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Set 26: RC oscillators II—1

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 and is indicated. The oscillators may

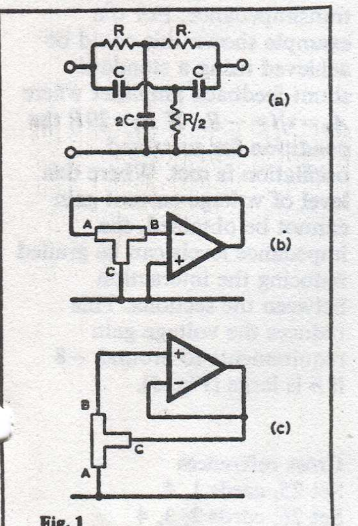

Fig. 1

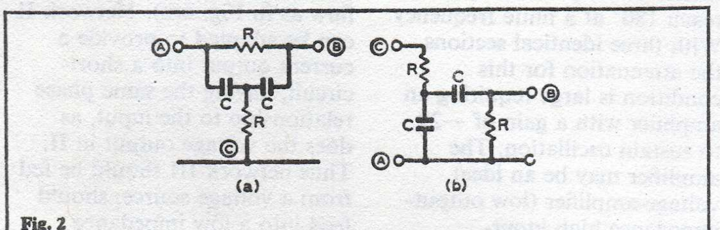

Fig. 2

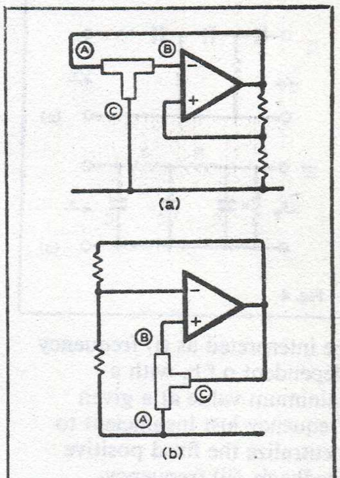

Fig. 3

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Set 26: RC oscillators II—2

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 $R_1=R_2=R$
$C_1=C_2=C$
$R_3=nR$
$C_3=C/n$

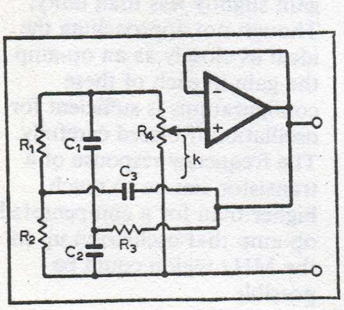
The passive network transfer function is then given by

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

where $f_0=1/2\pi RC$. For $n=1/5$ the response for zero phase-shift when $f=f_0$ is $T_V(f=f_0)\approx 1.094$.

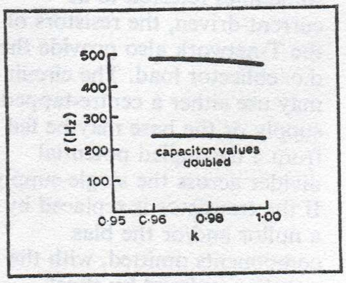
Similarly for $n=1/3$, $T_V(f=f_0)\approx 1.073$.

This suggests that the amplifier gain must not fall more than 5% or so below unity if oscillations are to be maintained.



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



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 $R_3 \to R_1/2$ 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.

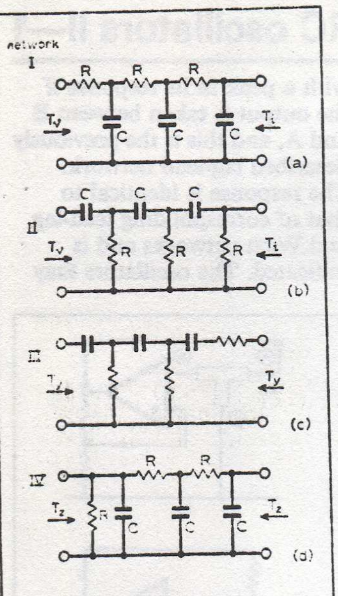


Fig. 4

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

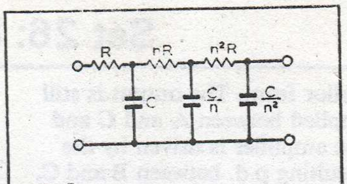


Fig. 5

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

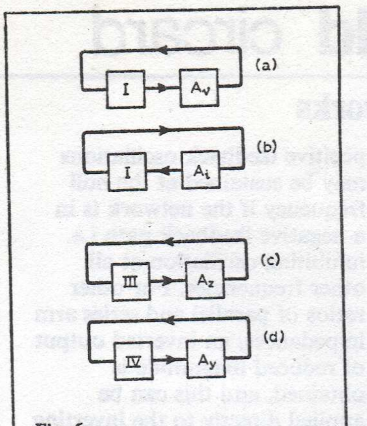


Fig. 6

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

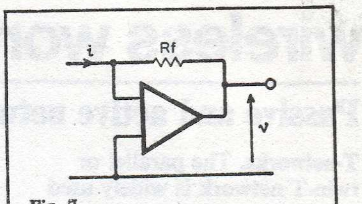
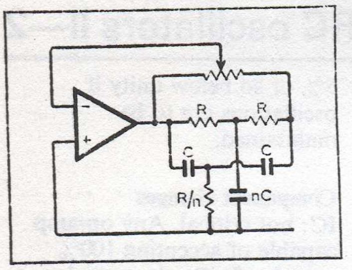


Fig. 7

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).

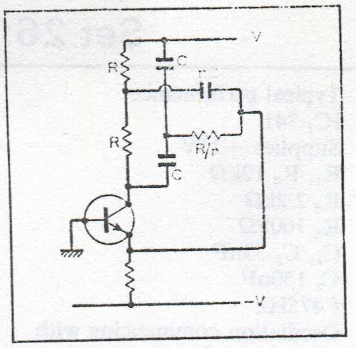
Cross references
Set 25, cards 1, 4
Set 26, cards 2, 3, 4
Set 17, card 1



n: typically 1/3 to 1/8. Frequency change in graphed results because increasing feedback brings increased distortion.

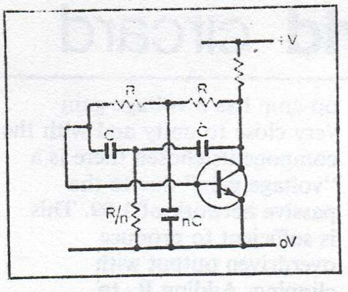
Circuit modifications

- 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 that. Re-drawing in op-amp form then leases 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.



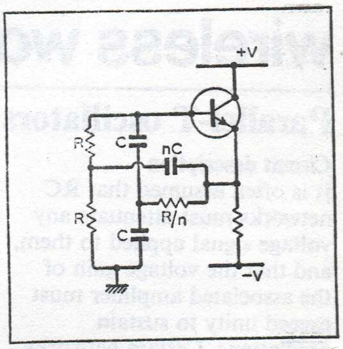
The same constraints on impedance levels apply, and the performance is very similar to the original.

- 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.

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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.

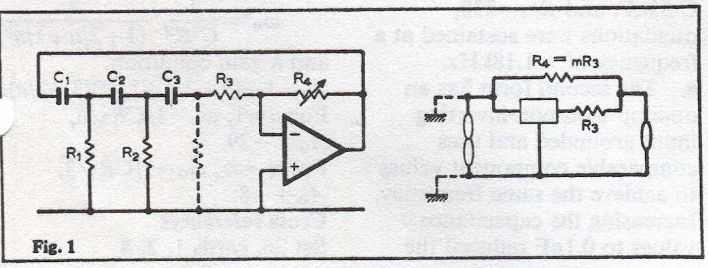

Fig. 1

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.

Typical performance
IC 741
Supplies $\pm$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.

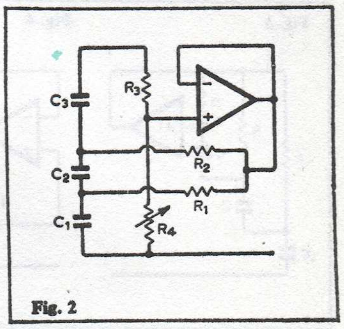

Fig. 2

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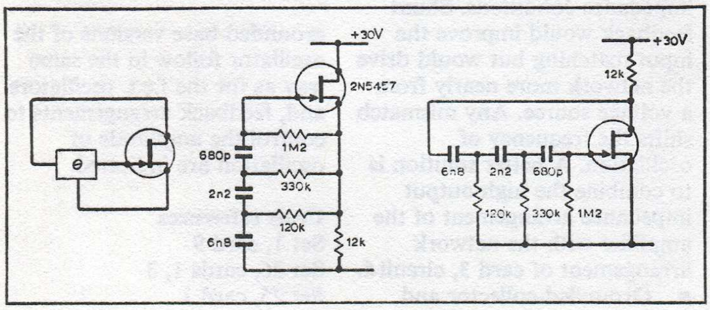
F.e.t. phase shift oscillators

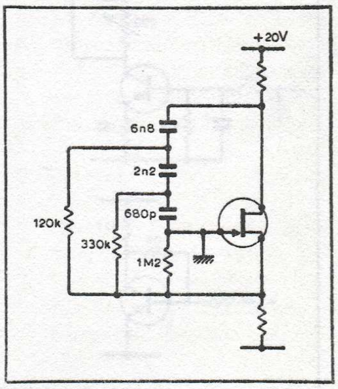
Circuit description
The basic principle is that of a passive network which has a frequency dependent transfer function such that in one configuration the output is 180° out of phase with the input at a particular frequency. The network contains a cascade of RC or CR sections and by grading the component values, the loading effects of successive sections is reduced. In the limit each section attenuates by ½ 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.





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_3: 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 $n \to \infty$ 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 $n \to 29$ is the appropriate condition. For R 10kΩ, C 33nF, and $R/n=330$, oscillations were sustained at a frequency of $\approx$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.

Fig. 3

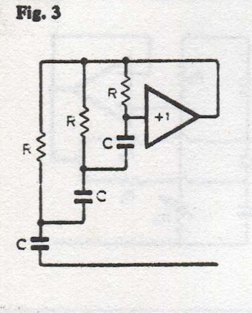


Fig. 4

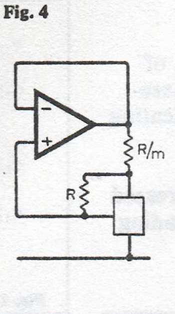
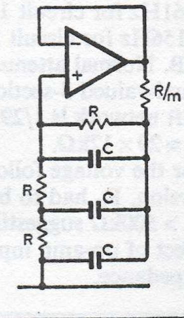


Fig. 5



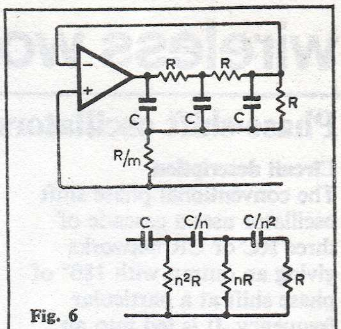


Fig. 6

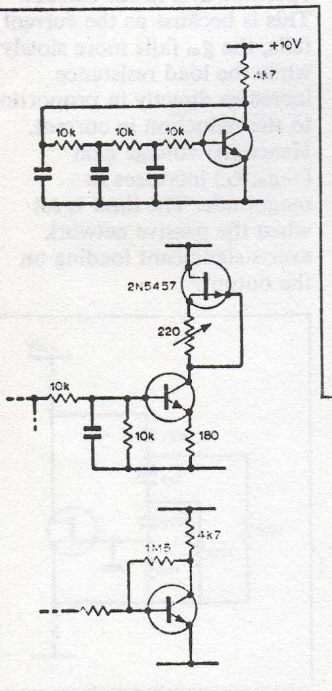
Theory. The transfer function of the CR network shown leads to a frequency of 180° phase-shift given by

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

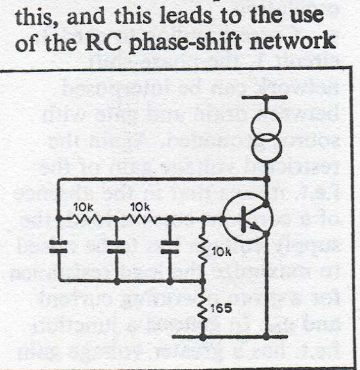
and a gain condition.
$A_0=1-(3+2/n+1/n^2)(3+2/n)$.
For $n=1$, $\omega_0=1/CR\sqrt{6}$,
$A_0=-29$
For $n\to\infty$, $\omega_0\to 1/CR\sqrt{3}$,
$A_0\to -8$.

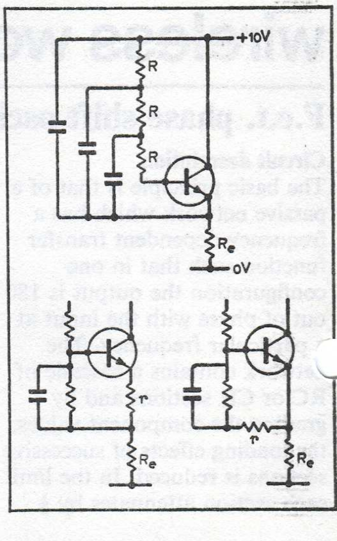
Cross references
Set 26, cards 1, 2, 4

- 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 $\approx$75Hz. Increasing the resistors to 330k, 1.2M and 3.3MΩ respectively reduced that frequency to $\approx$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

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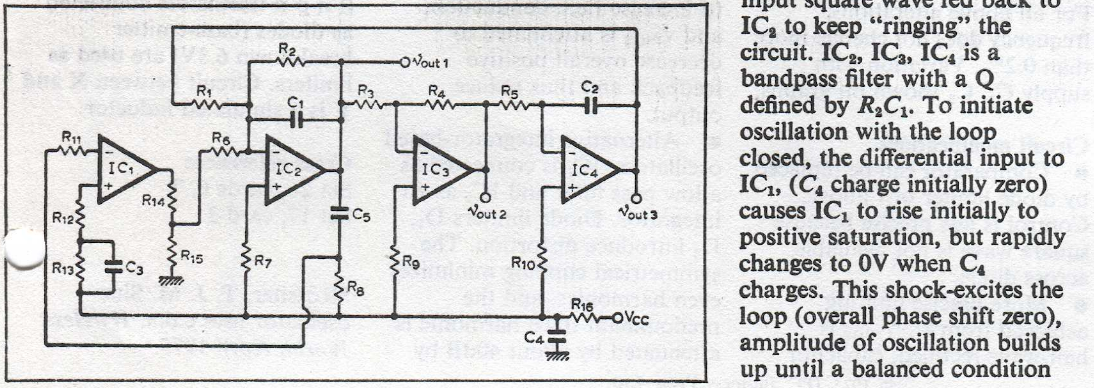
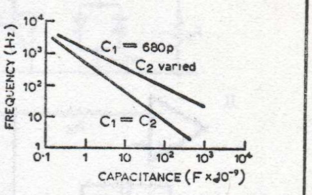
C.d.a. oscillator

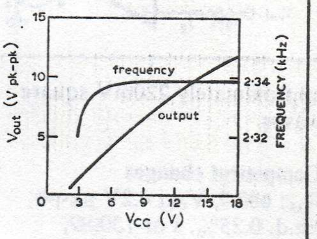
Circuit description
This circuit* takes advantage of the quad current-differencing amplifier package and is in the form of a two-integrator loop (IC_4, IC_2) plus inverter (IC_3), which does not contribute to the loop gain, but provides a 180° phase shift between v_{out1} and v_{out2}. For $R_3 = R_4$, 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,

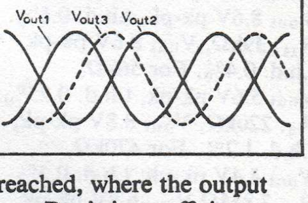
Typical data
IC_1 to IC_4 $\frac{1}{4}\times$LM3900N or MC3401
R_1, R_3, R_4, R_5, R_9 100kΩ
R_2 6.8MΩ, R_6 330kΩ
R_7, R_{10}, R_{12} 220kΩ, R_8 470kΩ
R_{11} 22kΩ, R_{13} 270kΩ
R_{14} 47kΩ, R_{15} 1kΩ, R_{16} 4.7kΩ
C_1, C_2 680pF, C_3 2.2μF
C_4 10μF

V_{CC} +10V
v_{out1} v_{out2} v_{out3} 6.9V pk-pk (≈1.3V offset)
Oscillation frequency 2338Hz
Phase difference between outputs 90° as shown in waveform diagram.
Total harmonic distortion: 0.55%.









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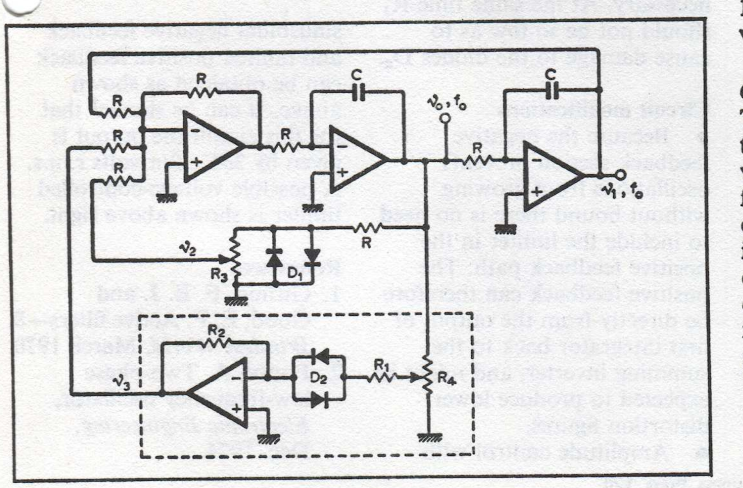
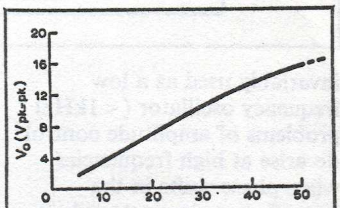
Two-integrator loop oscillator

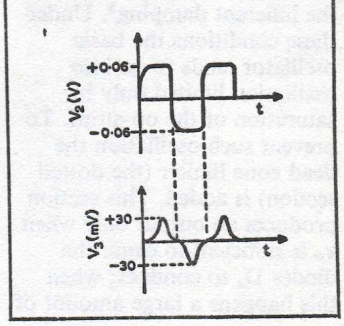
Components
Supply ±15V
IC 741 op-amps
R 10k ±5%
R_1 22k ±5%
R_2 220k ±5%
R_3 100kΩ, R_4 1kΩ
C 0.1μF ±5% polyester
D_1, D_2 1N914

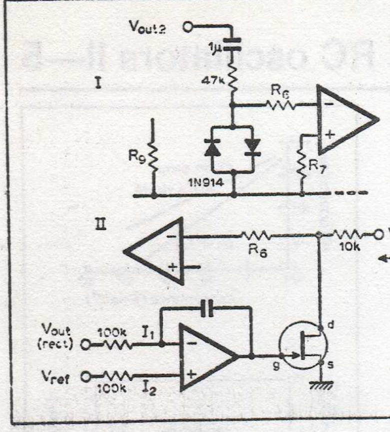
Performance
With R_3 set at 0.1, a value which caused the circuit to oscillate with slight clipping when R_4 was at zero, variation of v_o plotted against the inverse of R_4 setting is shown. Total harmonic distortion throughout this range lay between 0.96% and 0.86% and the frequency f_o was 159.35 ±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.

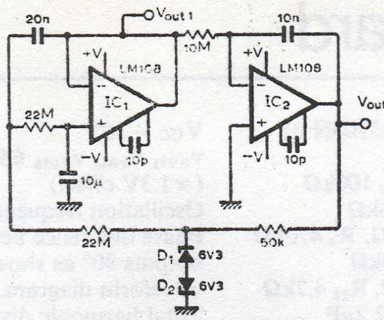
Circuit description
The circuit is a straightforward two-integrator oscillator[1] with v_2 providing sufficient positive feedback to overcome the damping inherent in the imperfect capacitors and op-amps thus ensuring oscillation. This positive feedback is of a constant nature in that once oscillation 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

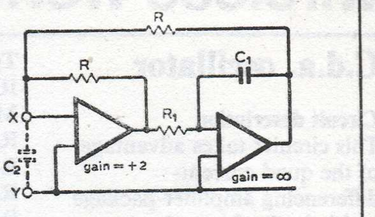












approximately 220mV square waves.

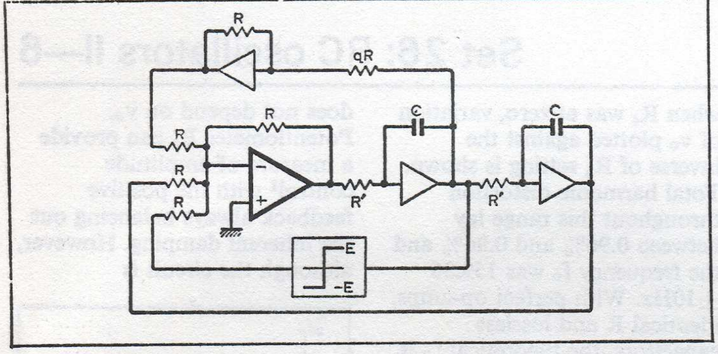
Component changes
R_{15}: 680Ω, V_{out} 4.2V pk-pk, t.h.d. 0.35%. For 1500Ω, V_{out} 8.6V pk-pk, t.h.d. 0.5%. R_{14}: 39kΩ, V_{out} 8.6V pk-pk, t.h.d. 0.4%. For 56kΩ, V_{out} 5.6V pk-pk, t.h.d. 0.25%. R_6: 220kΩ, V_{out} 8.8V pk-pk, t.h.d. 1.2%. For 470kΩ, V_{out} 4.4V pk-pk, t.h.d. 0.3%. R_2: 7.8MΩ 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.

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.

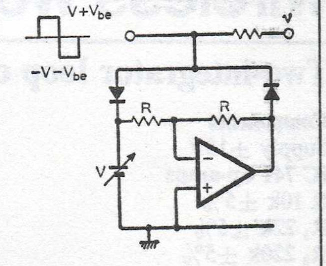


invariably used as a low frequency oscillator (<1kHz) problems of amplitude control do arise at high frequencies when phase shifts in the imperfect components reduce the inherent damping[2]. Under these conditions the basic oscillator tends to go into oscillation limited only by saturation of the op-amps. 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.

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 right.

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.

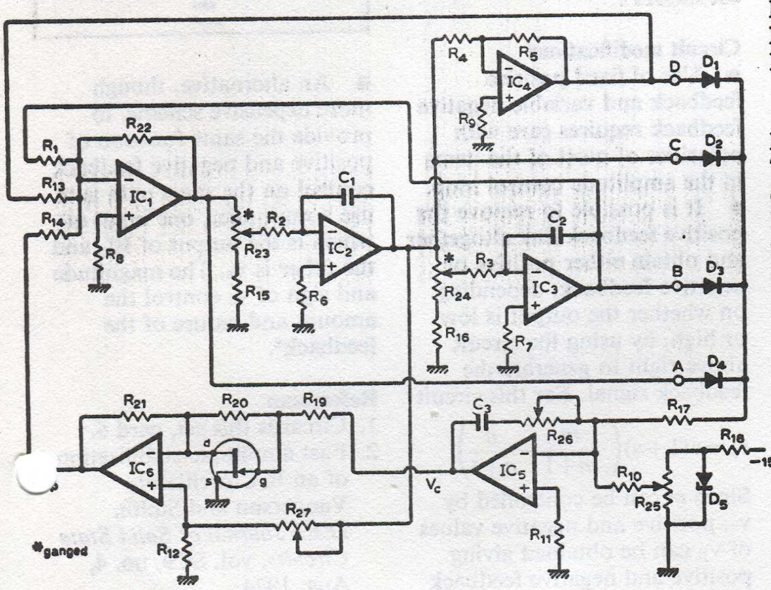
wireless world circard

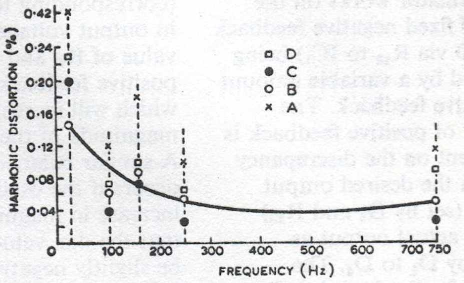
Set 26: RC oscillators II—7

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





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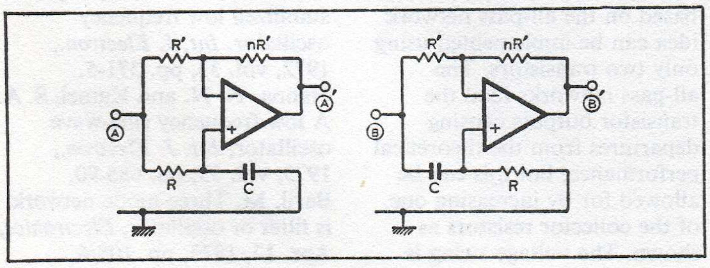
Set 26: RC oscillators II—8

All-pass network oscillator

Circuit description
Most oscillators have a peak in their look 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

$$\frac{v_0/n+v}{n+1}=\frac{v}{1+sCR}$$

$$\therefore T_{VA}=\frac{v_0}{v}=\frac{n-sCR}{1+sCR}$$

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 $\omega=1/CR$.

Typical performance
IC_1, IC_2 741
Supplies ±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 $n>1$.

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

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.

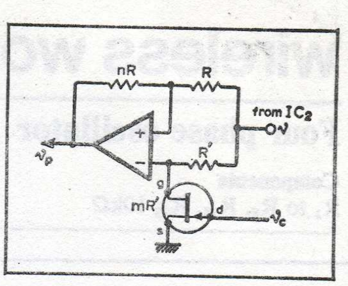
A four-phase oscillator allows rapid amplitude stabilization since any change in amplitude is quickly sensed; this is of importance in low frequency oscillators[2].

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[3].

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.

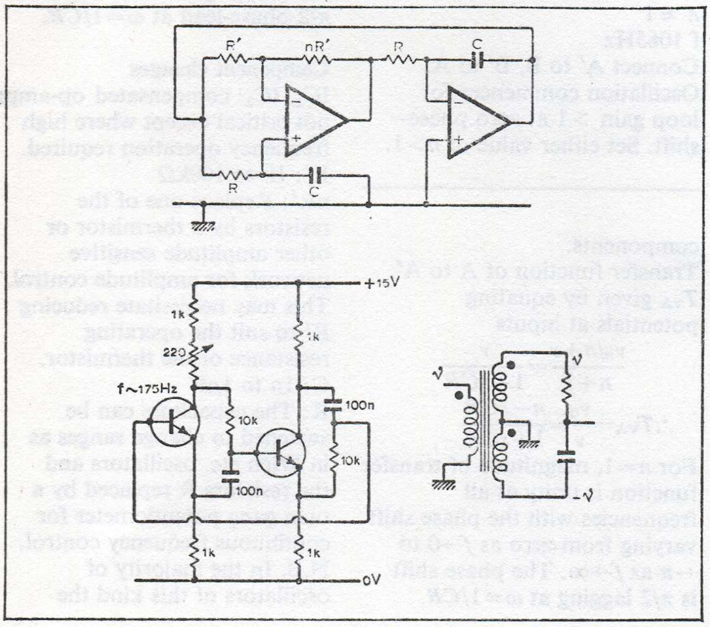
frequency of oscillation is of the form $f=1/2\pi RC$ where all Rs and Cs are equal.

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

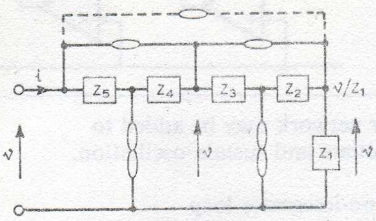
Circuit description

A gyrator is a circuit that synthesizes an impedance at one part 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. $Z_2=1/j\psi C$ then $Z_1=j\psi CR^2$ equivalent to a pure inductor having $L\equiv CR^2$. 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.

Equivalent circuit



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 produce sustained oscillation. For $Z_1=Z_2=Z_3=Z_5=R$ and $Z_2=1/j\omega C$, $Z_1=j\omega CR^2$.

Fig. 1

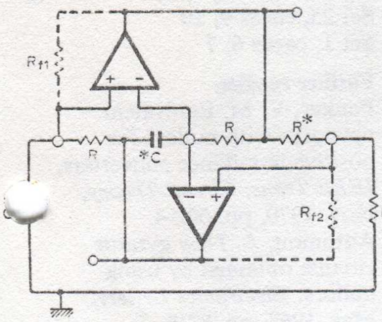


Fig. 2

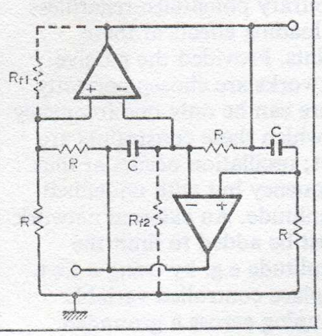
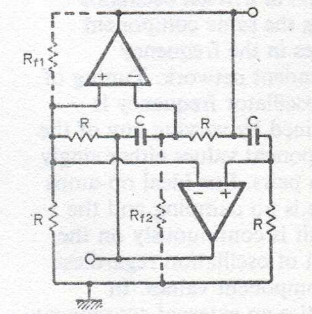


Fig. 3



Wide-range gyrator oscillator

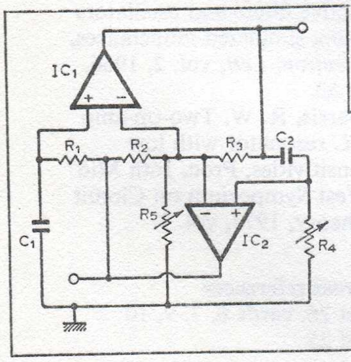
Circuit description

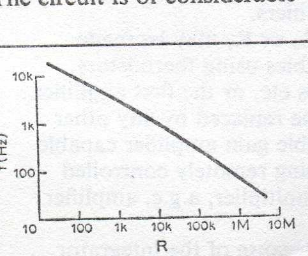
Another version of the gyrator oscillator, this one uses a single variable resistor, R_4, to control the frequency of oscillation. For ideal active and passive elements only the smallest amount of positive feedback would ensure that the amplitude builds up with peak clipping restores the loop gain condition.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

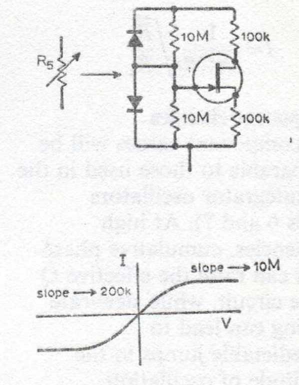
It is equally possible to change the frequency by varying both capacitors and/or a pair of resistors, making the components switched or continuously variable. This reduces the variation needed

Typical performance

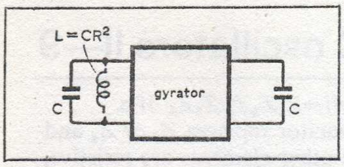
$IC_{1,2}$ 741
Supplies ±15V
R_1, R_2, R_3 10kΩ
C_1, C_2 15nF
R_4 10kΩ
R_5 470kΩ
f 1.05kHz

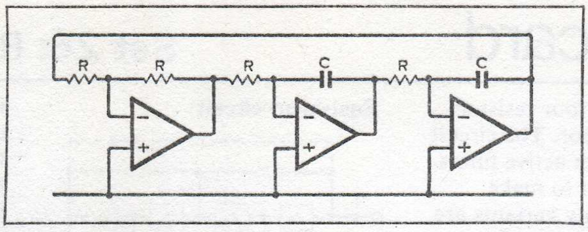


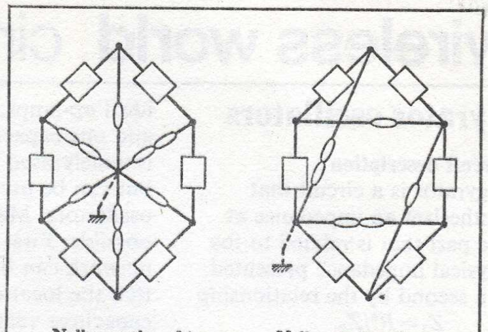




Amplitude control network







Nullor version of two-integrator oscillator

Nullor version of gyrator oscillator

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

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

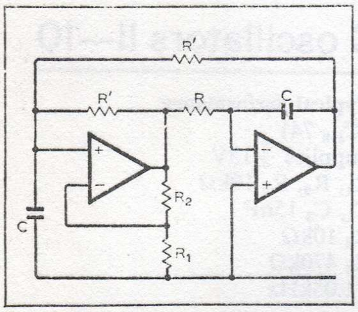
This is the same frequency as occurs in a Wien oscillator using the same component values in the frequency dependent network. Tuning of the oscillator frequency is obtained by varying any of the component values either singly or in pairs. For ideal op-amps there is no damping and the circuit is continuously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

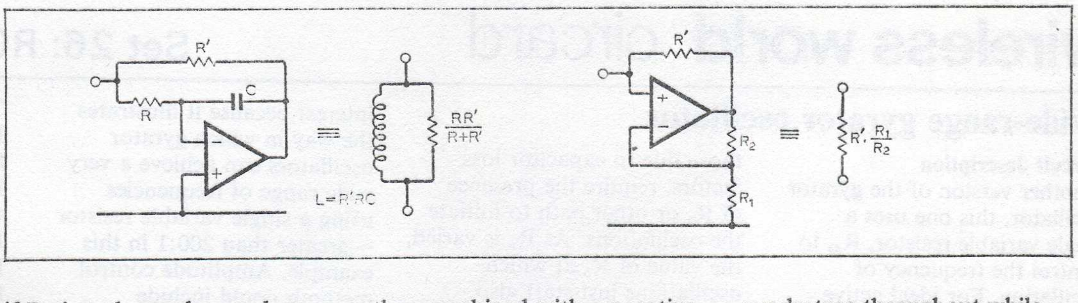
Two-integrator loop
This circuit, described on cards 6, 7, uses the same passive components and has the same frequency of oscillation. Though developed separately these two are different forms of the same basic circuit. This can be seen by re-drawing the passive networks in a similar format (where the 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.





in R_5 but does not eliminate it. In the original circuit, for $R_1=R_2=R_3=R$, $C_1=C_2=C$

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

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 100kΩ
C_1, C_2: In to 1μF
R_4: 47 to 4.7MΩ
R_5: Typically 220k to 2.2MΩ. Low resistance needed at high frequencies.
Supplies: ±5 to ±15V.

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