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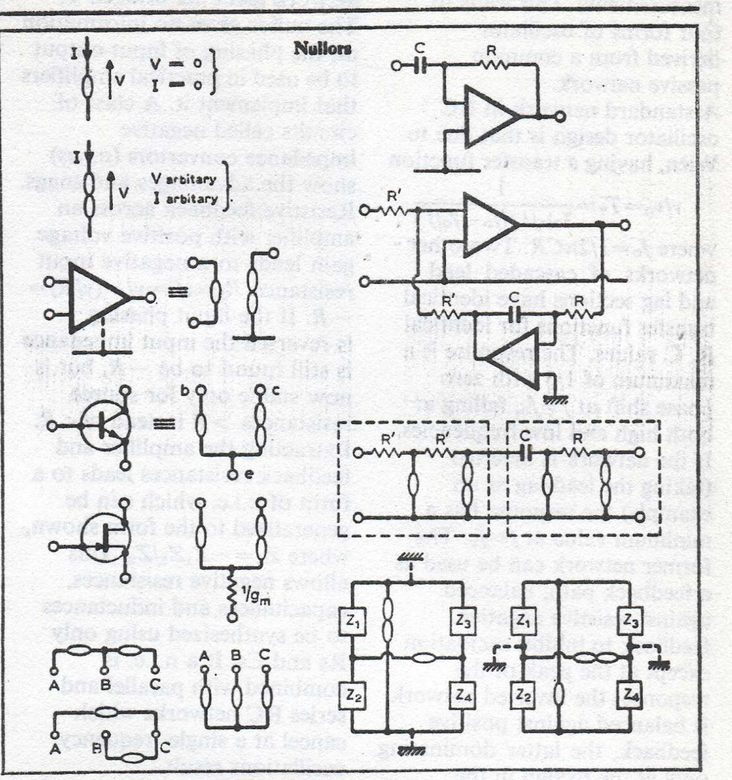
Set 25: RC Oscillators—1

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



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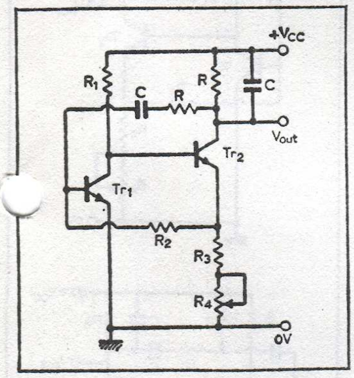
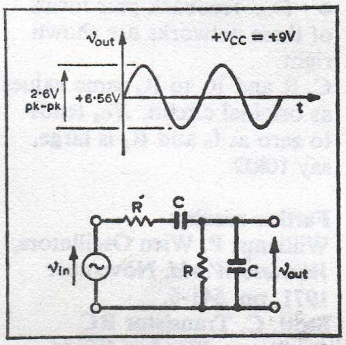
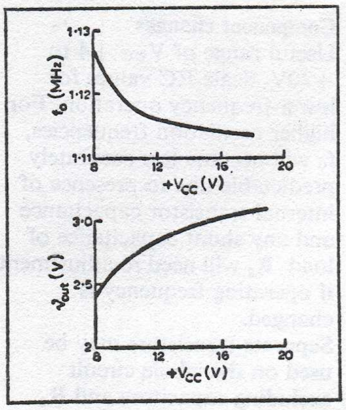
Set 25: RC Oscillators—2

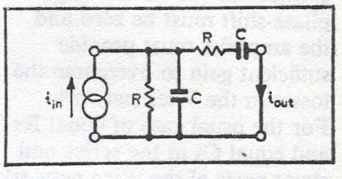
Current-driven Wien oscillator



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 **duality 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 the use of a relatively low value of R_2 provides a sufficient**

norator in changing the source and load in bridge measurements. This leads to four forms of oscillator derived from a common passive network.

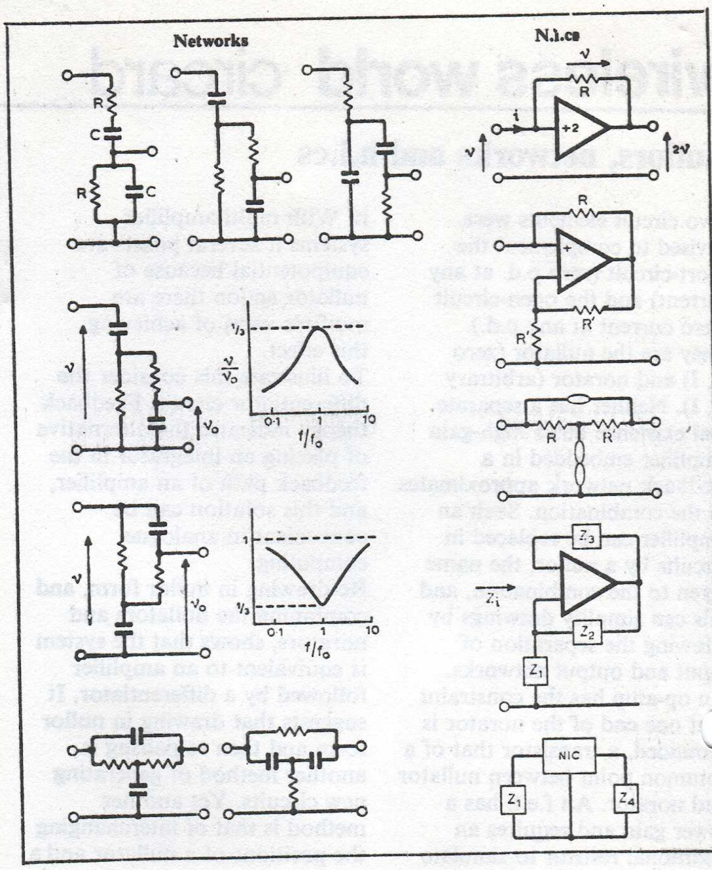
A standard network in RC oscillator design is that due to Wien, having a transfer function

$$v/v_0=T_v=\frac{1}{3+j(f/f_0-f_0/f)}$$

where $f_0=1/2\pi CR$. 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 $f=f_0$, 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 $f=f_0$. 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 generalized to the form shown, where $Z_1=-Z_1Z_3/Z_2$. 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.



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 overleaf this current gain is given approximately by $A_1=1+R_2/(R_3+R_4)$. 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 overleaf for effect of increasing supply voltage, for example, after setting the critical condition at a lower value of $+V_{CC}$.)

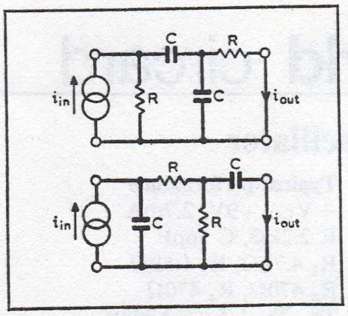
Component changes

Useful range of V_{CC} +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 on the whole circuit excluding capacitors and R_4 integrated on a single monolithic chip.

Circuit modifications

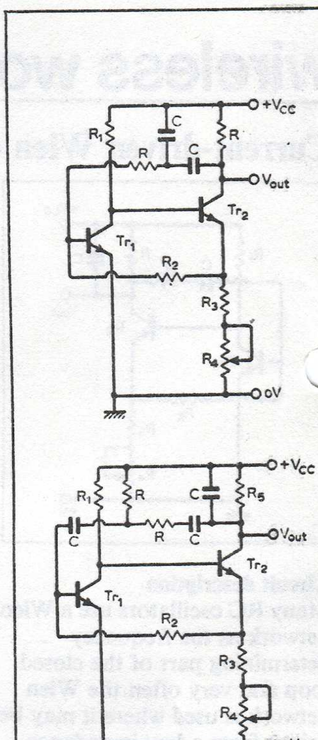
- Any other frequency-determining network which operates with an optimum source impedance tending to infinity (a current source) and an optimum load impedance tending to zero can be used in place of that shown overleaf.



Two examples are shown above.

- **D.c. feedback pair forms of these networks are shown right.**

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.

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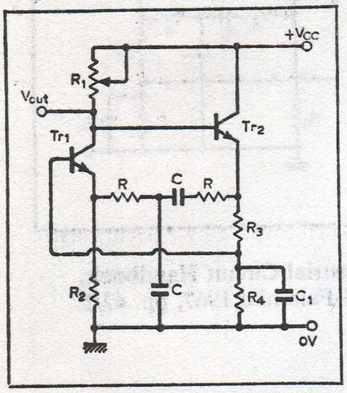
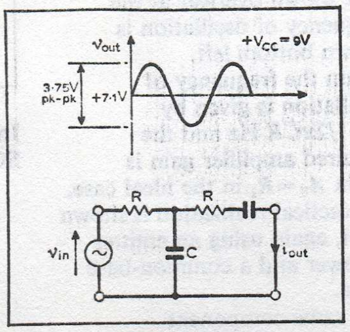
Set 25: RC Oscillators—3

Voltage-current Wien oscillator

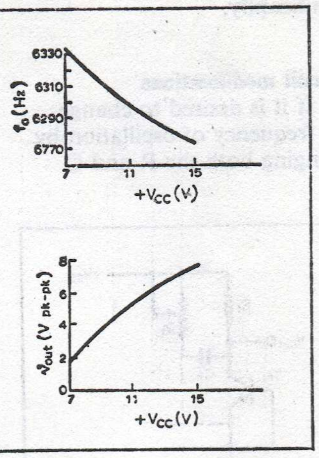


Circuit description
The above oscillator circuit uses a Wien-network which is fed from an emitter-follower and is loaded by a common-base Tr_1. This realization is an example of an RC oscillator which has the Wien-network ideally fed from a voltage source and loaded by

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)



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



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 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. In the above circuit, for example, where R was selected

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Set 25: RC Oscillators—4

Op-amp Wien oscillator

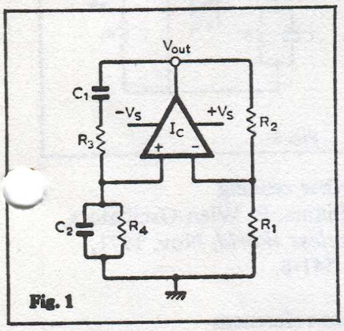
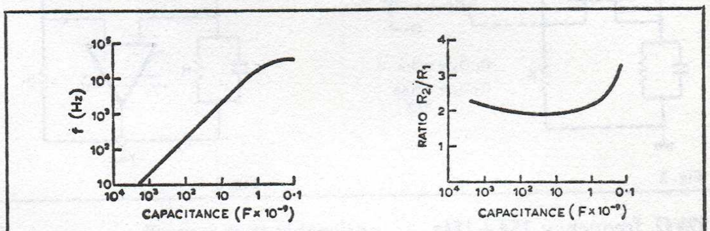


Fig. 1

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 $(R_2+R_1)/R_1$.

Typical data
$\pm V_S$ ±15V
IC 741
R_3, R_4 4.7kΩ ±5%
C_1, C_2 0.1μF
Frequency 355Hz
R_1 3.4kΩ, R_2 6.6kΩ
Harmonic distortion 0.4%
Output 26V pk-pk up to 14kHz
Drops to 16V pk-pk at 27kHz

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.

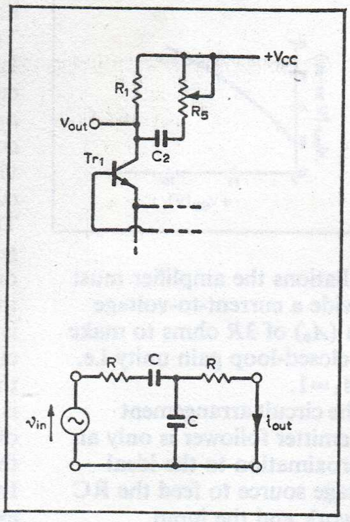
Component changes

- 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

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 $R_1\approx15k\Omega$.

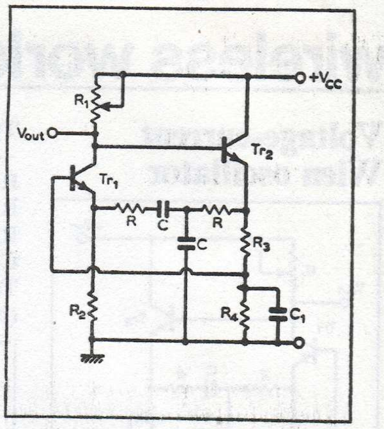
Component changes
The useful range of supply voltage is approximately +4 to +20V, but note that changing the supply will change the operating currents and hence the overall gain of the amplifier. Hence, large changes in supply voltage will change the loop gain sufficiently to either prevent oscillation 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 top left.



From an a.c. viewpoint R_5 is in shunt with R_1 provided $X_{c2}\ll R_5$ at the frequency of oscillation.
● Another network having the same transfer function as that shown overleaf at the frequency of oscillation is shown bottom 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 right, 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.

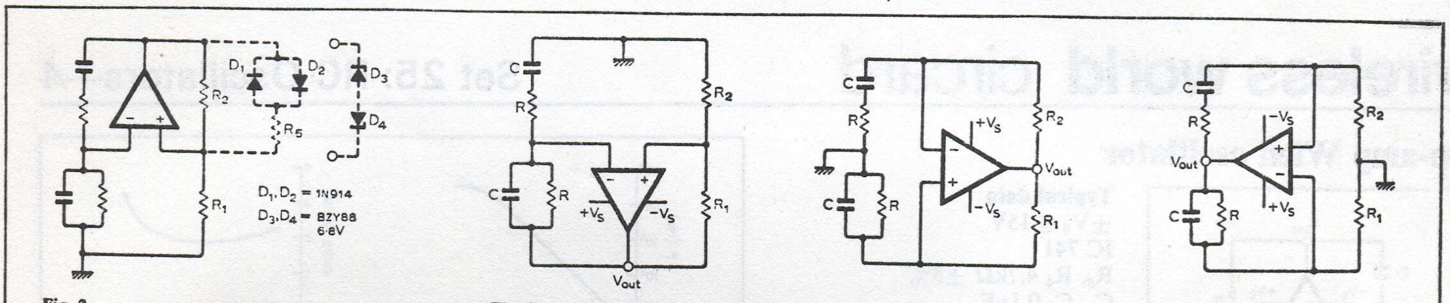


Fig. 2 Fig. 3 Fig. 4 Fig. 5

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.
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.

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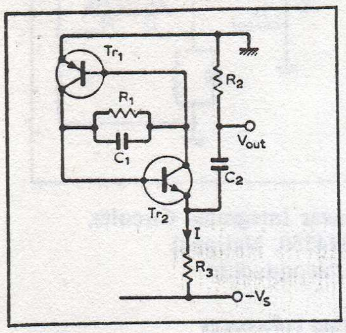
Micropower oscillator

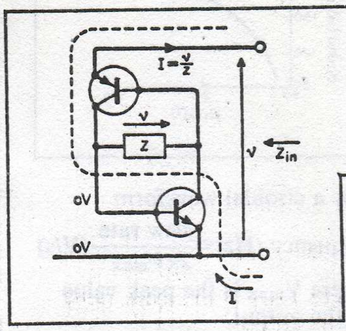


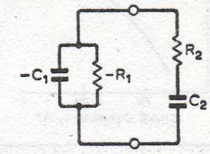
Circuit description
This is one form of a negative impedance converter circuit. It may be considered as a two terminal device, and be inserted in a constant current path that may already exist in a circuit. circuit will provide a low voltage amplitude sinusoidal oscillation (but not necessary low distortion) 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 zero giving $C_1/C_2+R_2/R_1=1$ or $R_1=2R_2$, $C_1=C_2/2$ and $f=1/2\pi\sqrt{C_1R_1C_2R_2}$.

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 ±5%
C_1 100nF, C_2 220nF
V_{out}: 54mV pk-pk
frequency 436Hz
Tr_1 BC126, Tr_2 BC125





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 onset of visible distortion.
Circuit will operate at lower currents but with lower output —see graph over.
Typical data for absolute minimum current of 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

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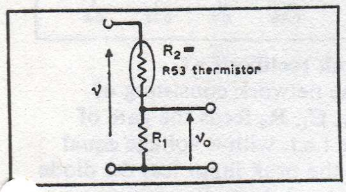
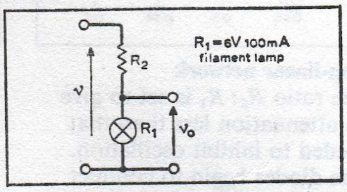
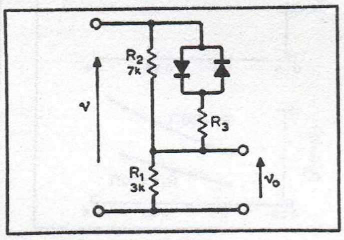
Amplitude-control methods



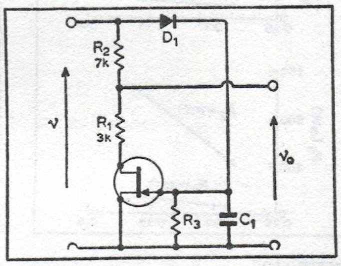
ermistor
A bulk semiconducting resistor with negative temperature coefficient. Placed in series with a resistor, and with an increasing voltage applied, the thermistor is heated by the resulting flow of current; its resistance falls and the attenuation of the network decreases. There will be a single 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.



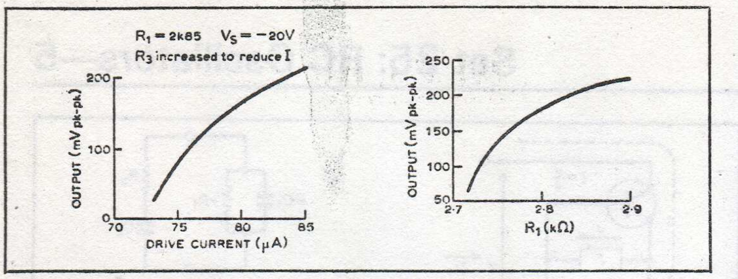
Lamp
Place in series with a resistor which is low enough to permit sufficient current to heat the lamp. As the voltage increases, the lamp resistance rises and again the attenuation decreases. The lamp can operate at a higher temperature than the thermistor and is less sensitive to ambient temperature 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.



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

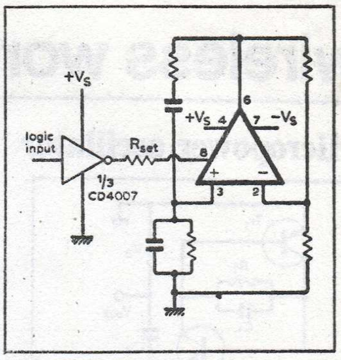
$$\text{frequency (Hz)} = \frac{\text{slew rate}}{2\pi V_{max}} \text{(V/s)}$$

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 below.



Further reading

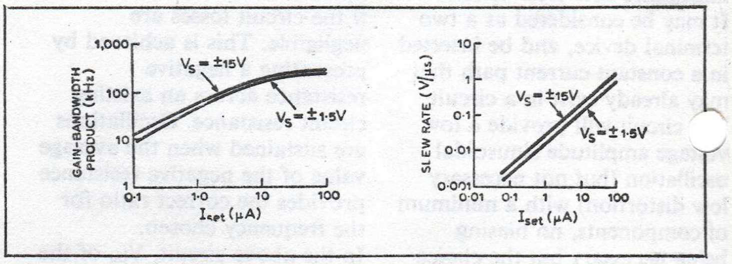
Brown, J. Equivalent n.i.c. networks, nullators and norators, *IEEE Trans.* vol. CT-14, 1967.

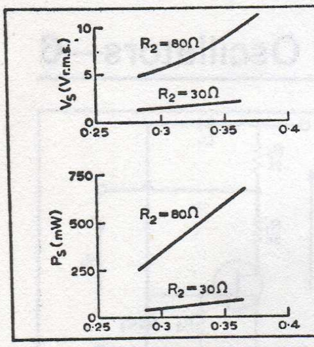
Linear Integrated Circuits, LM4250, National Semiconductor.

Cross references

Set 10, card 8.

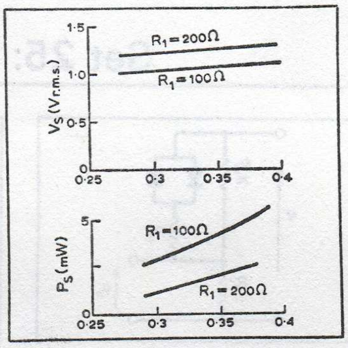
Set 25, cards 8, 1, 4.





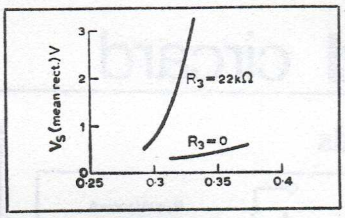
Thermistor

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.



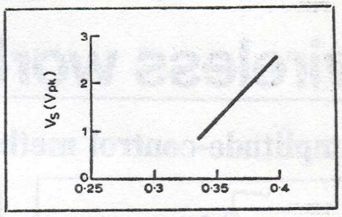
Lamp

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—1 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.

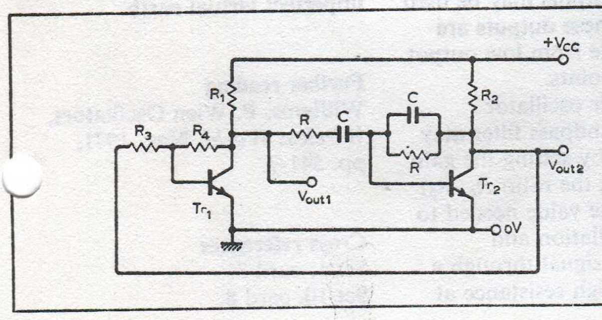


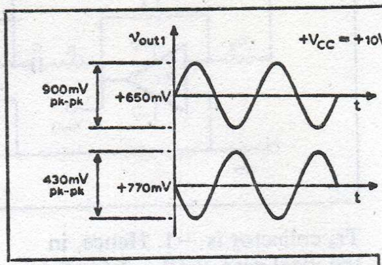
Peak rectifier/f.e.t.

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.

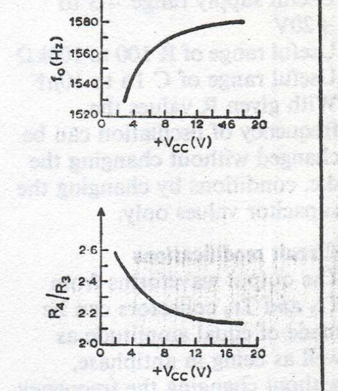
Baxandall RC oscillator

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



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)

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 then only one single frequency the serial and parallel RC networks have equivalent magnitudes and the

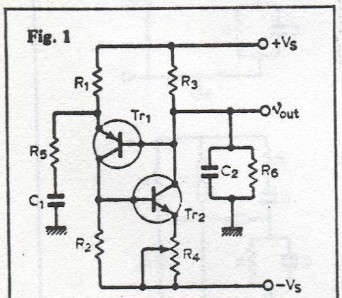


Fig. 1

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

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

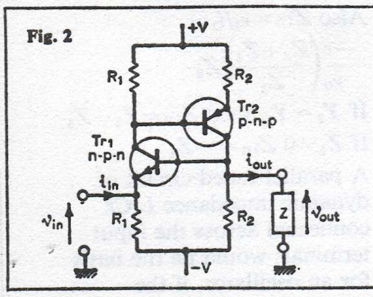


Fig. 2

of Tr_2 upwards is $i_{in}R_1/R_2$. 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

$$Z_{in} = v_{in}/i_{in} = v_{out}/i_{in} = -\frac{v_{out}}{i_{out}R_2/R_1} = -\frac{R_1 Z}{R_2}.$$

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

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.

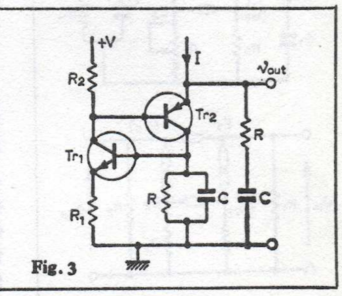
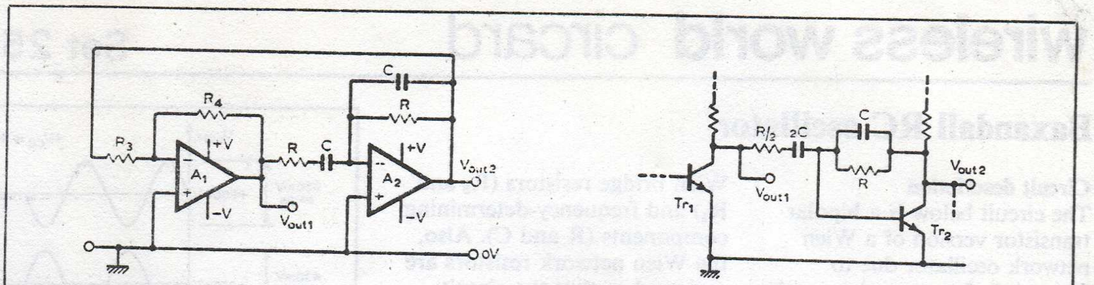


Fig. 3

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 above, showing nullator, norator interchange. Bridge oscillators 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.

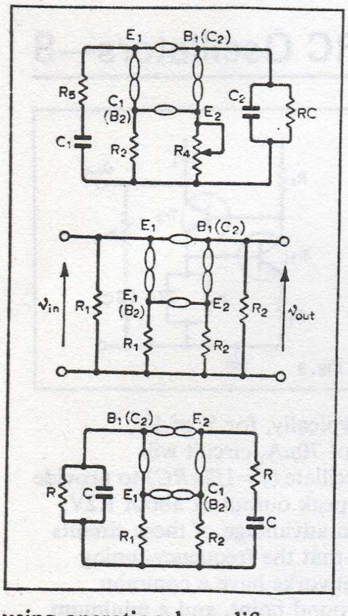
Circuit modifications
The output waveforms from Tr_1 and Tr_2 collectors can be made of equal amplitude as well as being in antiphase, without changing the frequency of oscillation, by scaling the frequency-determining RC components of the Wien network as shown below. With this arrangement the impedances 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.

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

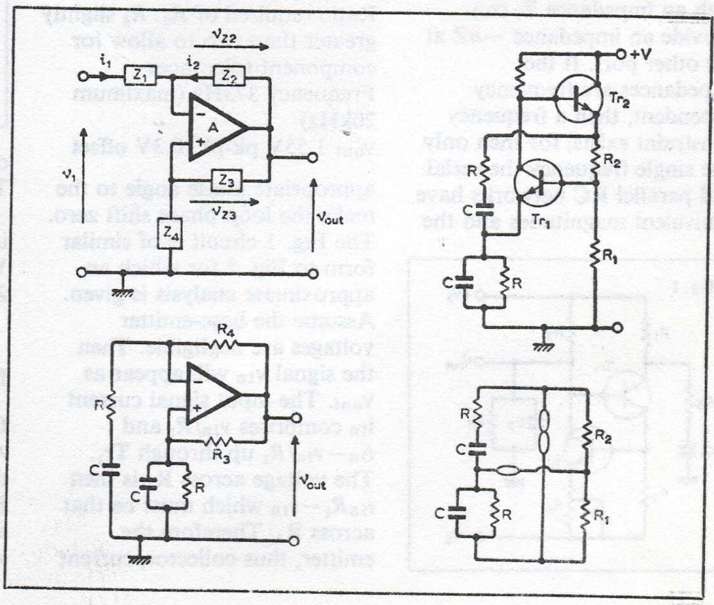
Cross references
Set 1, card 3.
Set 10, card 8.



using operational amplifiers (card 4) can also be considered as a form of n.i.c. oscillator.
For the above general case
$v_{Z_3}=v_0Z_3/(Z_3+Z_4)=v_{Z_2}$
$i_2=v_{Z_2}/Z_2=v_0Z_3/(Z_3+Z_1)Z_2=i_1$
From Millman's theorem,
$v_-=v_1y_1+v_0y_2(y_1+y_2)$
$v_+=v_0y_3(y_3+y_4)$

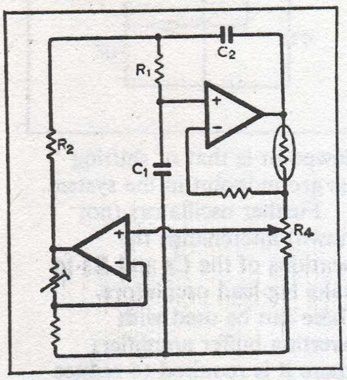
For high A, $v_-=v_+$ and
$$\frac{v_1}{v_0}=\left(\frac{y_1+y_2}{y_3+y_4}\right)\frac{y_3}{y_1}-\frac{y_2}{y_1}$$
Also $Z_{in}=v_1/i_1=$
$$\frac{-v}{v_0}\left(\frac{Z_3+Z_4}{Z_3}\right)Z_2$$
If $Y_2=Y_3$, then $Z_{in}=Z_1-Z_4$
If $Z_1=0$ $Z_{in}=-Z_4$
A parallel tuned circuit of dynamic impedance L/Cr connected across the input terminals would be the basis for an oscillator, if the magnitude of Z_4 is made equal to r. Specifically, $Z_m=-Z_4.Z_2/Z_3$. 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.

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



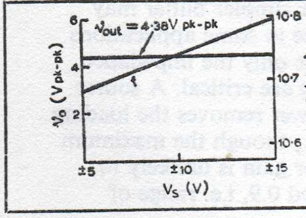
Circuit description
As described earlier the lead/lag network, C_2R_2 followed by C_1R_1, is an alternative to the Wien network. If the network is driven by two amplifiers (see ref.) then it is possible to vary the frequency of oscillation by changing the gain of one amplifier, without changing the condition 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 $k \rightarrow 1$ the frequency $\rightarrow 0$. 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.

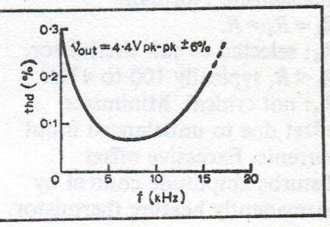
Typical performance
IC 741
Supplies ±15V
R_1, R_2 (=R) 10kΩ
R_3 15kΩ, R_4 1kΩ log
R_5 ITT thermistor R24
C_1, C_2 (=C) 1nF
f 1.5kHz to 20kHz

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

provided $R_2 \gg R_4$

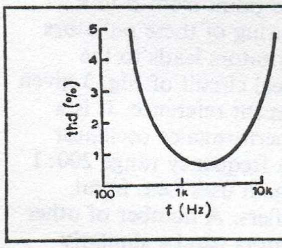
Component changes
IC: any general-purpose compensated op-amp. For lower frequencies, f.e.t. input op-amps allow the use of larger resistors. This also minimizes the loading on the potentiometer and widens the range that can be covered by variation of k alone.
Supplies: not critical. Should be appreciably greater than required peak-peak output if





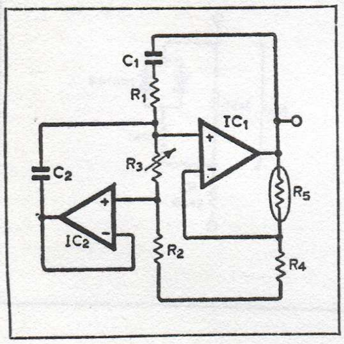
Single-element-control oscillators—2

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



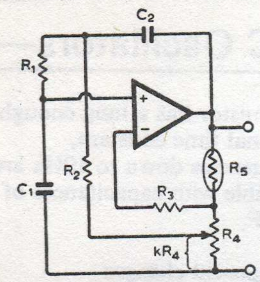
emitter or source-followers can be used. The frequency of oscillation is $1/2\pi\sqrt{R_1R_3C_1C_2}$ provided that $R_3=R_1$ and the amplitude-maintaining condition $R_5=R_4$ is maintained.

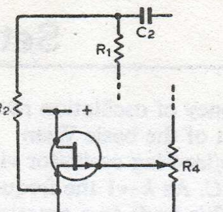
N.B. R_3 can be replaced by any other element or device that behaves as a linear resistor. If a photo-conductive cell is used the frequency becomes light sensitive and a wide range of light intensities can be covered. Oscillator amplitude stability ±1% from 120Hz to 15kHz, V_s 10V to ±15V.

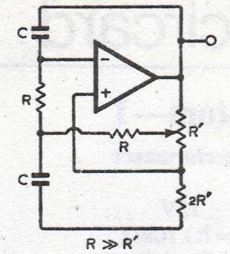


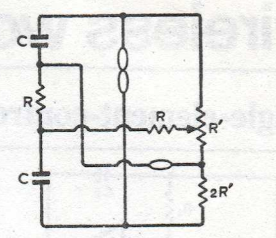
Typical performance
$IC_{1,2}$ 741
Supplies ±15V
R_1, R_2 2.2kΩ
C_1, C_2 0.047μF
R_3 2MΩ log
R_4 470Ω
R_5 ITT thermistor R54
At f 1kHz, v_o 6.4V pk-pk
T.h.d. 0.1%
f_{max} >18kHz as $R_3 \rightarrow 0$
f_{min} <120Hz as $R_3 \rightarrow 2M\Omega$

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









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 $k \rightarrow 1$ 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 but 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 $1/2\pi CR$.

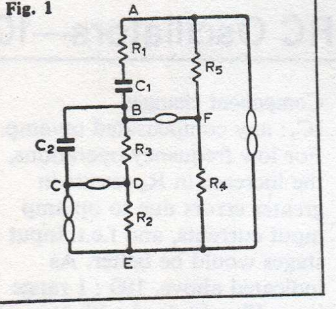
Further reading

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.

Fig. 1



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 above. 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.

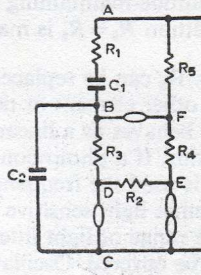
Further reading

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.

Fig. 2



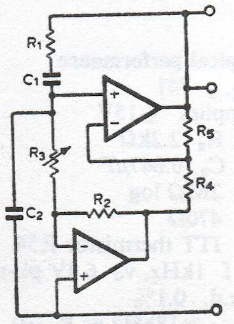


Fig. 3

Fig. 4