputs should not be heavily loaded or the frequency/amplitude will be disturbed but the circuit gives roughly quadrature outputs with a low component count. The transistors are general purpose silicon planar. The all-pass network behaves similarly to that using a transformer with a centre-tapped secondary as indicated.







Further reading
Yewen, J. Two-phase sinewave oscillator, *New Electronics* 1974, vol. 7, no. 6, p. 25.
Das, S. K. & Das G. Amplitude stabilized low frequency oscillator. *Int. J. Electron.*, 1972, vol. 33, pp. 371-5.
Hanna, N. N. and Kamel, S. A. A low frequency sine-wave oscillator, *Int. J. Electron.*, 1973, vol. 35, pp. 685-90.
Baril, M. Three-mode network is filter or oscillator, *Electronics*, Apr. 12, 1973, pp. 105/6.

Gyrator oscillators

Circuit description
A gyrator is a circuit that synthesizes an impedance at one port that is related to the physical impedance presented at a second by the relationship

$$Z_1 = R^2/Z_2.$$

If one impedance is capacitive e.g. $Z_2 = 1/j\omega C$ then $Z_1 = j\omega CR^2$ equivalent to a pure inductor having $L = CR^2$. Such a circuit can be constructed using two ideal op-amps, four resistors and one capacitor. The circuit is widely used in active filters but can be used to make oscillators. Many variants are possible. First the passive network can be changed so that the locations of the capacitors vary. Two capacitors are used occupying any pair of the starred locations—circuit 2 is an example. Then the amplifier inputs may be moved so long as they maintain the circled junctions at equal potentials—circuit 3 is derived from circuit 2. An additional feedback path labelled R_{f1} or R_{f2} initiates oscillation.

Typical values
R 10kΩ, C 15nF, R_{f1} 680kΩ, f 1055Hz.
The p.ds across Z_2 and Z_3, Z_4 and Z_5 are equal as are the currents in Z_1 and Z_2, Z_3 and Z_4, Z_5 and the source. Thus the input current is given by

$$\frac{v}{Z_1}\times Z_2\times\frac{1}{Z_3}\times Z_4\times\frac{1}{Z_5}$$

Input impedance of the circuit is $v/i = Z_1Z_3Z_5/Z_2Z_4$. If a capacitor replaces Z_2 or Z_4 and the other elements are resistive, the input impedance becomes inductive. If a capacitor is placed across the input port, it resonates with this synthesized inductor, and only a small amount of positive feedback is needed to overcome losses due to finite amplifier gain etc. and

Fig. 1

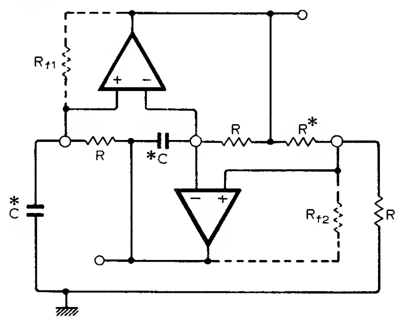
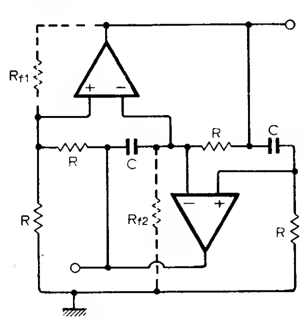


Fig. 2



Equivalent circuit

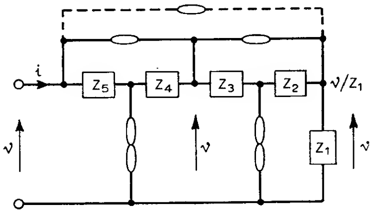
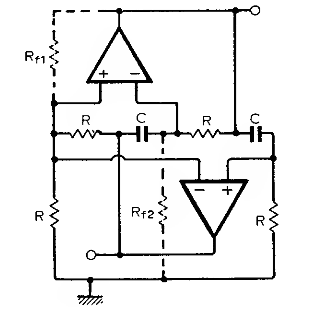


Fig. 3



produce sustained oscillation. For $Z_1 = Z_2 = Z_3 = Z_5 = R$ and $Z_2 = 1/j\omega C$, $Z_i = j\omega CR^2$.
The combination of a capacitor C and a synthesized inductor, $L = CR^2$, formed by gyrating the other capacitor produces an equivalent parallel tuned circuit. The self resonant frequency is

$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{CR^2C}} = \frac{1}{2\pi CR}$$

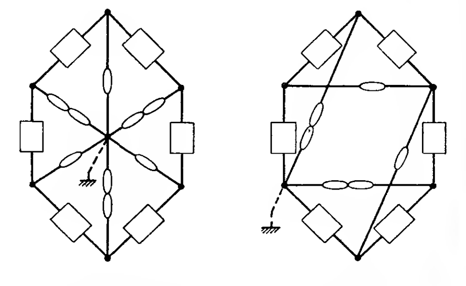
This is the same frequency as occurs in a Wien oscillator using the same component values in the frequency dependent network. Tuning of the oscillator frequency is obtained by varying any of the component values either singly or in pairs. For ideal op-amps there is no damping and the circuit is continuously on the point of oscillation regardless of component values. In practice an external component or network may be added to initiate and sustain oscillation.

Two-integrator loop
This circuit, described on cards 6, 7, uses the same passive components and has the same frequency of oscillation. Though developed separately these two are different forms of the same basic circuit. This can be seen by re-drawing the passive networks in a similar format (where the impedances in practice are composed of four resistors and two capacitors). Then replacing the input of each amplifier by a nullator and the output/ground port by a norator the identity can be seen. Of the six vertices, three are equipotential points with no current being fed into or taken out of the vertices (the properties of the nullators). The other three points are at arbitrary potentials, regardless of loading effects at these points. Provided the passive networks are chosen correctly there can be only one frequency at which these constraints are met; oscillation occurs at that frequency but with undefined amplitude. An external network must be added to limit the amplitude e.g. by using a f.e.t. to place controlled variable damping across a grounded capacitor with positive feedback to initiate the oscillations.

Cross references
Set 26, cards 6, 7, 10
Set 25, cards 9, 10
Set 1, cards 6, 7

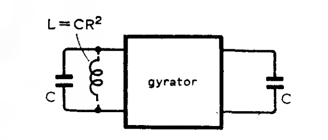
Further reading

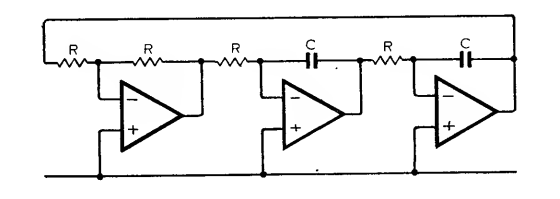
Pauker, V. M. Equivalent networks with nullors for positive immitance converters, *IEEE Trans. Circuit Theory*, Nov. 1970, pp. 642-4.
Antoniou, A. New gyrator circuits obtained by using nullors, *Electronics Letters*, Mar. 1968, pp. 87/8.



Nullor version of two-integrator oscillator

Nullor version of gyrator oscillator

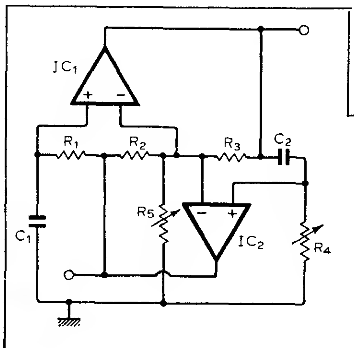


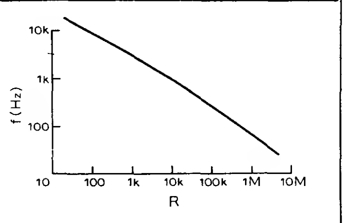


Wide-range gyrator oscillator

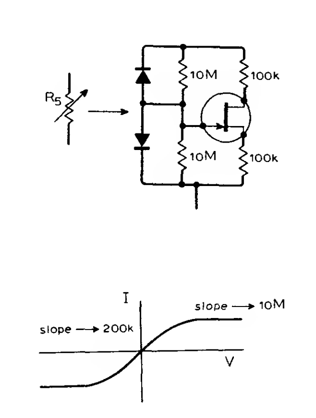
Circuit description
Another version of the gyrator oscillator, this one uses a single variable resistor, R_4, to control the frequency of oscillation. For ideal active and passive elements only the smallest amount of positive feedback would ensure that the amplitude builds up until peak clipping restores the loop gain condition. Losses including

Typical performance
$IC_{1,2}$ 741
Supplies $\pm15V$
R_1, R_2, R_3 $10k\Omega$
C_1, C_2 15nF
R_4 $10k\Omega$
R_5 $470k\Omega$
f 1.05kHz





those due to capacitor loss factors, require the presence of R_5 or other path to initiate the oscillations. As R_4 is varied, the value of R_5 at which oscillations just start also varies. Though much greater than R_1, R_2, R_3 at all settings of R_4, either R_5 has to be re-adjusted manually/automatically or set to a low enough value to ensure oscillation over the whole range. This brings increased clipping at high values of R_4. The circuit is of considerable interest because it illustrates the way in which gyrator oscillators can achieve a very wide range of frequencies using a single variable resistor —greater than 200:1 in this example. Amplitude control methods could include replacing R_5 by a network incorporating a f.e.t. whose on-resistance is varied via a peak-rectifier; or by replacing it by a non-linear network adjusted so that it has a low resistance at low amplitudes to excite the circuit, rising at higher amplitudes to levels where the loop gain is insufficient. A possible example for such a network is shown opposite.
It is equally possible to change the frequency by varying both capacitors and/or a pair of resistors, making the components switched or continuously variable. This reduces the variation needed in R_5 but does not eliminate it. In the original circuit, for



Amplitude control network

$$R_1 = R_2 = R_3 = R,\ C_1 = C_2 = C$$

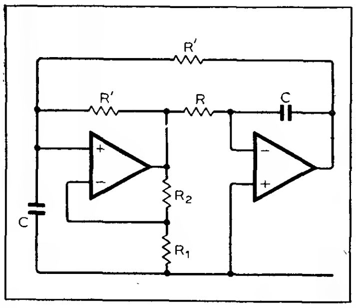
$$f = \frac{1}{2\pi CR}\sqrt{\frac{R}{R_4}}$$

Component changes
The component values will be comparable to those used in the two integrator oscillators (cards 6 and 7). At high frequencies, cumulative phase-shifts can raise the effective Q of the circuit, while slew-rate limiting can lead to unpredictable jumps in the amplitude of oscillation.
$IC_{1,2}$: Compensated op-amp. High input impedance necessary if R_4 is to be made very large.
R_1, R_2, R_3: 1k to $100k\Omega$
C_1, C_2: 1n to $1\mu F$
R_4: 47 to $4.7M\Omega$
R_5: Typically 220k to $2.2M\Omega$. Low resistance needed at high frequencies.
Supplies: ±5 to $\pm15V$.

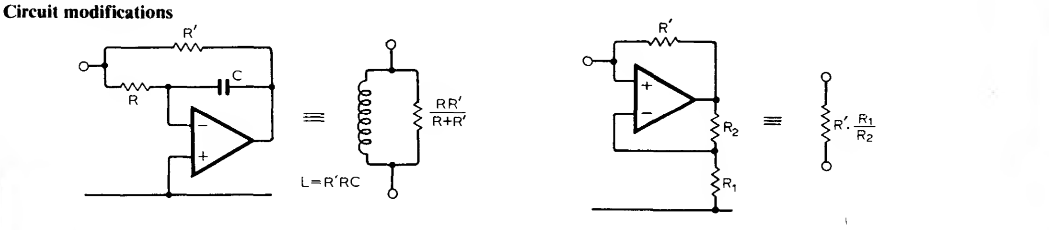
Circuit modifications
- Any other circuit that synthesizes a pure inductor can be combined with a capacitor to produce the effect of a self-resonant LC circuit. The example shown is best considered in two parts. First an integrator with overall resistive feedback from input to output via R′ simulates a parallel LR circuit. This is then combined with a negative resistance circuit (also viewable as a n.i.c. acting on R_1). If the integrator is fed from the output of the first amplifier, less compensation is required for the positive resistances in the system by this negative resistance circuit. For oscillation $R_2 = R_1$ with ideal amplifiers.
- R_2 or R_1 may be made variables using thermistors, lamps etc, or the first amplifier can be replaced by any other variable gain amplifier capable of being remotely controlled e.g. multiplier, a.g.c. amplifier etc.
- Because of the integrator, the two outputs are in quadrature throughout while control of frequency via a single component is again possible.

Further reading
Ford, R. L. and Girling, F. E. J. Active filters and oscillators using simulated inductances, *Electron. Lett*, vol. 2, 1966, p. 52.
Harris, R. W. Two-op-amp RC resonator with low sensitivities, Proc. 16th Mid West Symposium on Circuit Theory, 1973, vol. 1.

Cross references
Set 26, cards 6, 7, 9, 10
Set 25



Circuit modifications



wireless world circard

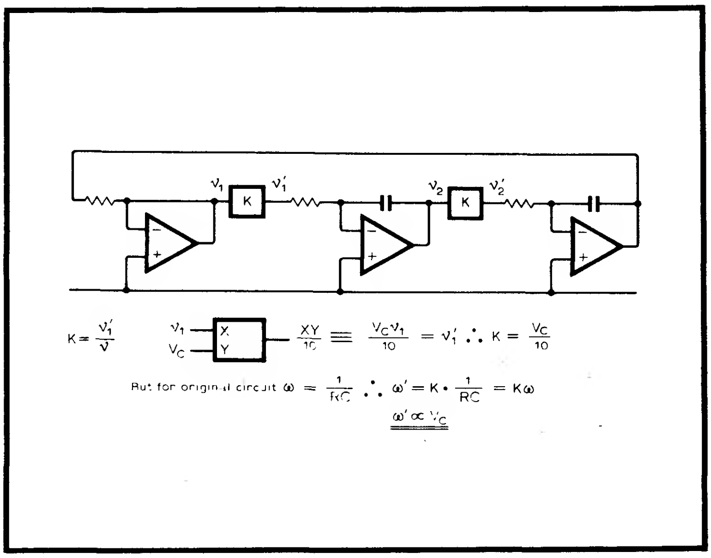
Set 26: Up-date

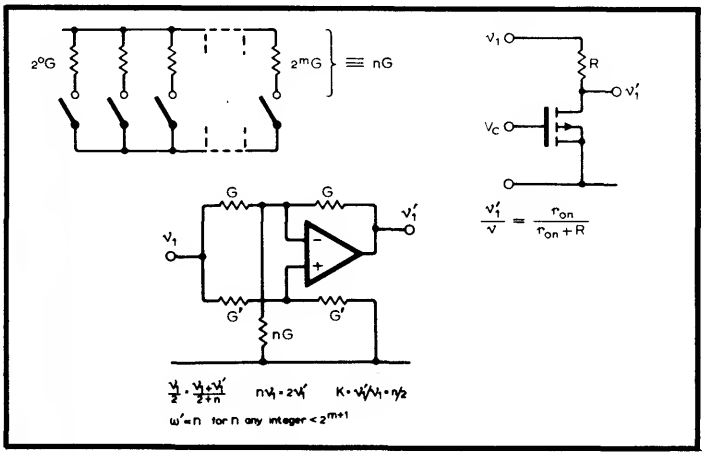
Control of oscillator frequency from an external signal is of increasing importance in such diverse applications as computer-controlled test systems, electronic musical instruments and communication receivers. Rather than giving detailed information on one or two circuits this update surveys the novel and ingenious methods proposed in recent publications. To aid comparison they are shown applied to a two-integrator loop, the principles then serving for both oscillatory and filtering functions. The same methods can be applied to most other RC oscillators.

Consider a pair of identical blocks K which change the voltage transferred from one stage to the next. Increasing the voltage by a factor K increases the current fed to the capacitor by K charging it more rapidly. It is equivalent to reducing the resistor by a factor K i.e. the frequency of oscillation is increased K times. (N.B. No amplitude control etc indicated; see details on preceding cards.) The block K can be provided by an analogue multiplier in which K becomes proportional to the control voltage V_c fed to one of the multiplier inputs.

Reference

Sparkes, R. G. and Sedra, A. S. Programmable active filters, *IEEE J. Solid State Circuits*, Feb. 1973, pp. 93-5.





Any other amplifier or attenuator can be substituted including a simple field-effect transistor, for example, a n-channel m.o.s.-f.e.t. from the CD 4007 package as shown. The control is then non-linear and to avoid distortion the signal levels must be low. A second remote-control method is via a digitally-controlled amplifier or attenuator. An $R/2R$ ladder network may be used directly if it can handle bipolar signals. A neat alternative is the balanced amplifier shown in which any additional conductance to ground produces a proportional output. The switches control a binary-weighted resistor network, and because the switches are grounded might be t.t.l. open-collector outputs or multiple m.o.s. devices from certain c.m.o.s. packages (Set 27 card 7), again provided the signal level is restricted. Any other electronic switches including analogue gates can be substituted.

Reference

Hamilton, T. A. Stable digitally-programmable active filters, *IEEE J. Solid-State Circuits*, Feb. 1974, pp. 27-9.

To summarize the methods so far: in place of direct variation of the resistances necessitating a twin-gang potentiometer we have substituted (i) voltage-controlled amplifiers or attenuators (ii) digitally-controled amplifiers that can be programmed in discrete, binary steps. A third principle that has been in use for more than twenty years is regularly re-invented. An electronic switch periodically opened and closed replaces the multiplier attenuator or amplifier. If the switching rate is high compared with the signal frequency the effect is to make the average current proportional to τ, the on-period of the switch. A pair of switches inserted at locations K (above) results in $\omega' \propto \tau$. If switching and signal frequencies are widely separated, simple passive filtering serves to remove the output transients. This is a form of p.w.m. control and can be implemented via a voltage-controlled monostable or by a member of digital techniques including the counter-based idea indicated or binary rate multipliers and dividers.

Reference

Harris, R. W. and Lee, H. T. Digitally controlled, conductance tunable active filters, *IEEE J. Solid State Circuits*, June 1975, pp. 182-5.