Generators

From output to inverting input sets the amplifier gain to overcome the losses in the passive network. Largest signal at amplifier output, least distortion across last capacitor. Make $R' = R$ to minimize loading of network. Any band-pass filter may be inserted into the feedback loop of a system, allowing the overall loop gain to reach the critical condition only at the filter centre frequency. To adapt the more usual Wien networks to current differencing amplifiers, it is sufficient to take any such network whose output has a grounded resistor, and use the current in the resistor as a feedback—the amplifier having a low enough input resistance to act as a virtual earth. No mechanism is shown for amplitude control, but the precisely defined input non-linearities allow for simple solutions as shown later.

Circuit description—1

Four types of generator are considered here. They can be summarized as sine, relaxation, square/triangle and d.c. The last of these, the group of circuits normally called voltage/current regulators, can be considered as generators of direct voltage or current. Alternatively, they can be dealt with as d.c. amplifiers with input signals provided by constant voltage/current reference elements, e.g. Zener diodes (see set 16). For cisooidal oscillators the basic requirement is a passive network giving a phase-shift of 0 or $\pi$ at a single frequency. This may or may not be accompanied by a maximum in the magnitude of the transfer function, but in either case it dictates to which input the feedback should be returned. The circuit oscillates at the frequency for which the loop has zero phase-shift and amplifier phase-shifts cause errors at high frequencies. The classic phase-shift oscillator can be used as above, with the amplifier providing more than enough gain. Negative feedback

Set 17: c.d.as—signal generation—1

Circuit description—2

The common element to many non-sinusoidal waveform generators is a Schmitt trigger.

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RC oscillators

Typical performance

$A_i, A_i$: 1 LM390

Supply: +20V

$R_1$: 220kΩ

$R_2$: 33kΩ

$R_3$: 120kΩ

$R_4$: 10kΩ

$R_5$: 2.2MΩ

$C_1$: 0.68μF

$C_2$: 10μF

$C_3$: 0.1μF

$C_4$: 1μF

$D_1$: 1N914

$f$ : 500Hz

$V_o$ : 7V pk-pk ± 0.2dB

for $V_o$ 12 to 20V.

Circuit description

The circuit has two distinct sections. Amplifier $A_1$, together with $R_4$ to $R_5$ and $C_0$, $C_2$ constitutes a complete RC oscillator, in which the positive feedback via $R_2$ exceeds the negative feedback via $R_4$. Choosing $R_2 \approx 2R_4$, the amplifier output is biased to $V_o/2$ since the d.c. feedback ensures almost equal currents in $R_1$, $R_2$. The second amplifier amplifies and filters the output of $A_1$, after a.c. coupling via $C_0$, $R_4$ and half-wave rectification via the non-inverting input of $A_2$. The addition of a reference voltage acting via $R_5$ causes the d.c. output of $A_2$ to vary the bias current in $D_1$ in such a way that it diverts a varying proportion of the a.c. feedback from the non-inverting input of $A_1$. Allowing for the time constants of the loop, the output of $A_2$ will vary until the half-wave rectified sine-wave output of $A_3$ is comparable with $V_{ref}$. The non-linearity of $D_1$'s characteristics does not introduce serious distortion, since it matches that of the input characteristics of $A_1$.

Because the amplifiers are current-differencing units, the conventional Wien bridge has been rearranged, i.e. instead of grounding $R_1$ and using the p.d. across it as the frequency-dependent feedback, the current through it is fed back, taking the amplifier input as a virtual-earth. The output voltage of $A_1$ has a d.c. term of $\approx V_o/2$ and the superposed a.c. may have a peak-peak value of up to 80% of the supply voltage before the onset of clipping. The relatively low open-loop gain of the amplifier does not allow very low distortion levels but the ability to control the amplitude of oscillation by an external direct voltage, while using only two out of the four amplifiers within the i.c.

Component changes

$R_1$, $R_4$: Ratio $\approx 2:1$ to bias $A_i$
implemented by an amplifier with positive feedback as shown above. In this amplifier the output changes state rapidly as the input passes through a given voltage level, reverting to its original state when the input returns to a different level. The difference between the switching levels is the hysteresis, and by using the circuit to monitor the voltage across a capacitor the charging current can have its direction reversed at each transition of the voltage through a switching level. The simplest arrangement is at top right where a single amplifier acts both as Schmitt trigger and as source of charging/discharging current. The current varies during the cycle according to the usual RC relationships since the output voltage has a constant value for each switched state. If the capacitor forms part of an integrator then the charging current has one of two fixed values, because the virtual earth action prevents the charging voltage across the capacitor from modifying the current. A further modification of this circuit uses an additional diode or transistor to reverse the net current flow to the amplifier by interrupting the current to either the inverting or non-inverting input. Selection of suitable resistors gives a net-charging current that reverses its direction while remaining of equal magnitude and being controlled by an external direct voltage i.e. a voltage controlled oscillator results. If the first circuit is adapted to have one stable state then it can be triggered by an external pulse into the unstable state where it remains until the capacitor has completed its charge (or discharge) sequence. The resulting monostable circuit may be edge- or level-triggered.

output $V_o/2$, $R_4$ is chosen to provide less negative feedback than the positive feedback via $R_5, R_9, C_3$ i.e. $R_4 > 3R_5, R_9, R_4 = R_5$. Together with $C_5, C_6$, defines the frequency of oscillation. $R_5, 10k$ to $1M\Omega$. $C_5, C_6 = C_7$, $1000pF$ to $10nF$. Large $R$ values can be used with small $C$ values for economy, but layout/screening need to be watched to avoid hum.

$C_7$: Not critical. $0.01\mu F$ to $100\mu F$. Higher values only required for lower load resistances. Circuit output current should not exceed $1mA$ for lowest distortion.

$C_7$: Must have low reactance at signal frequencies—at least $0.1\Omega \cdot F$ for $1kHz$. $R_4, R_5, R_9, R_4$: Control mean current in and hence slope resistance of $D$. High values to minimize required value of $C_7$. Time constant period of waveform.

$V_o$: Oscillations sustained at lower supply voltages but amplitude control difficult.

Circuit variations
- The principle of the circuit is that of an RC oscillator using $A_1$ and a modified Wien-network, followed by a signal level sensing circuit ($A_2$) that modifies the loop-gain of the oscillator feedback loop. Any other RC oscillator can be substituted for $A_1$ circuit, but amplitude control via a biased diode is optimum, as the diode and amplifier input non-linearities cancel.
- A peak-sensing circuit is an alternative, and a diode peak detector can replace $A_2$ with power control of amplitude.
- A recently described oscillator uses all four amplifiers and achieves the very low distortion of $0.1\%$.

Amplifier $A_4$ is a band-pass filter with $A_3$ setting the $Q$.

Feedback via the $100-k\Omega$ resistance sustains oscillations at the filter centre-frequency. Amplifier $A_4$ senses the mean-rectified output and controls the amount of feedback via the attenuation due to the diodes so as to keep the output as a well-defined function of the reference voltage. Resistor values are optimized to stabilize the output against temperature variation.
Voltage-controlled oscillators

Typical performance
IC: 1 LM3900
Supply: 10V
R1: 220kΩ
R2: 270kΩ
R3: R4: 1MΩ
R5: R6: 1.2MΩ
C2: 1nF
Square-wave output: 9V pk-pk
Triangular wave: 7.5V pk-pk
Frequency: 610Hz

Circuit description
A basic method of generating square/triangle waves is to re-connect an integrator and a Schmitt trigger circuit. With correct phasing the integrator output changes linearly with time between the threshold voltages of the Schmitt, causing the output to reverse and with the direction of the integrator output slope. To bring the frequency under the control of an external voltage the switched output is used to reverse the net current into the integrator, but leaving the magnitude defined by the control voltage. A current differencing amplifier is particularly suited to this technique, since the net charging current is the difference between the two input currents. If these are set in the ratio 2:1 and the larger current is switched off, the net current changes sign, remaining of equal magnitude i.e. a triangular wave results at the integrator output with peak-to-peak voltage equal to the hysteresis of the Schmitt trigger circuit.

Consider the Schmitt output positive; the diode is reverse-biased and with
R1 + R2 = R5/2, the net integrator input current is negative. Integrator output ramps negatively until the lower threshold voltage of the Schmitt is reached. Output of the Schmitt falls almost to zero and the diode conducts by passing the current in R1 to ground through the output of the Schmitt circuit. (A residual current flows in R5 since the output of the Schmitt is typically a 100mV but the current is small compared with the normal charging current.) The net charging current is now positive and the integrator output ramps positively until the cycle is completed on chargeover at the upper threshold level. The rate of

Voltage regulators

Typical performance
Supply: ±15V
IC: 1 LM3900
Tr1: BC125*
R1: 10kΩ
Vb: 7.7V
R2: 2kΩ
ΔVb: —15mV
(for ΔVb = —5V)
—25mV
(for ΔVb = —25 deg C)
*Use of reverse base-emitter junction gives good low-current breakdown. Using reverse collector-emitter raises voltage by 0.6V, changes drift by about —2mV/deg C.

Circuit description
When negative feedback is applied around a current-differencing amplifier such as the LM3900, the inverting input terminal remains at 0.6V at 0°C falling to nearly 0.4V at 75°C. In the voltage regulator shown the voltage across R1 is defined by these values, fixing the current in Tr1, which acts as a Zener diode. Amplifier input current is very small (<30nA) allowing currents between 1μA and 1mA to be adequately defined. Output voltage is the sum of the input Vb and the Zener voltage. Two sources of variation exist:
—direct variation in input voltage (Vb) as above of about —2.5mV/deg C together with a fall of 2 to 3mV for each 1V increase in the supply.
—corresponding variation in Zener voltage caused by the resulting current variation together with its own temperature coefficient. If a Zener diode is available with a positive temperature drift cancellation between that drift and the input drift is possible, though complicated by
charging is proportional to $V_{ref}$ which may be derived from a potentiometer across the supply or from an external source with $f \propto V_{ref}$ to a good approximation.

**Component changes**

$R_2$, $R_3$, $R_4$: For equal positive and negative slopes to the triangular wave (and hence unity mark-space ratio for the square wave). $R_1 + R_2 = R_3/2$. To minimize loading on source these resistors may range from 100kΩ to 10MΩ. $R_3 \approx R_2$ but may be different to use preferred values while meeting above conditions.

$C$: sets frequency of oscillation and $f \propto 1/C$ except at high frequencies where finite rise and fall times limit the response. Typically C is in to 10µF, $R_4$, $R_3$, $R_2$: control upper and lower threshold voltages and have peak-peak value of triangular wave. For $R_3 = R_4$ and $R_2 = kR_4$, the hysteresis is $\approx k$ (square-wave output). Reducing the hysteresis increases the frequency of oscillation, other components being equal. Amplifier slew-rate of 0.5V/µs makes it difficult to obtain outputs above 10kHz.

**D:** Not critical. Any general-purpose silicon diode. $V_{rs} = +4$ to +30V. $V_{ref}$: 0 to 30V (higher if $R_4$ to $R_3$ increased).

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The current variations. The use of a small-signal planar transistor with $V_{be}$ breakdown of about 7V gives partial cancellation while using the $V_{be}$ breakdown adds about 0.6V because of the forward biased base-collector diode. The additional negative drift over-compensates, for the particular device tested, but the method allows good cancellation. Conventional Zener diodes would be preferred where a quick range of voltages is required or close tolerance (but see circuit modifications).

**Component changes**

$T_{c1}$: Any Zener diode may be used in principle since the output current is limited low-current operation is an advantage, and any planar transistor may be used as reverse base-emitter breakdown typically lies between 6 and 12V, being well-defined for a particular transistor type.

$R_3$: defines diode current as $V_{be}/R_3$, where $V_{be}$ is that of the i.c. input transistor ($\approx 550$mV at room temperature, $-2.5$mV/°C). Select for Zener current of say 100µA to 2mA depending on type, down to 1µA for planar diode/transistor.

$V_4$: 4 to 30V (down to 3V in some cases), $V_5$: up to 30V. Limited by available diodes but see circuit modifications.

$I_5$: 0 to 10mA. May be boosted by adding emitter follower at output of amplifier (inside feedback loop).

**Circuit modifications**

- Additional temperature compensation is introduced by adding resistance in series with the Zener diode. Currents in the diode and resistors are almost equal since the amplifier input is small ($\approx 30$mA). Hence $V_{be} = V_{be} + nV_{be}$. Each equivalent $V_{be}$ increases the output by $\approx 0.55V$, adding a temperature drift term of $-2.5$mV/°C. (Top circuit.)

- To increase the output voltage Zener/transistor (e.g. selected for its temperature dependence) the diode is tapped onto a portion of the output voltage. The potential divider should not be significantly loaded, nor should it constitute a significant load on the output so a low current breakdown device is preferred. With a chain current of 1mA the diode current should not exceed 10mA for good accuracy. (Middle circuit.)

- Relying on the $V_{be}$ alone to define output voltage gives too large a drift with temperature for most applications. Adding a germanium transistor as shown gives first-order temperature compensation particularly for $m$ large and the difference in the $V_{be}$ values is in the range 400 to 450mV. (Bottom circuit.) The three techniques suggested may be used in various combinations as well as separately.

**Cross references**

Set 6, cards 1, 2, 7 & 10.
wireless world

Constant-current circuits

Circuit description
Many circuits used with conventional operational amplifiers are adaptable for use with current differencing amplifiers. To provide a constant-current drive to a grounded load is difficult in both cases since each amplifier has its output referred to ground so that feedback can only be taken in shunt with the load, i.e. lowering the output impedance. If the load can float then the feedback may be taken across a grounded resistor in series with the load. An alternative technique combines negative and positive feedback in such a way that at some point in the network the output impedance tends to infinity. By injecting voltages/currents at other points in the system, a load connected between that particular point and ground will carry a current independent of the load resistance. As shown, $R_4$ is used to adjust the negative feedback such that for very high load resistances, the negative and positive feedback are almost in balance. In the ideal case, this is achieved for $(R_3 + R_2)/R_1 = R/R_2$. Current in the load may then be varied in one of three ways, each of which has minimal effect on the output resistance: variation of the input voltage $V_i$; insertion of an alternative/additional voltage in series with $R_1$; variation in the tapping point of the

wireless world

Schmitts & comparators

Circuit description
The input section of current-differencing amplifiers rely on the accuracy of matching of transistors in a current mirror to define the operating levels. It is not possible to obtain the precision of switching levels by such a method as can be achieved using a long-tailed pair and more conventional circuitry. Where a moderately wide hysteresis, say, 1V or greater, is acceptable and neither upper or lower thresholds need be defined to within a small percentage, Schmitt circuits can be simply constructed offering the other advantages of positive and rapid switching. The limited slew-rate of the amplifier makes it comparable in performance in this respect to op-amps such as the 741, but the internal compensation cannot be removed to allow speeds comparable to the uncompensated op-amps (301, 748, etc) let alone high-speed comparators. The main uses are likely to be as part of signal processing systems, alarm circuits, waveform generators, etc.

Switching takes place, assuming ideal current mirror action, when the current in $R_4$ equals the sum of the currents in $R_2, R_3$. In each case the amplifier internal $V_{BE}$ of about 0.5V has to be taken into account, effectively raising both upper and lower thresholds by somewhat less than 0.5V. To a first order approximation, hysteresis depends only on the resistor values, while the input resistance may be made $>1M\Omega$. This is because the inverting stage input current of $\approx 30nA$ allows reasonable accuracy of switching levels even though the currents in the resistors are only a few microamperes. Increasing $R_3$ reduces the upper threshold without changing the lower one. Increasing $R_4$ reduces both thresholds by the same amount leaving the hysteresis unchanged. Increasing $R_4$ raises both threshold voltages proportionally,
resistor \( R \). The last method can be used to obtain a 5:1 spread in sensitivity while retaining a high output resistance. Using a single polarity supply restricts the output to one polarity (\(+\)) but dual-polarity supplies would permit bipolar outputs for handling d.c. signals without the need for blocking capacitors.

Component changes
\( V_s: \pm 4 \) to \( \pm 36 \)V. Because of losses in feedback-defining resistors load voltage swing is always less than amplifier capability.
\( R_1, R_2: 10k\Omega \) is near lower end of working range, with amplifier normally operating with input resistor of 100k\( \Omega \) upwards.
However, values required where current of several mA required in load.
\( R_3, R_4: \) If high impedance is sufficient and it is not required to trim for output impedance \( \rightarrow 0 \), replace by fixed resistance (\( \approx R \) for \( R_1 = R_2 \).) Negative output resistance attained if feedback resistance to inverting input is too large.
\( k: \) Can be varied from 0.2 to 0.9. At high values, relatively large currents obtainable but with some loss of accuracy.
\( V: \) Variable over range 2 to 10V.
For input voltages too low, \( V_{\text{in}} \) errors at input prevent linearity of 1V against \( V \).

Circuit modifications
- By adding transistors, the feedback may be taken from across one emitter resistor as would be the case for a voltage amplifier (top circuit.) The output is taken from the transistor collectors each of which carries a collector current almost equal to its emitter current if the transistors have high current gains.
In the simplest arrangement the current is defined by the amplifier internal \( V_{\text{be}} \) with a consequent temperature dependence to the currents (which may be set to any required ratios by varying \( R_1, R_2, \) etc.) Any of the techniques used in voltage regulators may be used to achieve better control over the voltage across \( R \).
- A voltage-controlled current sink as shown forces the voltage across the 1k\( \Omega \) resistor to match the input voltage giving an overall transconductance of 1mA/V. Bias provided by the 1M\( \Omega \) resistors linearizes the characteristic at low \( V_{\text{in}} \).
- A current source using this principle is possible but has some disadvantages. It is the voltage across \( R_1 \) that needs to be defined, but approximate equality of the p.n.s across the 1M\( \Omega \) resistors that is achieved. Where the voltage across \( R_3 \) is to be low (say 1V) to maximize load voltage, small unbalance in the input current sensitivities produce small voltage differences at \( A \) & \( B \). These small errors then represent a larger error in the small voltage across \( R_3 \) and hence in the load current. (Bottom circuit.)

Cross references
Set 6, cards 4 & 12.

leaving the hysteresis in the same ratio to each as prior to the increase.

Component changes
Supply \( V: \pm 4 \) to 36V. Usual range for given i.c.
\( R_1: 10k\Omega \) to 10M\( \Omega \). The lower value is particularly useful if voltage comparator action is required for voltages \( < 1V \).
Each input is pre-biased to \(+V\) through 1M to 5M\( \Omega \) resistors.
\( R_2: \) As above. If used as a comparator \( R_2 \) is absent, and comparison is possible between unequal voltages i.e. the output can be made to swing between its limits as \( V_{\text{in}} \) passes \( V_{\text{ref}}/n \) if \( R_2 = R_{\text{ref}}/n \). Comparison of negative voltages are also possible provided bias network accommodates common mode currents. Since currents are being compared, the use of suitably large resistors allows comparison of very large positive or negative voltages, but with the limited accuracy implied by the current-mirror input.
\( R_3: 1M \) to 22M\( \Omega \). Sets

hysteresis and would be absent for comparator action. Can generally be added to any comparator arrangement to introduce, for example, small amounts of hysteresis to prevent output jitter due to noise.

Circuit modifications
- The position of the input and reference voltages (\( \pm V \) used in the circuit overleaf) may be interchanged to give a non-inverting Schmitt. The availability of inverting and non-inverting Schmitts is of importance in the design of such circuits as wave-form generators. The thresholds and hysteresis are approximately:
\[ V_{\text{in}} = \pm V \frac{R_{\text{ref}}}{R_1} \]
\[ \text{hysteresis} = \pm V \frac{R_2}{R_1} \]
- If a p-n-p long-tailed pair is added at the input the resulting amplifier may be used as a comparator or as a Schmitt (as shown). The input impedance is extremely high if high-gain low current transistors are available, and any degree of matching for the input transistors, may be chosen for high precision. It is necessary to clip the output voltage swing more precisely (with a resistor-Zener combination for example) if the hysteresis is to be equally precise.
- If two amplifiers are used then the matching is that between corresponding inputs which can be better than between \(+\) and \(-\) inputs of a single amplifier. One amplifier plus the output transistor has overall negative feedback (transistor is an inverter) and the equal current forced into the two non-inverting inputs forces the transition point in the second amplifier output to occur for equal currents at the two input currents. Again comparison can be made between unequal voltages but more account has to be taken of the effects.

Cross references
Set 2, all cards.
Set 17, cards 3, 7, 8 & 9.
Astable multivibrators

Circuit description
The amplifier, together with resistors \( R_2, R_3, R_4 \) constitutes a Schmitt trigger. The output switches between the two extremes (\( \approx \pm 100\text{mV} \) and \( \approx 0.8\text{V} \) below \( V^+ \)). When the output is low \( C_1 \) is discharged through \( R_1 \) until the lower threshold of the Schmitt is reached. The output switches rapidly to its most positive value and the capacitor is charged positively. When the upper threshold is reached the Schmitt output is reversed and the cycle restarts. If \( R_2, R_3, R_4 \) are chosen to make the threshold voltages \( V_+/3 \) and \( 2V_+/3 \) respectively then the voltage across the capacitor swings between these limits, while the output approximates to a square wave with almost unity mark-space ratio. For the highest accuracy a number of conditions must apply: The current in \( R_2 \) must be very low compared with the charging current through \( R_1 \); the output voltage swing should be as nearly as possible equal to the supply voltage; and the accuracy of the current-differencing action should be high (this places a restriction on the minimum currents in \( R_2, R_3, R_4 \)). If these constraints are met then both the frequency and the mark-space ratio remain substantially constant over a wide range of supply voltages. Any increase in supply voltage produces proportional and

Monostable multivibrators

Circuit description
The circuit is closely related to the pulse generator modification of the astable circuit, and to the unjunction equivalent (card 7). The feedback network is designed so that the circuit normally latches into one particular state, with the output high. This is because in the low stage the diode inhibits current flow into the inverting input. Current in \( R_3 \) drives the output high initially and with the sum of the currents in \( R_3 \) and \( R_4 \) exceeding that in \( R_2 \). The first negative-going pulse drives the output temporarily low, the diode becoming reverse-biased. The capacitor discharges through \( R_1 \) with only \( R_2 \) contributing any current to the non-inverting input. Discharging continues until the voltage across the capacitor is down to a fraction of the supply given by \( R_2/R_1 \). At this point, the inverting input current comes into balance with that at the non-inverting input, the output begins to rise and the positive feedback causes the cycle to complete itself rapidly. The diode conducts, rapidly changing the capacitor to a voltage or so below the positive supply voltage that the circuit is ready for the following trigger pulse. The positive-going slew-rate limit of 0.5V/\( \mu \)s limits the minimum pulse-width that can be defined with any accuracy for a 20-V supply and a pulse of nominal length 100\( \mu \)s this slew rate should add up to 45% to its effective length.

Component changes
\( +V \): +4 to +36V. Monostable action with increasing pulse-width and lowered output is possible to below 2V with some samples though outside specification. \( C_1 \): Large values may be used
hence self-cancelling increases in the magnitude and rate-of-change of voltage across the capacitor. The potential difference across the capacitor is uni-directional permitting the use of electrolytic capacitors for low frequency generation.

Component changes

$R_1, R_2, R_4$: These must be chosen jointly such that the firing points of the Schmitt trigger they constitute are $V_{cc}/3$ and $2V_{cc}/3$ (see circuit description). $R_3 = R_1 = 3R_4$ gives this condition. Make resistors as high as possible to avoid loading $R_1$, $C_1$ charging circuit i.e. $R_2 > 100R_1$. Typically $R_1$ 1MΩ to 10MΩ.

$R_1$: This sets oscillator frequency together with $C$. Too low a value loads the output, too low is loaded by $R_2$. Typically 4.7kΩ to 100kΩ.

$C$: Max frequency of oscillation $\approx 20\text{kHz}$ corresponding to $C = 470\mu\text{F}$ with $R_1 = 15\text{kΩ}$. Capacitor voltage always positive allowing use of electrolytic to give low

$R_2$: When the output is low the diode is non-conducting, and the small current in $R_2$ gives a long discharge time, i.e. a long space. (Middle circuit.)

$R_3$: Other forms of multivibrator may require more than one amplifier but the bottom circuit has some similarity of behaviour to the programmable uni-junction. The capacitor charges through $R_1$ until the upper threshold is reached. The output swings towards zero rapidly discharging the capacitor. The long mark has its duration controlled by the direct voltage applied to $R_1$. ($4V$ may be replaced by an external control voltage giving a voltage controlled oscillator). The space is of short duration because of the relatively large discharge current.

Cross references

Set 3, cards 2, 4, 6 & 8.
Set 8, all cards.
Set 10, card 2.
Set 17, cards 3, 8 & 10.

(provided leakage resistance is 1MΩ) and with $R_1$ at 1MΩ time-delays of minutes are possible.

$C_2$: Used to differentiate input signal to provide short-duration positive transients. 47p to 220pF for first input pulses; increase for slower input rise times.

$R_2$: Ratio $R_2/R_3$ sets fraction of supply voltage across $C_2$ at which the pulse period is completed. Typically $R_2/R_3 = 0.1$ to 0.3. If ratio is too low the period lengthens considerably but is ill-defined.

$R_3$: Current in $R_3$ should not fall below 1μA if timing is to be controlled by RC values.

$R_4$: With $R_4 = R_1$, the smaller current in $R_4$ is sufficient to keep the output high in the quiescent state of voltage.

$D_2$: A general-purpose silicon diode.

Circuit modifications

$+$ Positive-going output pulses can be obtained with a similar circuit if the permanent bias is applied to the inverting input, the direction of the diode reversed and a changing path provided in parallel with the diode. The quiescent condition has the output low with the diode conducting and significant current only in the 10MΩ resistor. A positive-going input transition drives the input high, which state is held until the capacitor charges sufficiently to allow the current in the inverting input to exceed that in the non-inverting input. A major advantage of an amplifier having both inputs accessible is that the signal pulses can be transferred to the other input. In this version, the output pulses would then be triggered by negative rather than positive transitions, while in the previous case which required negative transitions the response would then be to positive edges.

$+$ Two-amplifier monostable circuits can have two opposite polarity outputs and allows separation of the triggering timing and output sections more readily. The circuit shown is equivalent to a Schmitt trigger followed by a monostable although the functions are combined in a single circuit. Out1 is normally low contributing no current to the inverting input of the lower amplifier. Hence Out2 is only driven low when $V_{in}$ reaches $\approx 0.8V_b$. Provided the input falls below this level prior to the end of the pulse then generated, the circuit is suitable for resetting an oscillator when its ramp output exceeds a given d.c. level.
Flip-flops

Circuit description
The circuit has strong affinities with standard two transistor flip-flop and could be triggered at either of the inverting inputs via diodes to provide a set/reset function. Non-inverting inputs provide the additional facility of triggering the flip-flop to provide a \( \pm \) 2 action. Equal currents are fed into the non-inverting inputs on receipt of a positive-going edge but that amplifier whose output is already high because its inverting input receives no current will be initially unaffected. The other amplifier receives a short-duration current flow at its non-inverting input that exceeds that at its inverting input, and its output begins its positive-going transition. The speed-up capacitor couples this to the inverting input of the other amplifier and the positive feedback around the loop causes the transition to complete itself. Thus each output changes state in opposite directions on the receipt of each positive-going input transition. Two such transitions are required to complete each cycle of the output, and the output waveforms are complementary square waves at half the frequency of the input.

Such circuits may be readily cascaded to produce lower frequency to develop simple counters. Their low speed offers no competition to c.m.o.s., t.l.l. etc. but the ease with which they can be combined with other analogue/digital functions

Staircase generators

Circuit description
A staircase waveform can be generated by feeding pulses of current to a capacitor between pulses. The resulting circuits are closely related to those used in frequency meters such as the diode-pump circuit and its variants. One version as shown feeds the pulses to either the inverting or the non-inverting input of the first amplifier depending on the state of the output of the second. Consider output 2 positive. The diode is reverse biased and with \( R_1 = R_2 \) the net positive current into amplifier 1 is positive. During each positive pulse output 1 ramps positively, remaining relatively constant between pulses. The smaller the mark-space ratio the more nearly the output waveform approaches a staircase waveform, with the rising steps occupying a small proportion of the time. At some positive voltage set by the resistors \( R_4, R_5 \), the Schmitt trigger operates, its output switching to near ground level. The diode becomes forward-biased diverting all subsequent positive pulses from the non-inverting input. The net negative input current via \( R_4 \) produces a staircase waveform in the negative direction until the lower threshold voltage of the Schmitt is reached and the cycle recommences. In essence the circuit is that of a square-triangle generator in which the direct control voltage is replaced by a succession of positive pulses giving a stepped characteristic to the triangular wave. The mean input voltage determines the mean charging current and the average slope of the ramp. If the frequency of the input signal is varied while maintaining a constant pulse height and constant mark-space ratio, then the basic repetition rate of the output remains unchanged. (It is composed of a number of steps dependent on the ratio of input frequency/output frequency.) Varying the pulse height, the mark-space ratio or Schmitt trigger points varies the output frequency.

Component changes
\( R_4, R_5 \): The magnitudes of
Component variations
+V: Standard supply range +4 to 36V. Samples may operate down to <2V.
R₁, R₄: Sensitivity can be controlled by varying R₁, R₄ for fixed values of R₂ and R₃, and vice versa. Values are not critical and may be adjusted to suit a wide range of input pulses. 10k to 1MΩ.
R₂, R₃: These provide the current to keep the opposing amplifier saturated and are again not critical. 330k to 10MΩ could be used but as in any pulse system high resistances can be accompanied by spurious triggering. Generally the lower resistances are required at low supply voltages.
C₁, C₂: Speed-up capacitors ensure positive transitions but the maximum speed is limited by amplifier slew-rate.
C₄: Not critical. 47p to 470pF.

Circuit modifications
© Using a single amplifier, one

these resistors are not critical and they may range from 10k
to 10MΩ. Design procedure for
the trigger levels is given on
Schmitts and comparators card.
Hysteresis ≈ VR/R₄.
R₁, R₄, R₃: For comparable
positive and negative slopes,
R₄ + R₃ = R₄/2. Ratio R₂/R₃ is
not critical but neither should be
< R₂/10. R₄: 10k to 1MΩ.
C₄: Controls the basic repetition
rate of the output for given
input waveform. 220p to 10μF.
Output frequency can be very
much less than input frequency
with division ratios in excess of
100.

Circuit modifications
© When the input is at zero,
current flow to ground through
R₁ allows the induced voltage on
C₂ to vary. Inserting a diode
prevents this at the expense of
a reduced effective pulse height.
Most other v.c.o. circuits can
be adapted to perform this
function. Replacing the diode
by a transistor and controlling
the pulse to the inverting input
leaves the phasing correct. (Top.)

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arrangements are possible by
phasin(g) feedback
appropriately. Switching and
timing actions can be combined
in the single monostable shown,
where the positive output pulse
is initiated at a particular input
amplitude, the timing remaining
correct if the input voltage to
the monostable falls below this
firing level before the completion
of the monostable output pulse. (Middle circuit.)

© Circuits designed for
tachometer applications give an
inherent staircase output.
Further, the step height is
controlled only by the input
pulse height and the value of
C₄. Provided that the input
pulse width is sufficient to
allow full charge and discharge
of the capacitor C₄. The version
shown has a frequency
doubling action with the
positive input transitions being
directed to the non-inverting inputs
and the negative transitions to
the inverting inputs.

Cross references
Set 3, cards 3 & 7.

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