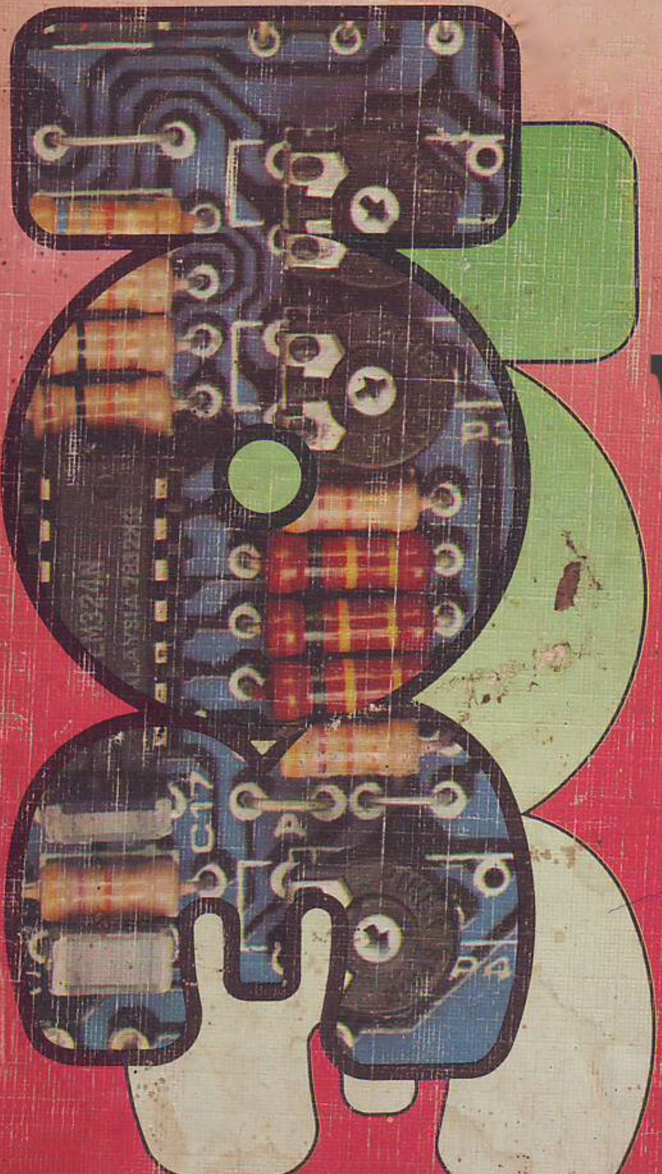
301 circuits

Circuits 79 to 179



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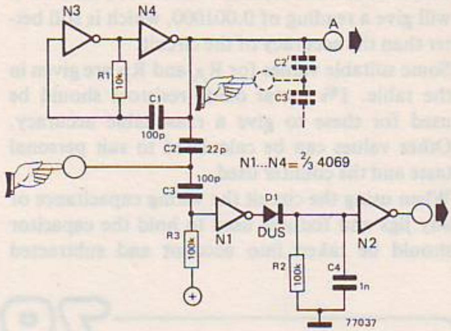
AN elektor PUBLICATION

79 AC touchswitch

There are many designs for touchswitches around these days. However, most of these operated on skin resistance and thus required a double contact that could be bridged by a finger. Single contact operation is possible using a capacitive pickup of mains hum, but this is not very reliable, and will not work at all with battery powered equipment! The design given here overcomes these difficulties and provides a reliable single-point touch switch.

N3 and N4 form a 1 MHz oscillator. When the contact is not touched the signal from the output of N4 is fed via C2 and C3 to the input of N1, which causes the output of N1 to go high and low at a 1 MHz rate. This charges up C4 via D1, holding the input of N2 high which causes the output to remain low.

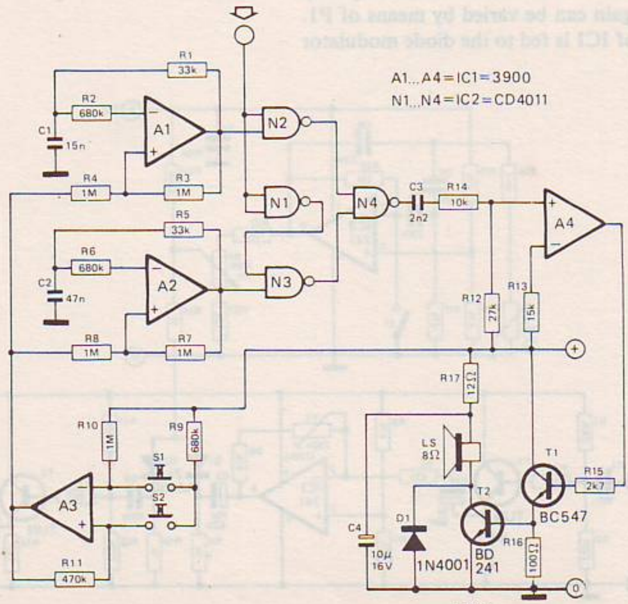
When the contact is touched, body capacitance 'shorts out' the 1 MHz signal. The input of N1 is pulled high by R3 and the output goes low. C4 discharges through R2 and the output of N2 goes high.



One oscillator will provide a 1 MHz signal for several touch switches, which may be connected to point A.

A.M. Bosschaert

80 Audible logic probe



This logic probe provides an audible rather than a visual indication of logic state by producing a high-frequency audio tone for a logic '1' state and a low-frequency tone for a logic '0' state.

The logic input signal is fed to N1 and N2. If the input is high then N2 will pass the high frequency signal from the oscillator built around A1. If the input is low then N2 will block, but the output of N1 will be high so N3 will pass the low frequency signal from the oscillator built around A2.

Depending on the input state one or other of these signals is fed through N4 to the input of a differentiator built around A4. This produces a train of short pulses from the squarewave input signal and these are fed to an audio amplifier comprising T1 and T2. The use of short pulses ensures a high peak audio output while keeping the average current consumption low.

To avoid the annoying 'bleeping' of the circuit when measurements are not being taken both oscillators may be switched on and off by a flipflop constructed around A3, which is controlled by two push buttons S1 and S2.

If the circuit is to be used exclusively with TTL circuits then N1 to N4 should be a 7400 IC and the supply voltage should be + 5 V, which can be derived from the circuit under test. If it is to be used with CMOS ICs then N1 to N4 should be a 4011 IC, and the circuit will operate over supply voltages of 5 to 10 V at a current consumption of between 4 and 10 mA.

H. Käser

81 Level shifter

It is often necessary, particularly when experimenting with circuits, to make connection between the output of one circuit and the input of another which is at a different DC level. If the signals involved in the circuit are AC signals this is no problem, a capacitor can be used to isolate the DC levels while allowing AC signals to pass. However, when dealing with DC or very low frequency AC signals the solution is not so easy, and it is in these cases that this little gimmick will prove useful.

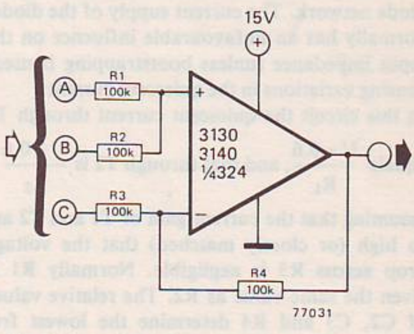
The circuit consists simply of an op-amp connected as a voltage follower whose quiescent output voltage can be set to any desired level within the output range of the op-amp.

Input A is connected to the output of the circuit in question while the output is connected to the input of the circuit which it is feeding. Input C is grounded, while input B is connected to a DC voltage equal to the difference between the input voltage of the second circuit and the output voltage of the first.

It can easily be proved that this works! Firstly, voltages appearing at the non-inverting input of the op-amp are amplified by a factor

$$\frac{R_3 + R_4}{R_3} = 2.$$

Secondly, suppose the output voltage of the first circuit is V_A and the input voltage of the second circuit is V_1. The voltage V_B applied to input B



is thus $V_1 - V_A$. The voltage appearing at the junction of R1 and R2 is thus $V_A + \frac{V_B - V_A}{2}$.

The voltage appearing at the op-amp output is twice this, i.e. $V_A + V_B$

But since $V_B = V_1 - V_A$ this equals V_1, the input voltage of the second circuit. Obviously, if V_1 is less than V_A then V_B will be a negative voltage.

Despite the difference in input and output levels the circuit functions as a voltage follower in that any change in the voltage at input A will produce the same voltage change at the output.

The circuit can also be used as an inverter. In this case the signal is fed to input C, B is grounded and A is fed with a DC reference voltage. To see what voltage must be applied to A it is simplest to treat the circuit as a unity gain differential amplifier. The output voltage V_O is equal to the difference between the voltages at the non-inverting and inverting inputs i.e. $V_O = V_A - V_C$ so $V_A = V_O + V_C$, i.e. input A must be fed with a voltage that is the sum of the voltage at C and the required output voltage. Any change in the input voltage at C will produce the same change at the output, but of opposite polarity.

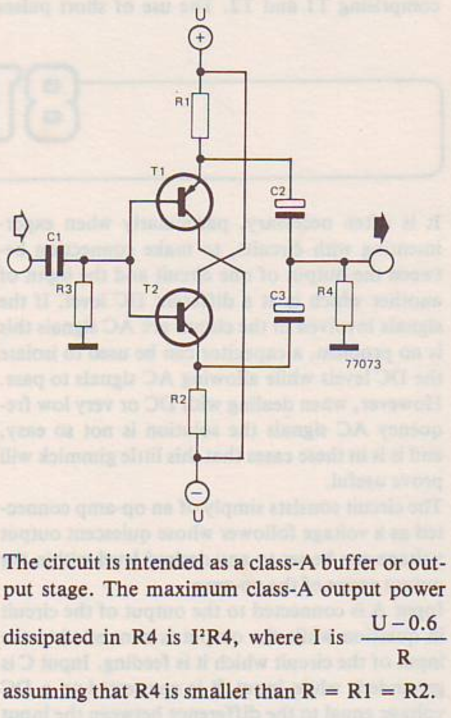
Two points must be noted when using this circuit. Firstly, care must be taken not to exceed the common-mode input rating of the op-amp used, especially with a single-ended (asymmetric) supply. Secondly, the values of R1 to R4 should be at least ten times the output resistance of the circuit feeding the level shifter to avoid excessive loading of the output.

82 Complementary emitter follower

This circuit presents an interesting alternative method of constructing a low-distortion buffer or output stage for use at low output powers. The quiescent current flowing through T1 and T2 is determined solely by the value of U and of R1 and R2 respectively. This contrasts with conventional circuits where the bases of T1 and T2 are connected to one another by means of a diode network. The current supply of the diodes normally has an unfavourable influence on the input impedance (unless bootstrapping is used) causing variations in the quiescent current.

In this circuit the quiescent current through T1 equals $\frac{U-0.6}{R_1}$, and that through T2 is $\frac{U-0.6}{R_2}$, assuming that the current gain of T1 and T2 are so high (or closely matched) that the voltage drop across R3 is negligible. Normally R1 is given the same value as R2. The relative values of C2, C3 and R4 determine the lowest frequency at which the circuit will function.

If T1 and T2 have the same current gain and R1 equals R2, then no DC voltage is produced across R3, and C1 may be omitted. If the circuit is fed from an op-amp then both C1 and R3 may be omitted.



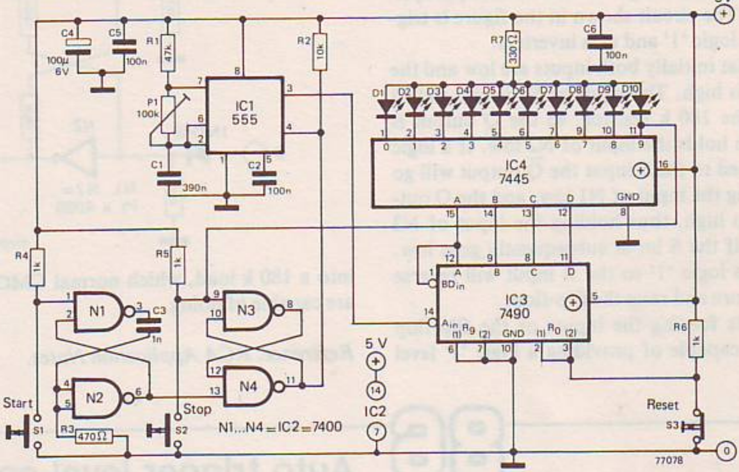
The circuit is intended as a class-A buffer or output stage. The maximum class-A output power dissipated in R4 is I^2R4, where I is $\frac{U-0.6}{R}$ assuming that R4 is smaller than R = R1 = R2.

83 Reaction speed tester

This circuit represents a simple design for one of the most popular types of electronic game, namely a reaction tester.

As soon as the 'start' button is pressed, IC1 feeds a train of pulses to the counter IC3, causing LEDs 1...10 to light up one after another. The sooner the 'stop' button is pressed, the smaller the number of LEDs which light up; the last LED to light up burns continuously. If the oscillator which generates the clock pulses is set

so that a pulse is produced say, every 20 ms, then the reaction time of the players can be calculated quite simply by observing which LED remains lit.
A new game can be started after pressing the reset button.
With the component values given in the diagram the circuit consumes 120 mA with a 5 V stabilised supply. The oscillator frequency may be adjusted by means of P1 between 10 and 80 Hz.
If desired, an additional LED with a 220 Ω series resistor can be included between the output of N3 and positive supply. This will light up as soon as the opponent presses the 'start' button.

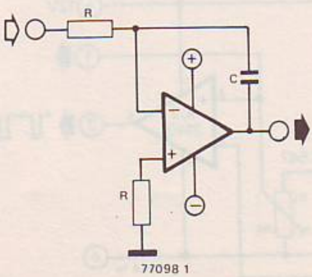


84 Non-inverting integrator

A drawback of conventional integrator circuits (figure a) is that the R-C junction is at virtual earth; this means that C appears as a capacitive load across the op-amp output, a fact that may adversely affect the stability and slew rate of the op-amp. Since the non-inverting character of an integrator is of minor importance in many applications the circuit shown in figure b offers a viable alternative to conventional arrangements.

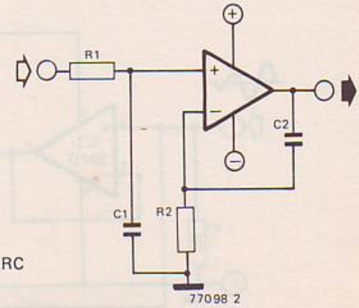
This integrator, unlike that in figure a, is non-inverting. The time constants R_1C_1 and R_2C_2 should be equal.
If both R_1 and C_1, and R_2 and C_2 are transposed then the result is a non-inverting differentiator.
For correct offset-compensation R_1 and R_2 should have the same value.

a



$R_1C_1 = R_2C_2 = RC$

b

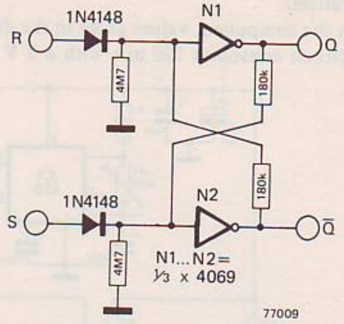


85 Positive-triggered set-reset flip-flop

The standard set-reset flip-flop circuit consists of two cross-coupled NAND gates and is set and reset by applying a logic '0' level to the appropriate input. The circuit shown in the figure is triggered by a logic '1' and uses inverters.

Assume that initially both inputs are low and the $\overline{Q}$ output is high. The input of N1 is also pulled high via the 180 k resistor, so the Q output is low, which holds the input of N2 low. If a logic '1' is applied to the S input the $\overline{Q}$ output will go low, pulling the input of N1 low, and the Q output will go high, thus holding the input of N2 high even if the S input subsequently goes low. Applying a logic '1' to the R input will reverse the procedure and reset the flip-flop.

The circuits feeding the inputs of the flip-flop should be capable of providing a logic '1' level into a 180 k load, which normal CMOS circuits are capable of doing.



Reference: RCA Application Notes.

86 Auto trigger level control

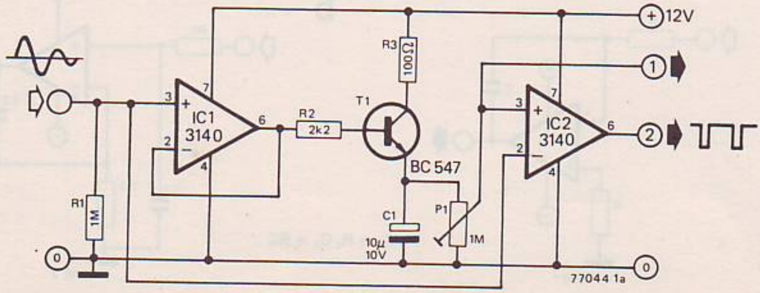
Oscilloscopes, frequency counters and other instruments triggered by AC signals almost invariably have a manual trigger level control, to adjust the point on the waveform at which triggering occurs. When making measurements where the signal level varies, for example at different places in a circuit, it is tedious to have to make frequent adjustments to this control.

The circuit described here provides a trigger signal at a fixed percentage of the peak input level, irrespective of what that level is, so the frustration of having the trace disappear from an oscilloscope when the signal level falls below the trigger level is avoided.

The circuit consists basically of a peak rectifier that provides one input of a comparator with a DC voltage equal to a fixed percentage of the peak signal level. The other input of the comparator is fed with the signal. When the signal level exceeds the DC reference level the comparator output will go low. When it falls below the reference level the comparator output will go high.

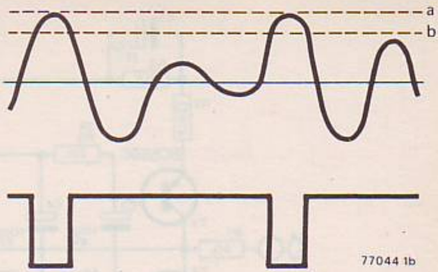
The peak rectifier consists of IC1 and T1. On positive half cycles of the signal waveform the output of IC1 will swing positive until T1 starts

1a



to conduct, after which IC1/T1 will act as a voltage follower, charging up C1 to the peak value of the signal.
A portion of this voltage is taken from the slider of P1 and applied to the non-inverting input of IC2, which functions as a comparator. The AC signal is fed to the inverting input. When the signal level exceeds the reference voltage the comparator output will go low; when the signal level falls below the reference level the comparator output will go high (see figure 1b).
P1 may be used to set the trigger level to any desired percentage of the signal level. The DC level at the slider of P1 may also be fed to the comparator input of an existing trigger level circuit.

1b

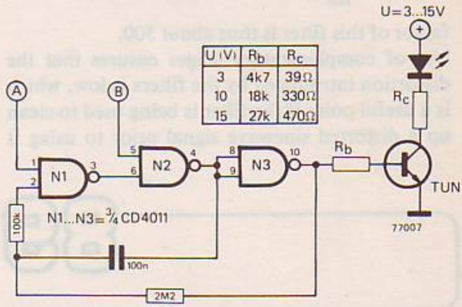


In this case this circuit should have a high input impedance to avoid discharging C1. Alternatively the output from P1 can be buffered by an op-amp connected as a voltage follower.

87 LED logic flasher

The condition of the LED is determined by the logic states of the two inputs A and B. If A is low and B is high then the LED will be lit continuously. If B is low then the LED will be extinguished, irrespective of the state of A. If A and B are both high then the astable multivibrator comprising N1, N2 and N3 will start to oscillate and the LED will flash at about 3.5 Hz. Component values are given for supply voltages of 3, 10 and 15 V. At the maximum supply voltage of 15 V the current consumption is less than 25 mA.

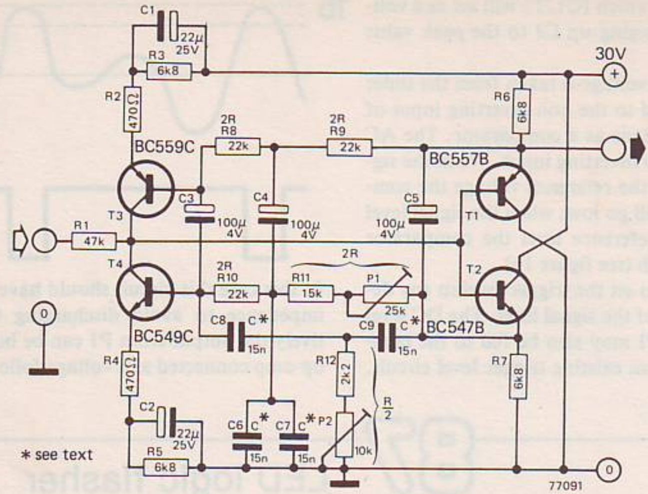
Source: RCA CMOS Application and design ideas.



U (V)	R_b	R_c
3	4k7	39Ω
10	18k	330Ω
15	27k	470Ω

88 Complementary twin-T selective filter

This filter will pass signals at its centre frequency while attenuating signals at all other frequencies. The input signal is fed via R1 to the bases of the complementary emitter follower T1/T2. Feedback is taken from the emitters of T1 and T2, through the twin-T network to the inputs of the complementary amplifier T3/T4. At frequencies removed from the centre frequency of the twin-T network, the feedback signals will pass through the twin-T unattenuated. These signals will be amplified by T3 and T4 and will appear at T3 and T4 collectors in antiphase with the input signal. The input signal, and hence the output signal from the emitters of T1 and T2, will be greatly attenuated.
At the centre frequency of the twin-T the feedback signal will be greatly attenuated, so little antiphase signal will appear at the collectors of T3 and T4 and the input signal will pass unattenuated. The output may be taken from the emitter of either T1 or T2.
The quality factor of the filter is approximately $\frac{A}{4}$, where A is the gain of the T3/T4 stage,



which is $\frac{2R1}{R2}$ (R2 and R4 are equal). The Q-factor of this filter is thus about 500.

Use of complementary stages ensures that the distortion introduced by the filters is low, which is a useful point if the filter is being used to clean up a distorted sinewave signal prior to using it for a distortion measurement.

The centre frequency of the filter is given by $f = \frac{1}{2\pi RC}$, and with the component values shown the centre frequency is about 1 kHz. P1 and P2 can be used to fine tune the filter for maximum output at the required frequency.

89 Negative supply from positive supply

It is sometimes necessary to provide a negative supply voltage in a circuit that otherwise uses all positive supply voltages, for example to provide a symmetrical supply for an op-amp in a circuit that is otherwise all logic ICs. Providing such a supply can be a problem, especially in battery operated equipment.

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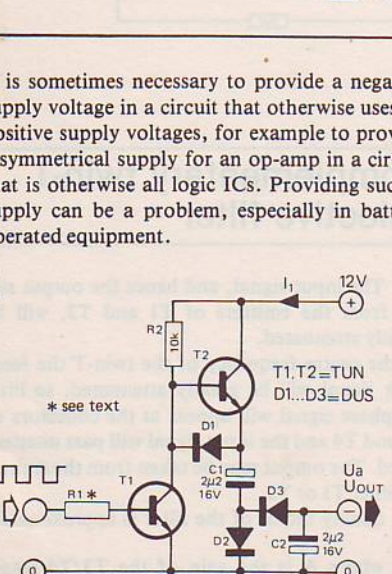

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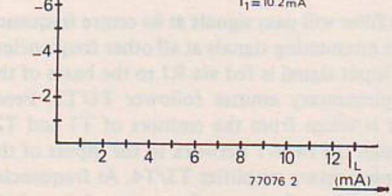
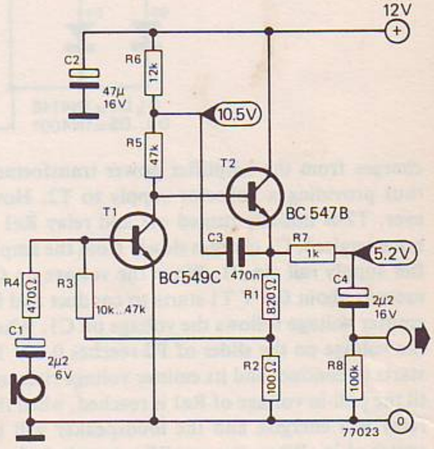


In the circuit shown here T1 is turned on and off by a squarewave signal of 50% dutycycle at approximately 10 kHz. In logic circuits it is quite conceivable that such a signal may already be

available as clock pulses. Otherwise an oscillator using two NAND gates may be constructed to provide it.
When T1 is turned off, T2 is turned on and C1 charges through T2 and D2 to about 11 V. When T1 turns on, T2 turns off and the positive end of C1 is pulled down to about +0.8 V via D1. The negative end of C1 is now about 10.2 V negative so C1 discharges through D3 into C2, thus charging it. If no current is drawn from C2 it will eventually charge to around −10 V. Of course, if a significant amount of current is drawn, the voltage across C2 will drop as shown in the graph and a 10 kHz ripple will appear on the output.

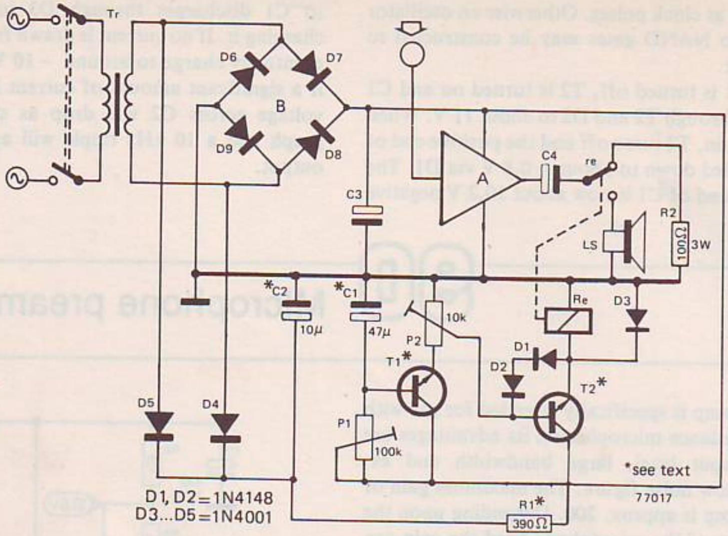
90 Microphone preamp

This preamp is specifically intended for use with low impedance microphones; its advantages are high output level, large bandwidth and extremely low noise figure. The maximum gain of the preamp is approx. 200. Depending upon the sensitivity of the microphone used the gain can be adjusted by altering the value of resistor R3 (for which a suitable typical value is around 22 k).
The low noise figure (virtually undetectable in the lab) is obtained by precise impedance-matching of the input. Optimal results are therefore obtained only with microphones of 500 to 600 Ω impedance. For 200 Ω microphones R4 should be reduced in value to 220 Ω, and C1 increased to 4µ7.
Sound 'perfectionists' may wish to use metal film resistors for R3...R6 and parallel-connected MKM capacitors in place of an electrolytic capacitor for C1. Further details: with an input signal of 3.5 mV_{pp} and maximum gain, an output signal of 800 mV_{pp} was obtained. The maximum output level is approx. 10 V_{pp} for an input of 50 mV_{pp}. The frequency response was flat within 3 dB from 50 Hz...100 kHz.



91 Loudspeaker delay circuit

Many owners of hi-fi equipment are plagued by switch-on and switch-off thumps which, while they rarely cause actual damage to the loudspeakers, are nonetheless very annoying. The simple solution is to switch the loudspeakers into circuit after the amplifier has been switched on and has settled down, and to switch them out of circuit before the amplifier is switched off.
This can be done manually, but there is always the chance that the user will forget, so an automatic switch seems the best answer. This can be achieved in a very simple manner. The circuit consists basically of a delay circuit and a relay that is energised to switch in the speakers a few seconds after the amplifier is switched on.
The DC supply to the relay has a very short time constant (much shorter than that of the amplifier power supply) so that when the amplifier is switched off the relay immediately drops out, disconnecting the speakers before a switch-off thump can occur.
The circuit functions as follows: at switch-on C2



charges from the amplifier power transformer, thus providing a collector supply to T2. However, T2 is initially turned off and relay Re1 is not energised. C1 charges slowly from the amplifier supply rail via P1. When the voltage on C1 exceeds about 0.6 V T1 starts to conduct and its emitter voltage follows the voltage on C1. When the voltage on the slider of P2 reaches 0.6 V T2 starts to conduct and its emitter voltage rises until the pull-in voltage of Re1 is reached, when the relay will energise and the loudspeaker will be switched in. When the amplifier is switched off the voltage on C2 will decay rapidly and Re1 will drop out, disconnecting the loudspeaker before the amplifier supply voltage has decayed and thus eliminating the switch-off thump.

The switch-on delay can be set by means of P1. P2 can be used to set the final voltage across Re1 to just above its pull-in voltage. This means that the relay voltage is not critical and any relay with a pull-in voltage less than the amplifier supply voltage may be used.

If the amplifier output is capacitor coupled then a 100 Ω 3 W resistor should be connected between the normally closed contact of the relay and ground to charge the output capacitor before the loudspeaker is connected. One set of relay contacts is, of course, required for each channel of the amplifier.

The ratings of transistors T1 and T2 should be chosen to suit the amplifier supply voltage. Medium power transistors such as BC142's should be adequate in most cases.

J. Rongen

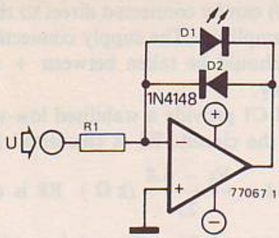
92 Voltage-controlled LED brightness

It is sometimes necessary to make the brightness of an LED vary in proportion to the magnitude of a DC control voltage, which in some cases may be less than the forward voltage drop of the LED.

The brightness of an LED is proportional to the current flowing through it, so the circuit required is a voltage/current converter which will provide a current through the LED independent of the forward voltage drop. This requirement is met by the well-known op-amp active rectifier circuit.

If a positive voltage is applied to the input in figure (a) then the output voltage will swing negative until the LED conducts. As the inverting input of the op-amp is a virtual earth point the cur-

1



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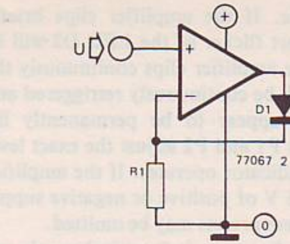


rent flowing through R1 and hence through the LED is $\frac{U}{R1}$. In the absence of an input the op-amp offset could cause the output to swing positive and exceed the reverse breakdown voltage of the LED. For this reason D2 is included to limit the maximum positive excursion to +0.6 V. Negative voltages may be used by reversing D1 and D2.

In figure (a) the input voltage must supply all the current taken by the LED, but figure (b) shows a circuit with a high input resistance that takes virtually no current from the input voltage.

The positive input voltage is applied to the non-inverting input of the op-amp, and the output voltage of the op-amp swings positive until the voltage on the inverting input is the same. A current $\frac{U}{R1}$ thus flows through R1 and since it is provided by the op-amp output it also flows through D1.

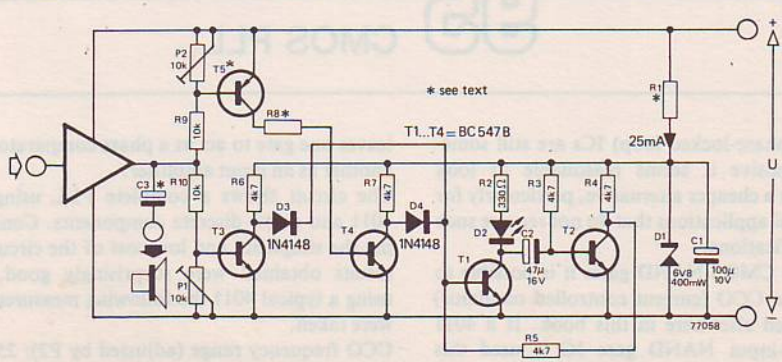
The value of R1 is simply $\frac{U_{max}}{I_{max}}$, where these are respectively the maximum input voltage and maximum LED current required. Any op-amp capable of supplying the required current may be used.

C. Chapman

93 Clipping indicator

This circuit is designed to indicate the onset of clipping in an amplifier on both positive and negative peaks. Clipping occurs when the output of an amplifier swings up or down to its positive or negative limit, which is usually just below the supply voltage, so the circuit is designed to detect this.

When the amplifier output clips positive T5 will turn off, which will cause T4 to turn off, triggering the monostable T1/T2, which will cause D2 to light for about 200 ms. If the amplifier clips negative then T3 will turn off, again triggering



the monostable. If the amplifier clips briefly then just a short flicker of the LED D2 will be seen, but if the amplifier clips continuously the monostable will be continuously retriggered and the LED will appear to be permanently lit. Potentiometers P1 and P2 adjust the exact level at which the indicator operates. If the amplifier clips within 0.6 V of positive or negative supply voltage then these presets may be omitted.

If the amplifier has a single (positive) supply rail and capacitor-coupled output then the input of the circuit should be connected to the 'hot' end of this capacitor, i.e. to the top of C3 in the diagram (C3 is the output capacitor of the amplifier). The supply rails to the circuit should be connected between + supply and 0 V.

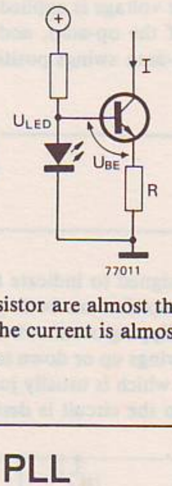
If the amplifier has a symmetrical supply and direct-coupled output then the input (junction of R9 and R10) can be connected direct to the output of the amplifier. The supply connections to the circuit should be taken between + supply and − supply.

R1, D1 and C1 provide a stabilised low-voltage supply for the circuit. R1 is calculated by the equation: $R1 = \frac{Vs - 6.8}{25}$ (kΩ). R8 is chosen so that about 1 mA flows through it and is given by $R8 = \frac{Vs}{1}$ (kΩ). Where Vs is the total supply voltage between + and 0 or + and − rails as appropriate. For values of Vs up to 45 V T5 may be a BC157B or BC557B and for voltages up to 65 V a BC556N.

94 Using LEDs as reference diodes

Depending on type and the current flowing, the forward voltage drop of an LED may lie between 1.4 and 2 volts. The temperature coefficient of this voltage is about −1.5 mV/°C.

As this is virtually the same as the temperature coefficient of the base-emitter voltage of a silicon transistor, it is very easy to construct a constant current source with almost zero temperature coefficient, as shown in the accompanying circuit.



The current is approximately $\frac{U_{LED} - U_{BE}}{R}$.

Since the temperature coefficients of the LED and the transistor are almost the same they cancel out and the current is almost independent of temperature.

95 CMOS PLL

As PLL (phase-locked loop) ICs are still somewhat expensive it seems reasonable to look around for a cheaper alternative, particularly for non-critical applications that do not require such high specifications.

Using two CMOS NAND gates it is possible to construct a CCO (current controlled oscillator) as described elsewhere in this book. If a 4011 quad two-input NAND gate IC is used this leaves one gate to act as a phase comparator and another as an input amplifier.

The circuit shows a complete PLL using one 4011 and a few discrete components. Considering the simplicity and low cost of the circuit the results obtained were surprisingly good, and using a typical 4011 the following measurements were taken.

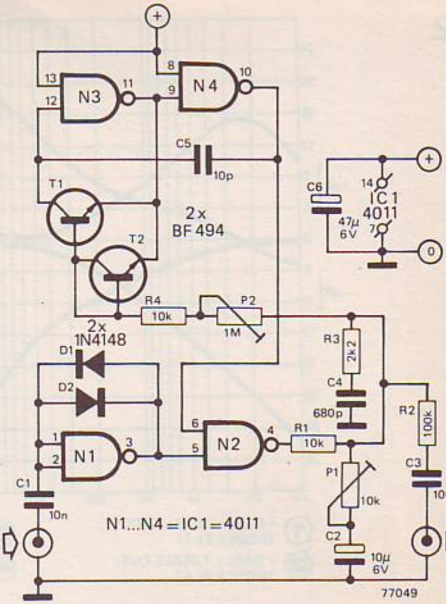
CCO frequency range (adjusted by P2): 25 kHz

– 800 kHz. Hold range: 20% of CCO free-running frequency. Output level: 45 mV measured at f_{in} = 500 kHz, deviation = ± 30 kHz, modulation frequency = 1 kHz. AM suppression for 30% AM: better than 40 dB. Minimum input level: less than 2 mV from 50 Ω source.

These measurements were taken at a supply voltage of 6 V, when the current consumption was 600 μA.

Since different IC manufacturers use different processes and different chip geometries it might be expected that results would vary when using different types of IC. The best results were obtained using ICs in which the gates had a steep transfer characteristic (better approach to an ideal switch) and lowest crosstalk between gates. In our experience, the Solid State Scientific SCL 4011 is a good example of this type of chip.

The 4011 PLL is particularly suitable for narrow-band FM demodulation, and in fact proved superior, in terms of s/n ratio and impulse noise rejection, to several monolithic PLL ICs.



96 Guitar preamp

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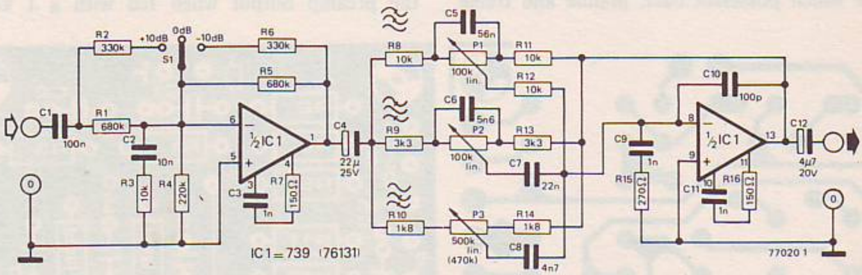

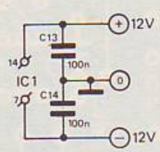

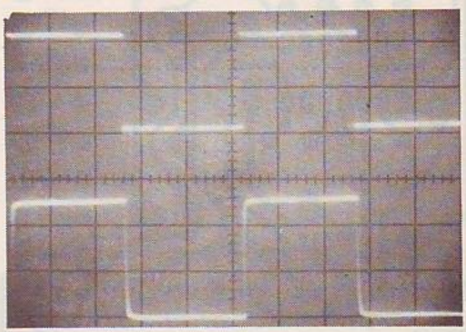

Using a μ A739 dual, low-noise op-amp it is a simple matter to construct a versatile preamplifier for a magnetic guitar pick-up. The input stage consists of an amplifier with a flat frequency response having switchable gains of – 10 dB, 0 dB and + 10 dB, so that it may be

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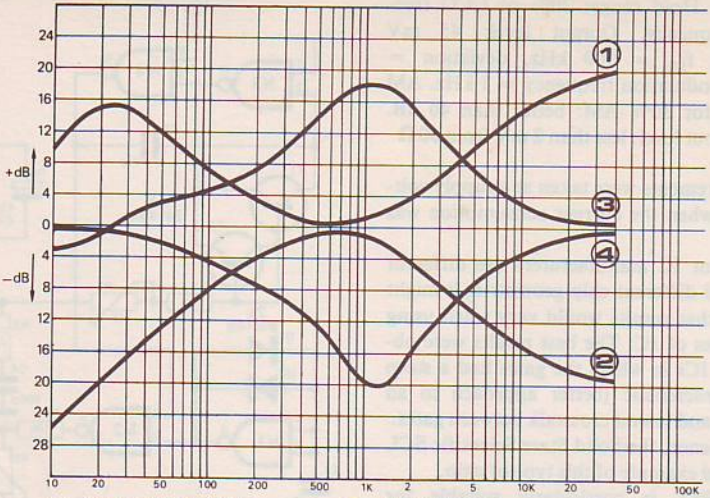

① = BASS + TREBLE BOOST, MIDDLE FLAT
② = BASS + TREBLE CUT, MIDDLE FLAT
③ = MIDDLE BOOST, BASS + TREBLE FLAT
④ = MIDDLE CUT, BASS + TREBLE FLAT

used with pick-ups having a variety of output levels. The switchable gain also makes possible feedback when the guitar is brought close to the loudspeakers. This effect, much favoured by guitarists, can be achieved if the guitar amplifier power is around 20 watts or greater.

The input stage is followed by a tone control stage which possesses bass, middle and treble controls. As the frequency response of many guitar pick-ups is far from flat these controls can be used to compensate for any peaks or dips. The response of the tone control networks for different settings of the control pots is shown in the accompanying graph.

The oscillogram, the lower trace of which shows the preamp output when fed with a 1 kHz

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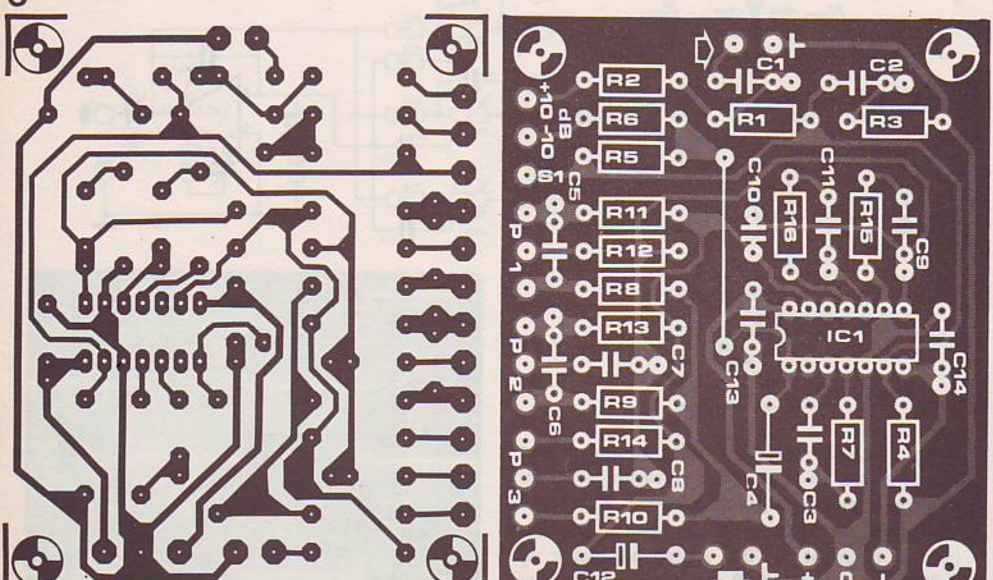

squarewave (upper trace) illustrates that the h.f. response of the preamp is fairly good. Indeed, the performance of the preamp is so good that, as well as its intended use, it may also be used in hi-fi systems.

97 Osculometer

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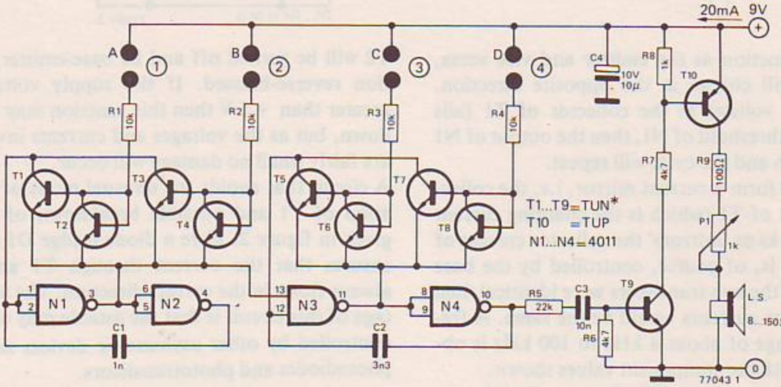

'Lie Detector' machines, which measure skin resistance, can provide great amusement at parties, especially if two participants each hold one electrode and indulge in some form of physical contact (e.g. kissing). The meter can then be calibrated in degrees of passion.

A variation on this theme is a circuit that produces an audible output rather than a meter indication, which is even more amusing. The circuit consists of two current controlled oscillators (described elsewhere in this book). The output of oscillator N1/N2 gates oscillator N3/N4 which produces some interesting effects. The output of N4 is used to drive an audio amplifier comprising T9 and T10.

The circuit has provision for eight electrodes for up to four pairs of participants. As the resistance

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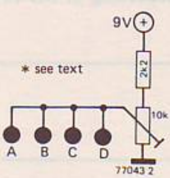

between a pair of electrodes decreases then the frequency of the corresponding oscillator will rise, so the more ardent the embrace the higher the oscillator frequency. The gating effect between the two oscillators produces some unusual sounds.

For safety reasons the circuit should be battery powered by a 9 V transistor 'power pack', such as a PP3, PP6, PP9 etc.

98 CMOS CCO

Using two CMOS NAND gates (or inverters) and two transistors it is possible to construct a simple current-controlled oscillator (CCO). The circuit of figure 1 is based on a normal two-inverter astable multivibrator. When the output of N1 is high the output of N2 will be low and C1 will charge through T1 until the threshold voltage of N1 is exceeded, when the output of N1 will go low and the output of N2 high. T1 will now operate in a reverse direction, i.e. the collec-

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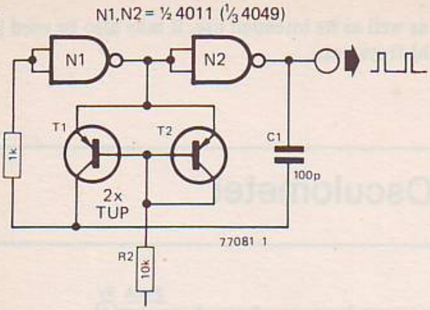

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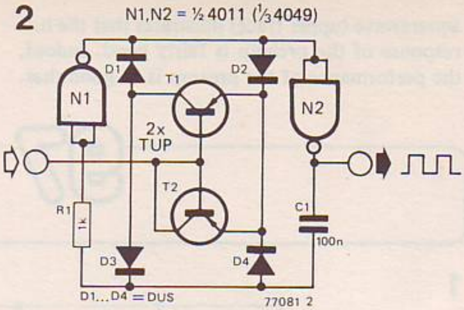

tor will function as the emitter and vice versa, and C1 will charge in the opposite direction. When the voltage at the collector of T1 falls below the threshold of N1, then the output of N1 will go high and the cycle will repeat.

T1 and T2 form a current mirror, i.e. the collector current of T1 (which is the charging current of C1) tracks or 'mirrors' the collector current of T2, which is, of course, controlled by the base current. If the two transistors were identical then the collector currents would be the same. A frequency range of about 4 kHz to 100 kHz is obtainable with the component values shown.

When T1 is conducting in the reverse direction T2 will be turned off and its base-emitter junction reverse-biassed. If the supply voltage is greater than +5 V then this junction may break down, but as the voltages and currents involved are fairly small no damage will occur.

A circuit that avoids the unusual mode of operation of T1 and possible breakdown of T2 is given in figure 2. Here a diode bridge D1 to D4 ensures that the current through T1 and T2 always flows in the correct direction. The advantage of this circuit is that the astable may also be controlled by other asymmetric devices such as photodiodes and phototransistors.

99 LED tuner

This circuit can be used as a tuning indicator, instead of the more common pointer instrument. It gives a three-LED indication of correct tuning: 'off-to-one-side', 'correctly tuned', 'off-to-the-other-side'.

A voltage is derived from the AFC control voltage in the FM receiver and fed to two comparators (IC1 and IC2). The divider chain R2, R3, P1 and R4 produces two reference voltages. If the input voltage is higher than the greater of the two reference voltages T1 will be turned on and LED D1 will light. In the other extreme case, where the input voltage is lower than the lower reference voltage, T2 will be turned on and LED D2 will light. In the in-between range, where the receiver is correctly tuned and the input voltage is somewhere between the two reference voltages, T1 and T2 will both be turned off. In this case the output of N1 will go low, the trigger circuit N2/N3 will switch, the output of N4 will go high and T3 will be turned on – lighting LED D3.

Since the AFC voltage corresponding to 'correctly tuned' varies considerably from one receiver to the next, the values of R2, R3, R4 and P1 are not given in the circuit. It is a simple matter to calculate these values for any particular application. If the total resistance is to be 20...30 k (a reasonable assumption), the voltage midway along R3 should correspond to the AFC voltage for correct tuning. To give two examples:

– assume that the 'correct' AFC voltage is 9.5 V. In this case the voltage across R2 + ½R3 should equal 2.5 V and the voltage across the selected as 4k7 and a 1 k preset is used for R3, the sum of P1 plus R4 should be approximately 20 k with P1 in the mid position. A good choice in this case would be R4 = 18 k and P1 = 4k7 (preset).

– assume that the correct AFC voltage is 5.6 V (as for the CA 3089!). In this case the voltage across the upper half of the divider chain should be approximately 6.5 V; reasonable values are

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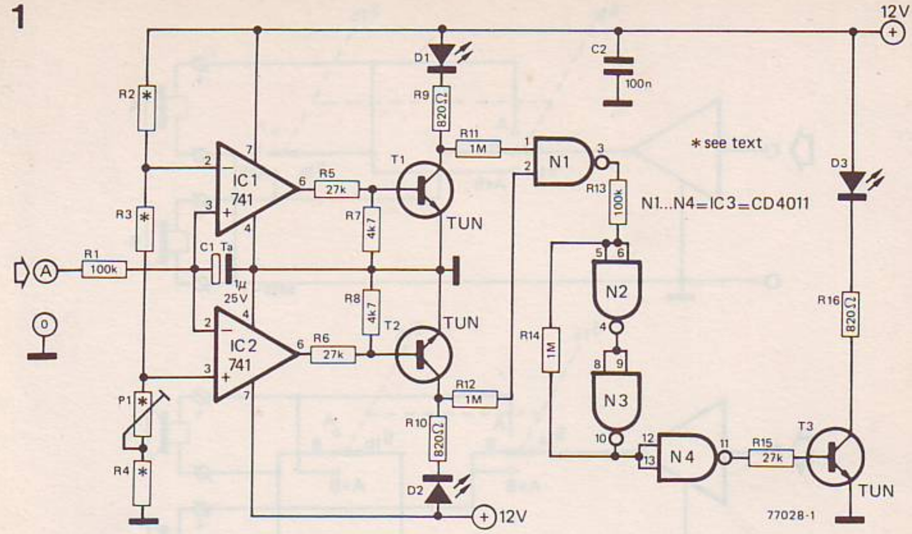

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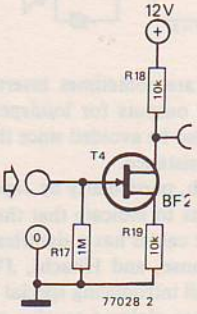

R2 = 12 k and R3 = 2k2. R4 + ½P1 should be approximately 10 k, so R4 can be 8k2 and P1 can be 4k7.

Note that R3 sets the sensitivity of the indicator, whereas P1 is used for correct calibration.

Some FM detectors, notably ratio discriminators, give a 0 V output when correctly tuned. In this case the circuit shown in figure 2 can be added, between the AFC output and the input to the circuit shown in figure 1.

W. Auffermann

100 Loudspeaker connections

Many hi-fi enthusiasts may not realise that significant distortion may be introduced into an audio signal by the connections between the amplifier output and the loudspeakers. In the first place, output current from the amplifier has to travel across several non-soldered metal-to-metal contacts, for example plug and socket connections at the amplifier outputs and the loudspeaker inputs, and loudspeaker switches within the amplifier (of which more later). For minimum distortion these contacts should not only have a very low resistance, but must also have a constant, linear resistance.

Oxidation of the metal surfaces of plugs, sockets and switch contacts can produce a non-linear resistance which varies with the current flowing through it, thus distorting the signal fed to the loudspeakers. DIN loudspeaker plugs and sockets are particularly bad in this respect due to their very small contact area, and should be avoided. Where non-soldered connections must be made the use of screw terminals or robust 4 mm 'banana' plugs and sockets is to be preferred.

The second area which can cause degradation of the audio signal is the connecting cable itself. When a loudspeaker is being driven by an amplifier the loudspeaker cone should move exactly in

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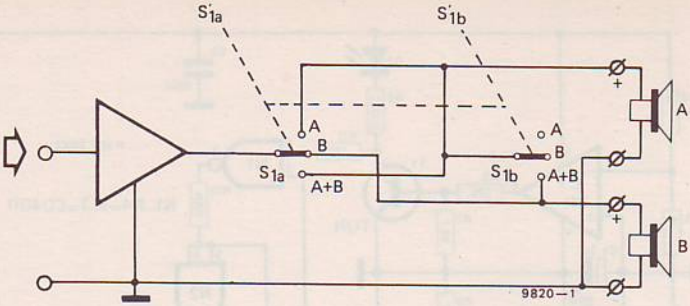

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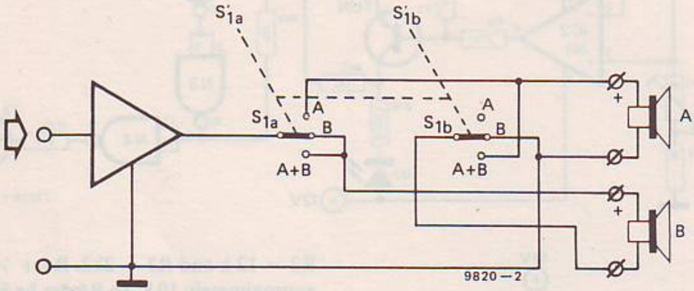


sympathy with variations of the amplifier output voltage. Ideally, if a loudspeaker is fed with, say, a step input, the cone should move quickly to the appropriate position and stop. In practice, of course, this does not happen. A loudspeaker possesses inertia and compliance, so that the cone will tend to oscillate about its final position before settling down. Whilst this 'ringing' is in progress the loudspeaker acts as a generator and tries to pump current back into the amplifier output. If the amplifier output impedance is low (and it generally is) the loudspeaker sees a short-circuit and the cone movement is quickly damped by electromagnetic braking. The 'damping factor' of an amplifier is defined as the ratio of the load impedance to amplifier output impedance. As the output impedance of a modern transistor amplifier is generally a fraction of an ohm, damping factors are typically between 50 and 200 with an 8 ohm load. However, the resistance of the loudspeaker connecting cable appears in series with the amplifier output and must be considered as part of the amplifier output impedance. If the loudspeaker cable is thin its resistance will be high and the damping factor will be considerably reduced. In addition, some of the amplifier's output voltage will be dropped across the cable resistance rather than appearing across the loudspeaker.

Thus the second rule when connecting loudspeakers is to use heavy-duty cable.

Fuses, which are sometimes inserted in series with amplifier outputs for loudspeaker protection, should also be avoided since they can have a significant resistance.

Recent research, particularly by Japanese manufacturers, seems to indicate that the inductance of loudspeaker cables has a significant effect on transient response, and Hitachi, JVC, Pioneer and Sony are all introducing special loudspeaker cables which are claimed to give an improved sound. Whether or not these claims are true is still a matter for conjecture.

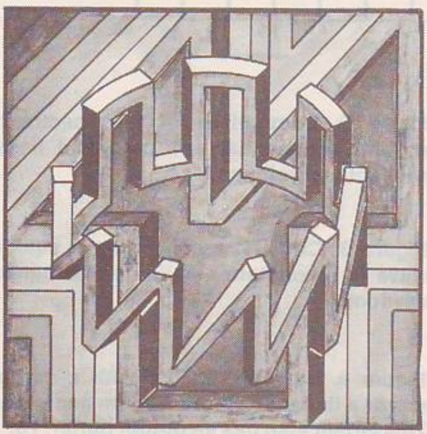
Returning to the subject of loudspeaker switching, figures 1 and 2 show two typical switching arrangements which allow two sets of speakers to be connected to an amplifier, either independently or simultaneously. One channel only is shown and the circuits are identical for the other channel. Although such switching arrangements offer convenience of use, they may not be such a good idea from a sound quality point of view due to the contact resistance of the switches. If loudspeaker switching is employed in an amplifier then the switches used should be rated at several amps to ensure minimum contact resistance. Both the switching arrangements shown in figures 1 and 2 have their advantages and disadvantages. In figure 1 both speakers appear in parallel across the amplifier output in the A + B position. Whilst this does mean that the damping factor is maintained the reduced load impedance

can cause overloading.
In figure 2 the speakers are connected in series in the A + B position. Assuming that both speakers have the same impedance this connection, of course, doubles the load impedance, so there is no risk of overload. However the available output power is halved (since $P = U^2/R$) and the damping factor is reduced to less than unity, since each loudspeaker has the other in series with it as a source impedance.
In conclusion, anyone contemplating the building of an audio amplifier and/or loudspeakers would be well advised to bear in mind all the points raised in this article. To summarise:
1. Connection to the loudspeakers should be made with the minimum number of non-soldered connections (plug and socket connections and switches) in series with the signal path.
2. The cable to the loudspeakers should have as low a resistance as possible. Fuses in series with the loudspeakers, although seemingly desirable from a circuit protection point of view, have a detrimental effect on sound quality and should be avoided.

101 Constant amplitude squarewave to sawtooth

Most electronic organs use squarewaves as the basic signal from which all the organ voices are obtained by filtering, simply because squarewaves are easy to generate and process. However, from a musical point of view, the sawtooth is a much more useful waveform, since it contains both odd and even harmonics of the fundamental frequency, whereas the squarewave contains only the odd harmonics. The main problems involved in generating sawtooth waveforms for organ circuits have been those of cost and reproducibility. However, the circuit described here, for which a patent is pending, suffers from none of these drawbacks and could, in principle, be integrated into a microcircuit.

Most electronic organs use octave dividers, which produce a symmetrical squarewave output. The harmonic content of this waveform is then altered by filtering to give the required organ voices. However, a symmetrical squarewave contains only the odd harmonics of the fundamental frequency, and those voices requiring even harmonics cannot be realistically imitated, since no amount of filtering can add harmonics which are absent. For this reason a sawtooth waveform, which contains both odd and even harmonics, is preferred as the 'raw material' for many organ voices.



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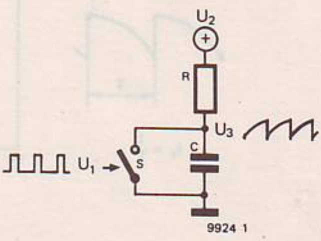

Figure 1. A sawtooth can be generated by charging a capacitor through a resistor and rapidly discharging it at regular intervals.

A sawtooth waveform can be obtained from a squarewave as illustrated in figure 1. A capacitor is allowed to charge, either from a voltage source in series with a resistor or from a constant current source. The positive-going edge of the squarewave is used momentarily to close a (electronic) switch, which rapidly discharges the capacitor. This charging and instantaneous discharging of the capacitor produces the familiar sawtooth waveform.
Figure 2 illustrates the differences in the spectra

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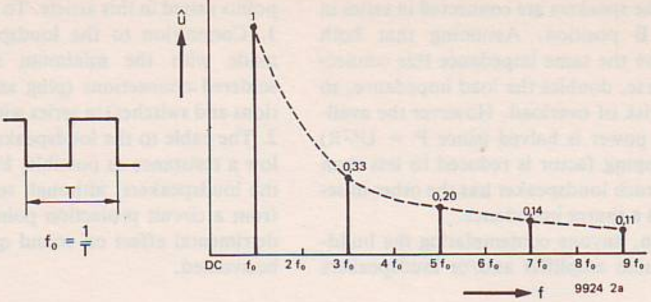
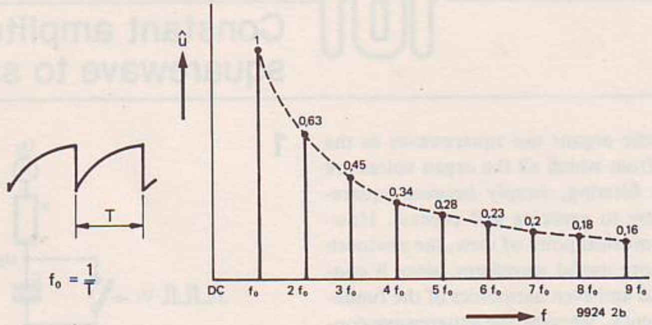
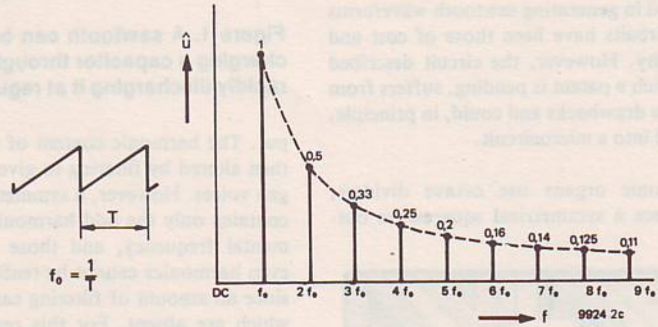


2b



2c



of the square and sawtooth waveforms. If the capacitor is charged from a voltage source then a sawtooth with an exponential curvature results, the spectrum of which is shown in figure 1b. If a constant current source is used then the sawtooth is linear, and has the spectrum shown in figure 1c.
For musical purposes an exponential sawtooth is preferred.
The disadvantage of this simple method is that the amplitude of the sawtooth waveform falls as the frequency on the input squarewave is increased, since the capacitor has less time to

Figure 2. Comparison of the spectra of square, exponential sawtooth and linear sawtooth waveforms.

Figure 3. Circuit for converting a squarewave to a constant amplitude exponential sawtooth.

Figure 4. By replacing R4 with a current mirror a linear sawtooth may be generated. This circuit also has an output buffer, T5.

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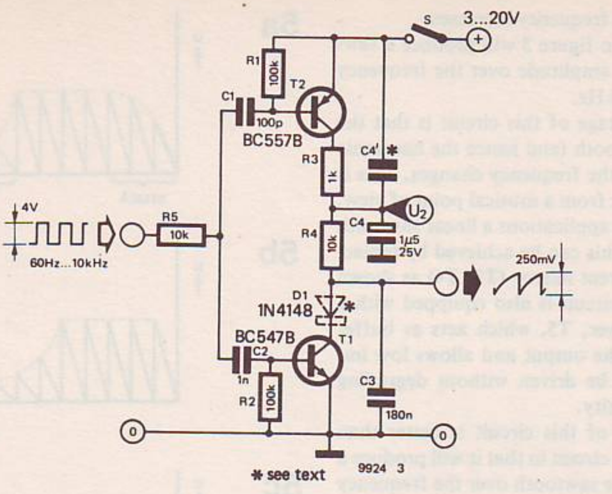

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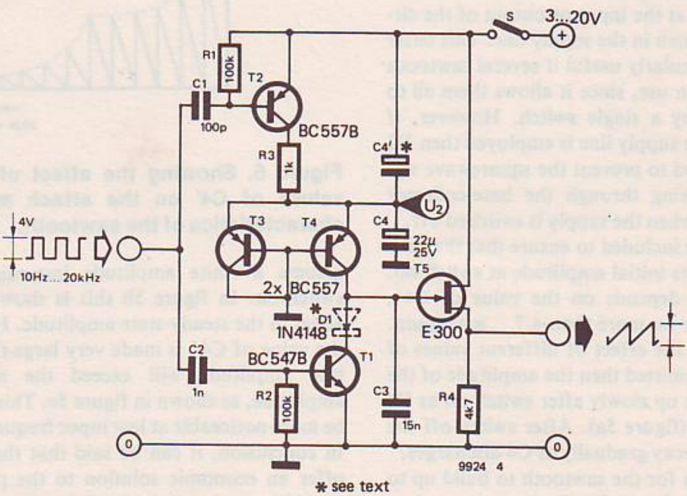


charge. This means that a different capacitor or charging resistor value would have to be used for each note of the organ to maintain equal amplitude over the entire compass of the instrument.

This problem can be overcome by arranging that the voltage of the source from which the capacitor is charged automatically increases as the frequency increases. This increases the charging current, which causes the capacitor to charge more rapidly and thus maintains a constant amplitude. A practical, constant amplitude, squarewave-to-sawtooth converter is shown in figure 3. Capacitor C3 is charged via R4 from the voltage U_2, present on C4. The leading (positive-going) edge of the input squarewave is differentiated by C2 and R2, producing a short pulse which briefly turns on T1 to discharge C3.

The trailing edge of the squarewave is differentiated by C1 and R1, producing a short pulse which briefly turns on T2 and charges C4 via R3. Since T2 is turned on for a fixed time, as the input frequency increases T2 will be turned on for a greater proportion of the total time, so that C4 will charge to a higher voltage. This causes the charging current into C3 to increase, thus compensating for the fact that C3 charges for a

shorter time as the frequency increases.
The circuit given in figure 3 will produce a sawtooth of constant amplitude over the frequency range 60 Hz to 10 kHz.
A slight disadvantage of this circuit is that the shape of the sawtooth (and hence the harmonic content) alters as the frequency changes. This is no great drawback from a musical point of view. However, in some applications a linear sawtooth is preferred, and this can be achieved by replacing R4 with a current mirror (T3, T4) as shown in figure 4. This circuit is also equipped with a FET-source-follower, T5, which acts as buffer between C3 and the output and allows low impedance loads to be driven without degrading the sawtooth linearity.
The performance of this circuit is better than that of the simpler circuit in that it will produce a constant amplitude sawtooth over the frequency range 10 Hz to 20 kHz.
The sawtooth output may be switched on and off by a switch at the input or output of the circuit, or by a switch in the supply line. This latter method is particularly useful if several sawtooth converters are in use, since it allows them all to be controlled by a single switch. However, if switching of the supply line is employed then D1 must be included to prevent the squarewave signal from breaking through the base-collector junction of T1 when the supply is switched off.
C4′ can also be included to ensure that the sawtooth has a finite initial amplitude at switch-on. The amplitude depends on the value of C4′, which should be no more than 4.7 maximum. Figure 5 shows the effect of different values of C4′. If C4′ is omitted then the amplitude of the sawtooth builds up slowly after switch-on as C4 charges via T2 (figure 5a). After switch-off the sawtooth will decay gradually as C4 discharges.
The times taken for the sawtooth to build up to its steady-state value and to decay are dependent on the input frequency, being longest at low frequencies and shortest at high frequencies. This behaviour corresponds quite closely to that of conventional musical instruments.
If C4′ is included then the sawtooth signal will

5a
5b
5c

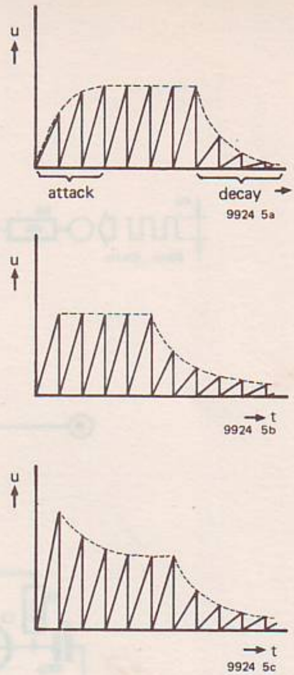


Figure 5. Showing the effect of different values of C4′ on the attack and decay characteristics of the sawtooth.

assume a finite amplitude immediately after switch-on. In figure 5b this is shown as being equal to the steady-state amplitude. However, if the value of C4′ is made very large then the initial amplitude will exceed the steady-state amplitude, as shown in figure 5c. This effect will be more noticeable at low input frequencies.
In conclusion, it can be said that these circuits offer an economic solution to the problem of providing a sawtooth signal in squarewave divider instruments. By taking advantage of the high input resistances offered by MOS technology it should be possible to make the capacitor values sufficiently small to allow total integration of the circuit.

102 Ohmmeter

Using a CA 3140 FET op-amp it is easy to construct a simple, linear-scale ohmmeter. The op-amp is connected in the non-inverting mode, with the non-inverting input fed from a 3.9 V zener. The op-amp output voltage is thus given by

$$\frac{(R_X + R_2)}{R_2} \times 3.9\ \text{V}.$$

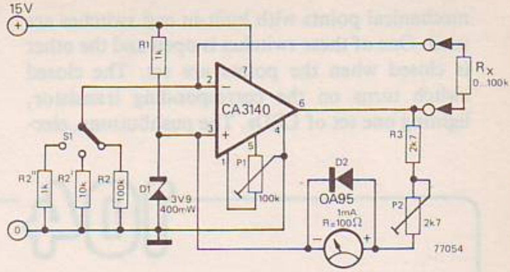
Since one end of the meter is returned to the zener the meter voltage is

$$\frac{R_X}{R_2} \times 3.9 + \frac{R_2}{R_2} \times 3.9 - 3.9 \text{ i.e. } \frac{R_X}{R_2} \times 3.9.$$

Since the zener voltage and R2 are fixed the voltage measured by the meter is proportional to R2. The full-scale deflection of the meter is about 3.9 V, but the exact value will depend on the tolerance of the zener.

Three ranges are provided by using different values of R2. With R2 = 1 k the full-scale reading of 3.9 V is obviously obtained when R_X = 1 k. With R2 equal to 10 k and 100 k full-scale readings are obtained at 10 k and 100 k. The voltmeter is simply a 1 mA meter with a nominal 3k9 series resistor, so the 0 to 1 mA scale can easily be converted to read 0 to 1 k, 0 to 10 k and 0 to 100 k. The germanium diode connected across the meter protects it in the event of an overload.

To calibrate the meter it is first necessary to zero it by nulling the opamp offset voltage. To do this P2 is first set to minimum resistance to make the meter most sensitive, and a wire link is connected across the R_X terminals. P1 is then adjusted to give a zero reading on the meter.



The meter may then be calibrated by connecting a close tolerance resistor of known value (e.g. 100 k 1%) across the R_X terminals and adjusting P2 until the meter reads correctly. To ensure good accuracy on all ranges R2, R2′ and R2′′ should be close tolerance types, 2% or better.

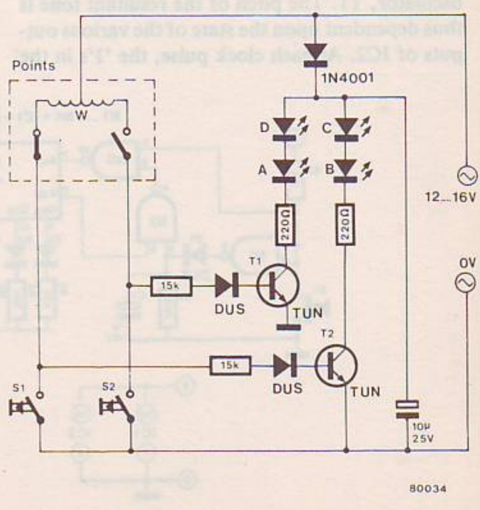
The maximum value which can be used for R2 and hence the maximum value of R_X that can be measured depends on the input resistance of the opamp, since any current flowing into the opamp input will cause errors. However, with the 1.5 T Ω input resistance of the 3140 it should be possible to use values up to 10 M, assuming 10 M close tolerance resistors can be obtained.

103 I see your point!

Whether a model railway is microprocessor-controlled or hand-operated, a visual display of the 'system status' is always worth while. If nothing else, it makes for an impressive control panel. For some functions, it is even essential to have a clear overview – unless, of course, your main aim is to realistically imitate crashes and derailments.

The points, in particular, are extremely important. As many model railway enthusiasts will have discovered, it is not at all easy to see what position the points are in from a distance. Even mechanical 'point position indicators' are not always particularly clear.

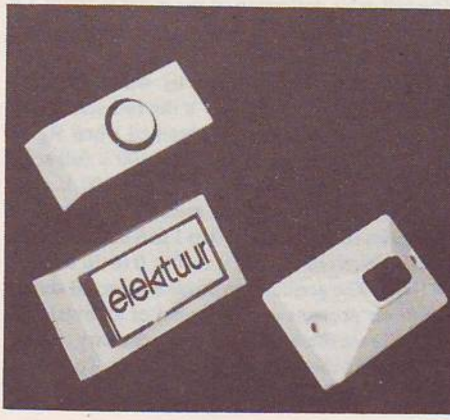
The indicator described here provides an unambiguous display on the main control panel. Different coloured LEDs can be used to provide a clear indication at a single glance.



The circuit could hardly be simpler. Electromechanical points with built-in end switches are used. One of these switches is open and the other is closed when the points are set. The closed switch turns on the corresponding transistor, lighting one set of LEDs. The pushbuttons, electronics and one LED out of each pair can be mounted in the control panel; the other LED in each pair can be mounted alongside the tracks near the corresponding set of points, to give an on-the-spot indication.

104 Musical doorbell

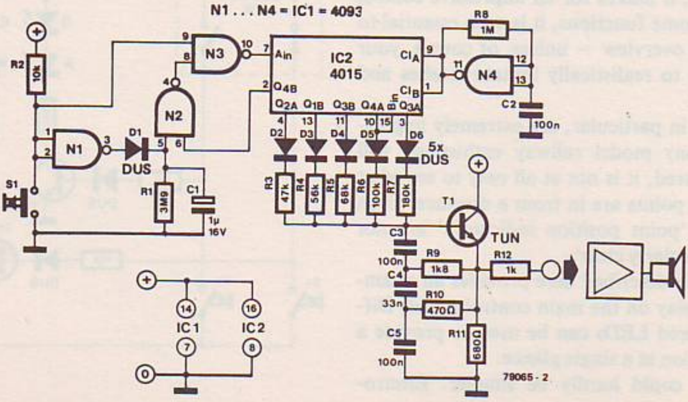
While on the subject of doorbells any kind of alternative gimmick is always worth considering. With the circuit described here, a pleasing effect is obtained. After even briefly pressing the bell-push, a short tune will be played. Holding the button down (or pressing it repeatedly in rapid succession) has two effects: a different melody is obtained, and it lasts longer. The circuit operates as follows.



By pressing pushbutton switch S1, the inputs of N1 and one of the inputs of N3 are taken low, with the result that pin 7 of IC2 (data input A) is taken high. IC2 is a four-bit static shift register, so that upon each successive clock pulse (provided by the clock generator, N4), this logic '1' is transferred to successive outputs. The clock frequency is approximately 5 Hz. The number of '1's clocked through the shift register will be directly proportional to the length of time that S1 is held down.

Each time that one of the outputs of IC2 goes high, a current is supplied via the corresponding resistor to the base of the current controlled oscillator, T1. The pitch of the resultant tone is thus dependent upon the state of the various outputs of IC2. At each clock pulse, the '1's in the shift register move up one place, causing a change in pitch; if the pushbutton is depressed at that time, a new '1' will also be entered. One of the outputs (Q4B) is fed back via N2 and N3, so that the '1's in the register will keep going round the loop.

After the pushbutton is released, the circuit will keep running until C1 is discharged (through R1); if the button is pressed repeatedly, the ca-

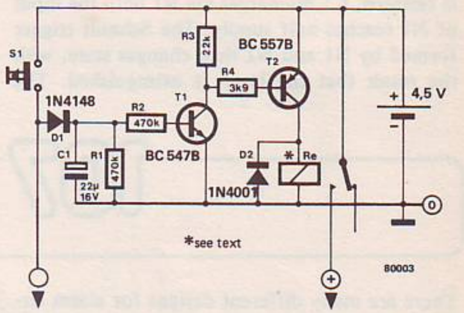


pacitor will remain charged and so the bell will 'run' continuously. The only difference between pressing repeatedly and holding the button down is therefore that a different succession of '1's will be entered, giving a different tune.
With this doorbell, it is necessary to add an output buffer amplifier.
The supply requirements are not critical (5... 15 V, 10 mA).

Lucas Witkam

105 Battery saver

With many electronic games, such as heads-or-tails, roulette, or any of the versions of electronic dice, a considerable saving in battery life can be obtained by ensuring that the circuit, or at least the current-guzzling displays, are switched off after each throw or turn. Naturally enough, it would be somewhat tiresome to have to do this by hand, so the following circuit is intended to take care of this chore automatically.
Basically the circuit is a simple timer. Push-button switch S1 is the start button for the die, roulette wheel, etc. When depressed, it causes capacitor C1 to charge up rapidly via D1. Transistor T1 is turned on, so that, via T2, the relay is pulled in, thereby providing the circuit of the game with supply voltage.
When the switch is released, initially nothing will happen. C1 discharges via R1, R2 and the base-emitter of T1, however it takes several secondes until it has discharged sufficiently to turn of T1. When it does so, however, the relay drops out, cutting out the power supply to the die, etc.



With the component values shown in the circuit diagram, a delay of roughly 3 seconds is provided in which to read off the display. If that interval is too short (or too long), it can be modified as desired by choosing different values for C1 and/or R1/R2.

W. Jitschin

106 Short-interval light switch

Even in today's well-equipped modern houses there are various 'corners' where additional lighting is required. For dark cupboards, meter boxes etc. temporary lighting is usually sufficient, so that making a connection to the mains is hardly worthwhile; a simpler and cheaper solution is to use a battery-powered circuit which will light a lamp for a short period of time. As is apparent from the accompanying circuit diagram, such a circuit is by no means complicated. Using only one CMOS IC, three resistors and one capacitor, the circuit will switch on a lamp for a presettable interval.

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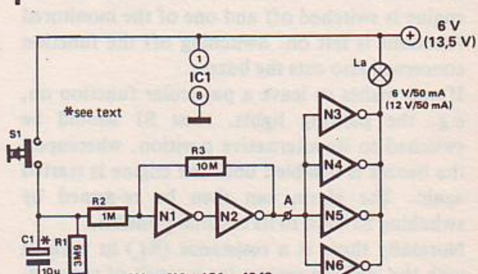

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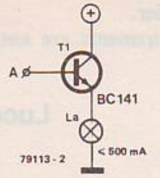
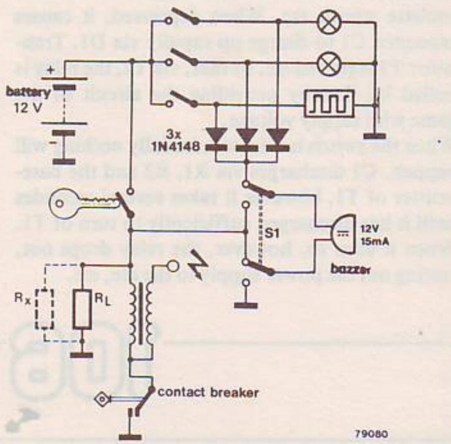


The operation of the circuit is perfectly straightforward: when the button is pushed C1 charges up to the supply voltage. The outputs of the four parallel-connected inverters (N3...N6) are then low, so that the lamp will be lit. When the button is released, C1 discharges via R1 until the input of N1 reaches half supply. The Schmitt trigger formed by N1 and N2 then changes state, with the result that the lamp is extinguished. The positive feedback resistor R3 ensures that the Schmitt trigger changes state very quickly.
With the resistor values shown in the circuit diagram, the lamp will remain lit for roughly 2.5 seconds per μ F of C1. Thus a 10 μ capacitor would give an interval of roughly 25 seconds.
The circuit can be powered by four 1.5 V cells connected in series. If a larger lamp is required, three 4.5 V cells connected in series can be employed. Alternatively, for really 'heavy-duty' applications, the four parallel-connected inverters can be replaced by a transistor, as shown in figure 2. The supply voltage should be matched to the voltage rating of the lamp and may lie between 4.5 and 15 V. The current through the lamp should not exceed 500 mA in that case.

107 Car light reminder

There are many different designs for alarm circuits which remind the motorist to switch off his lights before leaving the car. The advantage of the circuit presented here is that no extra components have to be connected in series with the existing wiring, so that it will not effect the operation of the car electrical system. Furthermore the circuit is extremely simple, consisting solely of a DC buzzer, a double-pole double-throw switch, and a handful of diodes (depending upon the number of functions to be monitored).
The accompanying circuit diagram shows how the car headlamps, sidelights, fog lamps, and heated rear windscreen can be monitored. Note however that the circuit will not indicate if the above functions are actually working!
With S1 in the position shown in the diagram, the buzzer will be activated the moment the engine is switched off and one of the monitored functions is left on. Switching off the function concerned also cuts the buzzer.
If one wishes to leave a particular function on, e.g. the parking lights, then S1 should be switched to its alternative position, whereupon the buzzer is disabled until the engine is started again. The alarm can then be re-armed by switching S1 back to its original position.
Normally there is a resistance (R_L) in parallel with the ignition system (as a result of the various dashboard indicator lights, fuel guage etc.)



which is sufficiently small to ensure that the buzzer will be activated in the event of the contact breaker coming to rest in the open position when the car is stopped. If, however this resistance is too large, a 100...200 Ω resistor (2 W) can be connected in parallel with R_L. It may be preferable to use a small lamp – roughly 0.1 W/12 V – since due to its positive temperature coefficient, the more power it dissipates, the greater its resistance becomes.

M. Penrose

108 TAP switch

The advantage of this circuit for a touch activated switch is that it requires only one set of contacts and uses only two inverters, two resistors and a pair of capacitors. The circuit functions as follows: At switch on, the input of N1 is low, since C1 is discharged. Since the input of N1 is low, the input of N2 must be high and the output of N2 low, which of course holds the input of N1 low – thus the circuit is latched in a stable state.

In the meantime capacitor C2 charges up, via R2, to logic '1'. If the touch contacts are now bridged, the logic '1' on C2 is applied to the input of N1 (C2 > C1), taking the output low (and the output of N2 high). The state of the Q

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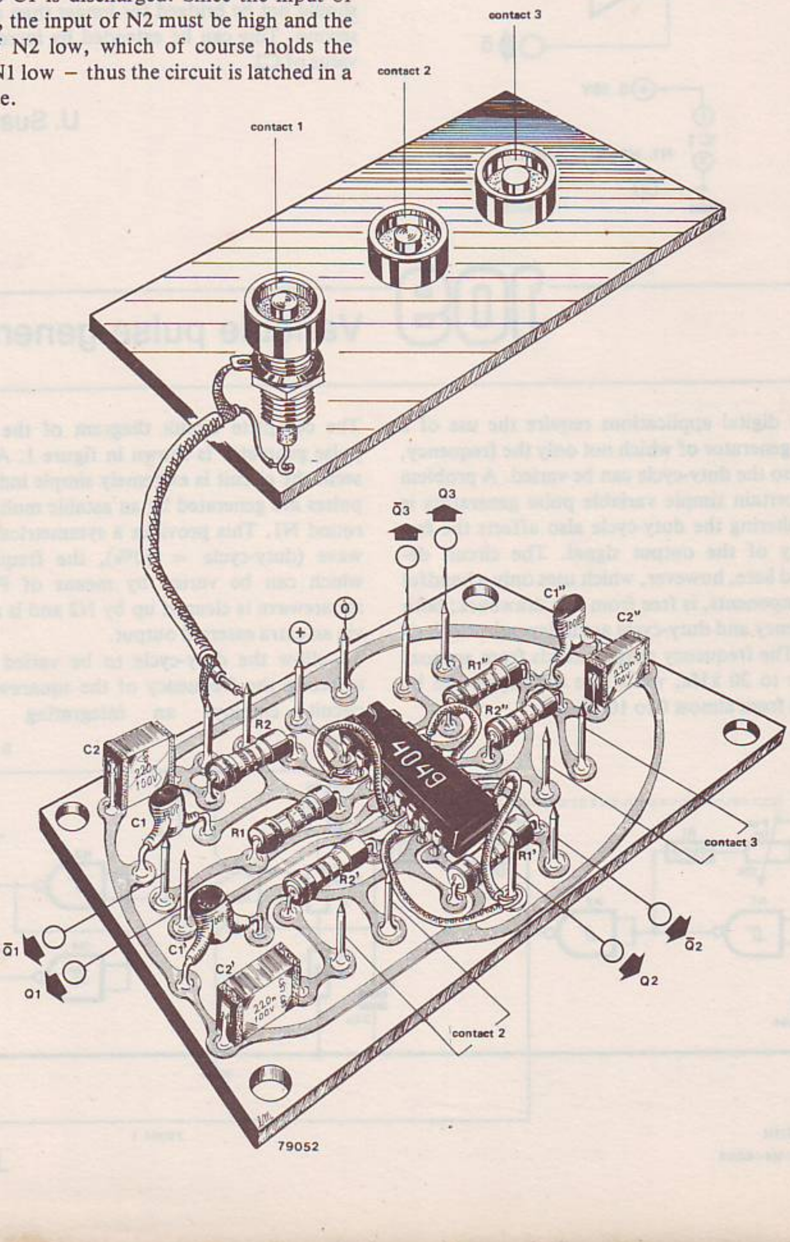

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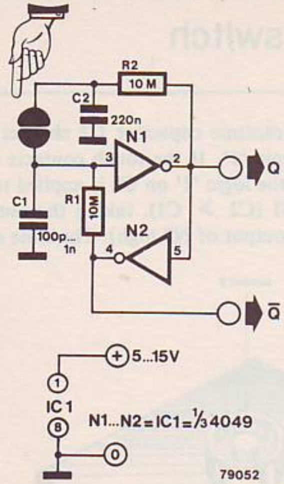

and $\overline{Q}$ outputs is thus inverted.

Bridging the contacts again causes C1 to discharge into C2 so that the outputs revert back to their original state. If the contacts are bridged for longer than the time constant R2 · C2, then the outputs will change state again. If the contacts are permanently bridged, the circuit will in fact oscillate at a frequency determined by the above time constant.

With the component values shown, the contacts should not be bridged for longer than approx. 1 second. This can be extended by increasing the value of C2.

U. Sussbauer

109 Variable pulse generator

Many digital applications require the use of a pulse generator of which not only the frequency, but also the duty-cycle can be varied. A problem with certain simple variable pulse generators is that altering the duty-cycle also affects the frequency of the output signal. The circuit described here, however, which uses only a handful of components, is free from this drawback; both frequency and duty-cycle are independently variable. The frequency range extends from approx. 1 kHz to 20 kHz, whilst the duty-cycle can be varied from almost 0 to 100%.

The complete circuit diagram of the variable pulse generator is shown in figure 1. As can be seen. the circuit is extremely simple indeed. The pulses are generated by an astable multivibrator round N1. This provides a symmetrical squarewave (duty-cycle = 50%), the frequency of which can be varied by means of P1a. The squarewave is cleaned up by N2 and is available via an extra external output.

To allow the duty-cycle to be varied without affecting the frequency of the squarewave, the circuit employs an integrating network

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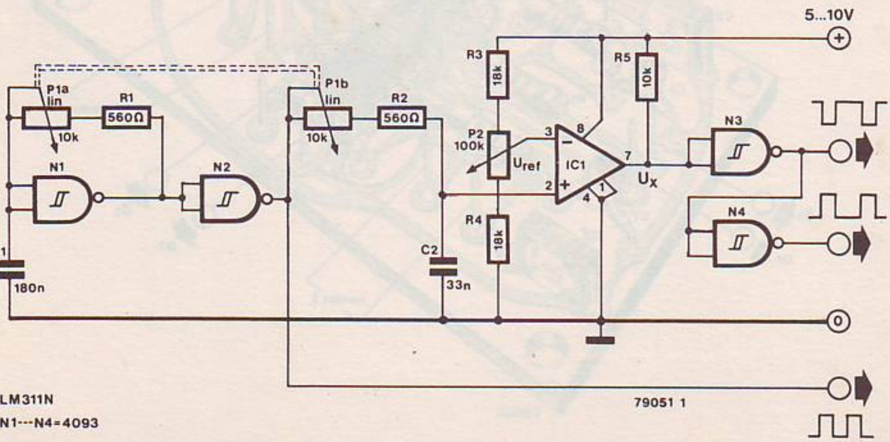

IC1 = LM311N
IC2 = N1---N4 = 4093

(P1b/R2/C2) and a comparator (IC1). The RC-constant of the integrating network (C2 = 1/6 · C1) is chosen such that the voltage across C2 may vary between approx. 20 and 80% of the supply voltage, U_b. Whenever this voltage exceeds the reference voltage on the inverting input of the comparator, the output of the latter changes state. The result is therefore a squarewave signal (U_x) whose duty-cycle is determined by the reference voltage (U_{ref}) of the comparator. This process is clearly illustrated in the timing diagram of figure 2. By varying the voltage at the inverting input of the comparator it is therefore possible to adjust the duty-cycle of the squarewave as desired without affecting the frequency.

There now remains the question of what happens to the duty-cycle if the frequency of the squarewave is varied. Normally the duty-cycle would be influenced by the frequency change, however due to the use of a twin-ganged potentiometer (P1a/P1b), in the circuit shown here, the RC-constant of the integrating network will vary in sympathy with that of the multivibrator. If the frequency, f, of the multivibrator is increased to x·f, the period of the resulting squarewave will be reduced by a factor x. However since the RC-constant of the integrating network is likewise reduced by a factor x, the duty-cycle of the squarewave at the output of the comparator will remain unchanged. It is not difficult to see that altering the RC-constant of the integrating network will not affect the shape of the charge curve of C2, so that the pulse diagram of figure 2 is also valid for any frequency x·f. The ratio T1/T2 and thus the duty-cycle (= T1/T2 x 100%) is therefore constant.

The values of R3, R4 and P2 are chosen such that the reference voltage at the inverting input of IC1 may vary between 13 and 87% of the supply voltage. As already mentioned, the voltage across C2 can vary between 20 and 80% of supply. Thus it is possible to vary the duty-cycle of the output signal between virtually 0 (i.e. no output signal) and 100% (DC voltage).

The two remaining Schmitt-trigger gates of IC2 are used at the output, N3 to further square up the output signal and N4 to provide an inverted version. Thus if a squarewave with a duty-cycle of 30% is present at the output of N3, the output of N4 will provide a squarewave of identical frequency but with a duty-cycle of 70%.

With the component values as shown in figure 1, the frequency range of the circuit extends from approx. 1 kHz to 20 kHz.

The frequency range can be altered if desired; the essential parameters of the circuit are given by the following equations:

$$C1 = 6 \times C2$$

$$P1a = P1b \text{ and } R1 = R2$$

$$f = \frac{1}{(P1a + R1) \cdot C1 \cdot 0.4}$$

It is also possible to control the amplitude of the output signal by connecting a 22 k potentiometer between the output of N3 or N4 and ground. The output signal can then be taken from the wiper of the potentiometer.

The supply voltage for the ciruit need not necessarily be stabilised, however if any sort of demands are to be placed upon the stability of the frequency, amplitude or duty-cycle, it is best to employ a voltage regulator. Since the entire circuit consumes no more than roughly 20 mA, a regulator from the 78L-series is the obvious choice. Depending upon the supply voltage, the 78L05, 78L06, 78L08, 78L09 and 78L010 should prove suitable.

K. Kraft

Figure 1. The circuit of the variable pulse generator employs only two ICs, yet both the frequency and duty-cycle are independently variable.

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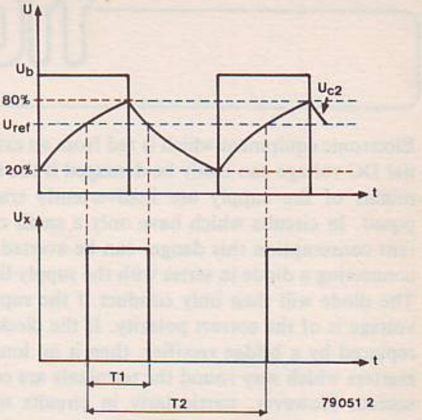


Figure 2. This timing diagram illustrates how the duty-cycle of the output signal is determined by the reference voltage of the comparator (U_{ref}). Furthermore, by varying the RC-constant of the integrating network in sympathy with that of the multivibrator it is possible to make the duty-cycle independent of the frequency.

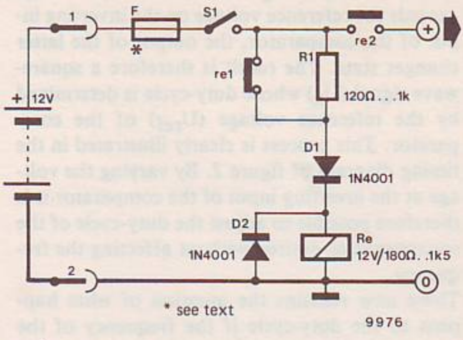
110 DC polarity protection

Electronic equipment which is fed from an external DC voltage can easily be damaged if the terminals of the supply are inadvertently transposed. In circuits which have only a small current consumption this danger can be averted by connecting a diode in series with the supply line. The diode will then only conduct if the supply voltage is of the correct polarity. If the diode is replaced by a bridge rectifier, then it no longer matters which way round the terminals are connected. However, particularly in circuits with larger current consumptions, this approach is somewhat unsatisfactory, since it leads to noticeable power losses.

A more elegant solution, which results in no voltage loss and virtually no power loss, and hence is suitable for circuits carrying relatively large currents, is shown in the accompanying diagram. The component values were chosen for a DC supply of 12 V.

The circuit should be mounted inside the equipment it is meant to protect and the external supply voltage connected to terminals 1 and 2. Assuming the polarity of the supply is correct, once the on/off switch, S1, is closed, the relay, Re, will pull in, causing two things to happen. The normally closed contact, re1, will open, reducing the relay current through R1. Since the drop-out current is less than the pull-in current, assuming R1 is the correct value, relay Re will remain energised. This little trick reduces the dissipation in the protection circuit.

Secondly, the normally open contact, re2, will close, thereby applying power to the rest of the equipment.



However, if the terminals of the supply are transposed, diode D1 will be reverse-biased, preventing the relay from being pulled in. Diode D2 suppresses any inductive voltages produced when the relay coil is de-energised.

If there is a fuse in the supply line of the equipment, then it is recommended that this be inserted between the supply and the protection circuit, so that it will blow should a fault occur in the latter. The current consumption of the protection circuit is so small compared with that of the equipment it guards that there is no need to alter the rating of the fuse.

The values of the components in the circuit can of course be modified to suit other supply voltages. One should bear in mind that the pull-in voltage of the relay, Re, should be the same as the supply voltage.

The value of R1 will depend to a certain extent on the type of relay used, and is best determined experimentally.

111 Disco lights

Flashing lights are very much an integral part of the disco scene nowadays. Usually the lights are controlled or modulated in some way by the music, i.e. the lights turn on and off or become dimmer or brighter in accordance with the volume or pitch of the audio signal. The circuit described here can be used either as a dimmer, 'running light' controller, or form the basis of a light organ.

The circuit, as shown in figure 1, is divided into a number of separate blocks, each of which has a distinct function. The supply stage is of course an essential, although if the circuit is used exclusively as a dimmer, IC2 and C6 can be omitted. The remainder of the dimmer circuit is contained in block (a).

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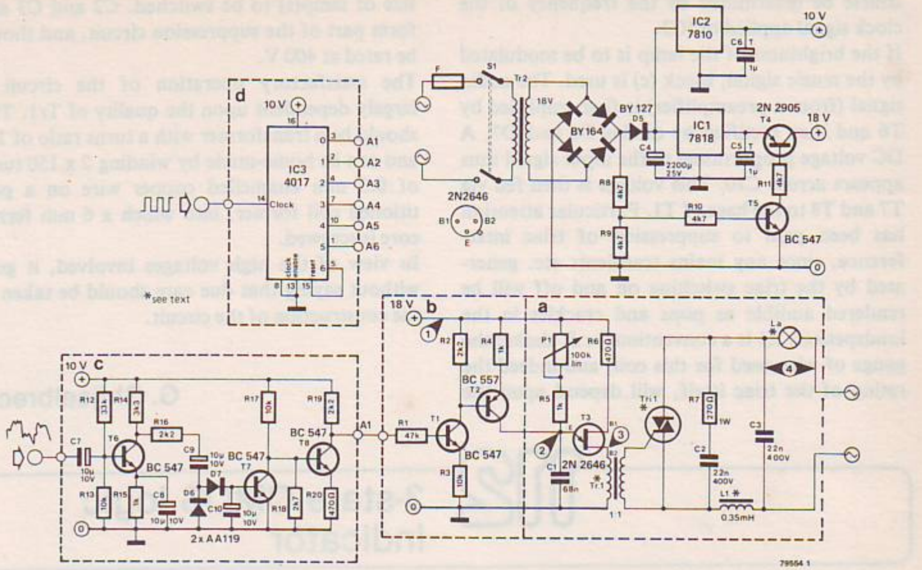

Together with T3, components P1, R5, R6 and C1 form a sawtooth generator which, via the pulse transformer Tr1, is used to trigger the triac. To ensure good synchronisation with the mains waveform the triac is turned off every 10 ms. This is achieved by transistors T4 and T5 momentarily removing the supply to the oscillator (see figure 2). The position of P1 determines the brightness of the lamp, which is continuously variable from zero to full on. With the aid of block (b), the brightness of the lamp can be varied by an external control voltage (4...8 V) which can be derived from a variety of add-on circuits. An example of one such control circuit is shown in block (d). By connecting each A-output of the 4017 to a circuit consisting of blocks (a) and (b), a running light effect is obtained. The 'speed' of the running light will of

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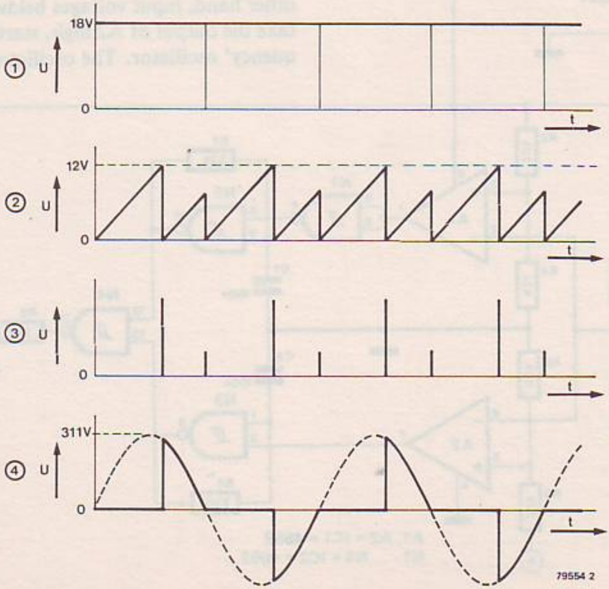
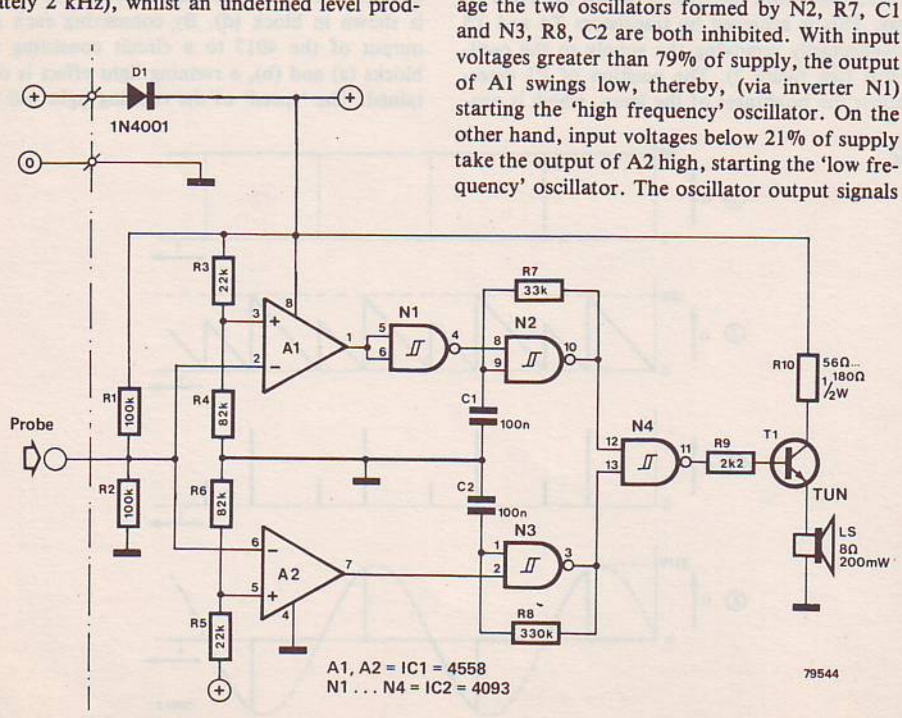


course be determined by the frequency of the clock signal applied to IC3.
If the brightness of the lamp is to be modulated by the music signal, block (c) is used. The audio signal (from a preamplifier) is first amplified by T6 and then rectified by diodes D6 and D7. A DC voltage proportional to the input signal thus appears across C10. This voltage is then fed via T7 and T8 to the base of T1. Particular attention has been paid to suppression of triac interference, since any mains transients etc. generated by the triac switching on and off will be rendered audible as pops and crackles in the loudspeaker. L1 is a conventional r.f. choke; the gauge of wire used for this coil, and indeed the rating of the triac itself, will depend upon the size of lamp(s) to be switched. C2 and C3 also form part of the suppression circuit, and should be rated at 400 V.
The satisfactory operation of the circuit is largely dependent upon the quality of Tr1. This should be a transformer with a turns ratio of 1 : 1 and can be home-made by winding 2 x 150 turns of 0.3 mm enamelled copper wire on a partitioned coil former, into which a 6 mm ferrite core is screwed.
In view of the high voltages involved, it goes without saying that due care should be taken in the construction of the circuit.

G. Ghijselbrecht

112 3-state CMOS logic indicator

The following circuit will provide an audible indication of CMOS logic states. Logic '0' is represented by a low frequency tone (roughly 200 Hz), logic '1' by a high frequency tone (approximately 2 kHz), whilst an undefined level produces no output signal.
The circuit functions as follows: two comparators are connected such that at voltage levels between roughly 21 and 79% of the supply voltage the two oscillators formed by N2, R7, C1 and N3, R8, C2 are both inhibited. With input voltages greater than 79% of supply, the output of A1 swings low, thereby, (via inverter N1) starting the 'high frequency' oscillator. On the other hand, input voltages below 21% of supply take the output of A2 high, starting the 'low frequency' oscillator. The oscillator output signals

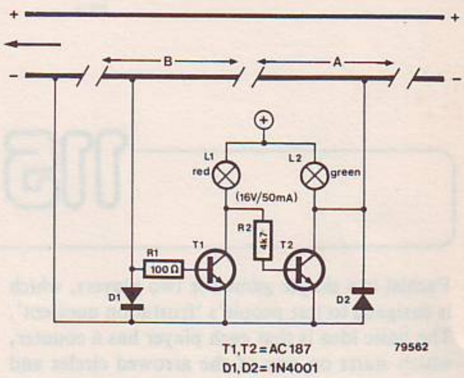


are fed to a simple buffer stage and then to a suitable loudspeaker.
The power supply should be drawn from the circuit under test, and must lie between roughly 5 and 15 V.

D. Hackspiel

113 Model railway block section controller

This simple circuit offers model railway enthusiasts a cheap alternative to the fairly expensive block section controllers which are available commercially. The circuit suffers from one disadvantage, namely that it can be used to control traffic in just one direction. However, cost may dictate that this is acceptable.
The circuit and how it is connected to the rails, is shown in the accompanying diagram, where the direction of the trains is assumed to be from right to left. As can be seen, the 'earth' rail is broken at three places (using insulating track sections which are available in model shops). The lengths of rail sections A and B will influence at what point the train stops, and should be chosen to suit individual circumstances (the length of the train(s) for example). The red and green lamps (L1 and L2) are built into a set of signals.
The circuit works as follows: As long as there is no train in the vicinity, the green lamp (L2) will be lit and section A of the track is connected to earth via the circuit. Transistor T1 is turned off, so that transistor T2 is turned on via L1 and R2.
Should a train then approach, nothing will happen as long as it remains on block A of the track. When the train advances to block B, however, diode D1 is forward biased via the motor of the train, which will slow down slightly since the diode drops 0.7 V of the supply voltage. The voltage dropped across the diode also turns on T1, causing the red lamp (L1) to light up. At the same time T2 turns off, extinguishing the green

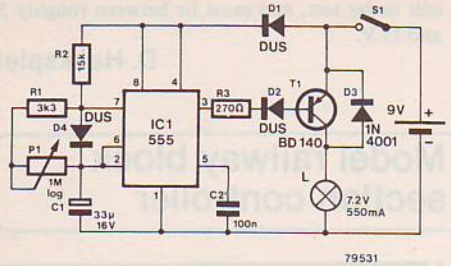


lamp and breaking the connection between block A of the track and earth. A subsequent train entering block A of the track is therefore forced to a stop.
As soon as the first train leaves block B, the initial situation is restored, i.e. T2 conducts, the green lamp is turned on and the connection between block A of the track and earth is restored. The train waiting in block A can therefore continue on its way.
The circuit can also be used to control a crossing. The 'A' sections of track are laid before the crossing and the B section forms the crossing itself. The signals are of course positioned on the approach to the crossing.

A. van Kollenburg

114 Burglar's battery saver

Elektor attempt to cater for everyone and included here is a circuit for gentlemen in the nocturnal profession. Put an end to stumbling in the shrubbery with the torch light controller described here. Incidentally it is also an excellent battery saver. Varying the brightness of a torch appears simple enough but using a series resistor or potentiometer is out of the question since power is dissipated in the form of heat. One solution is not to use a DC supply voltage but



rather a squarewave with a variable duty cycle. The brightness of the lamp then depends upon the length of the duty cycle.
In the circuit shown, a 555 timer is connected as an astable multivibrator and used to supply the squarewave. The duty cycle of the squarewave can be varied by potentiometer P1. Diodes D1... D3 protect the circuit if the polarity of the battery is reversed in which case the circuit will not operate and the torch will be 'full on'. Gentlemen, do not change your batteries in the dark!

C. Hentschel

115 Pachisi

Pachisi is a simple game for two players, which is designed to test people's 'frustation quotient'. The basic idea is that each player has a counter, which starts on one of the arrowed circles and then attempts to move round the board to the white rectangle in the centre of the 'M'. The players move alternately and the first person's counter to reach 'home' is the winner. Four different types of move are possible: forwards, backwards, onto the next white circle, and onto

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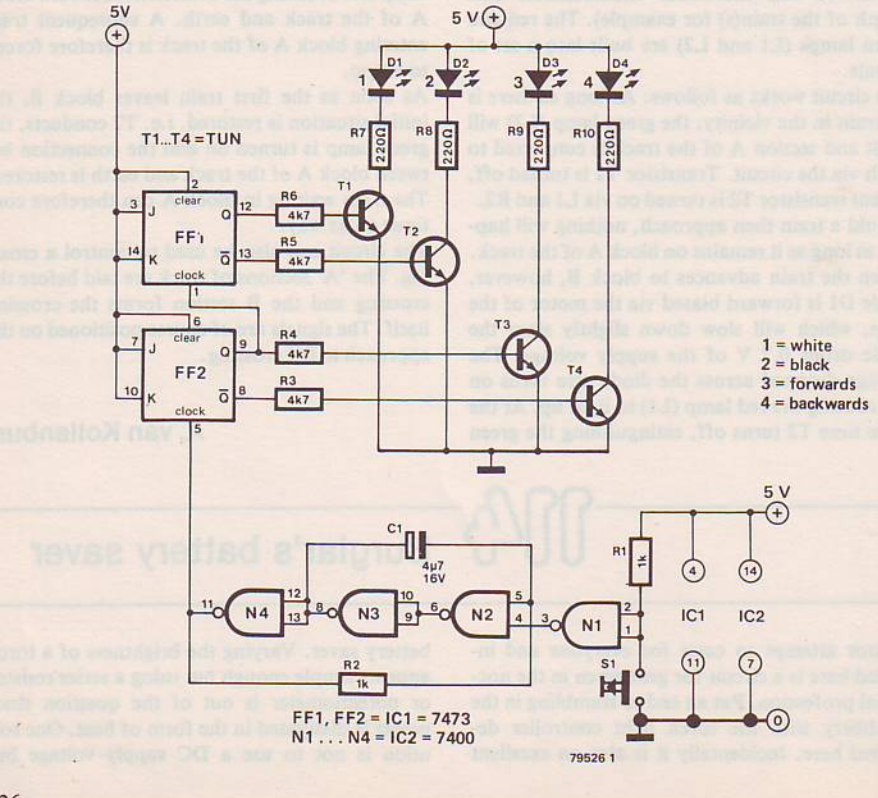

the next black circle. Thus it is effectively possible to move either one or two steps forwards or backwards each turn. If one player lands on the circle currently occupied by his opponent, the former is declared the winner, whilst if a player retreats backwards off the edge of the board, he is deemed to have lost.

The player's moves are determined by two pairs of LEDs. One pair decides whether the move is forwards or backwards, and the other pair whether it is to a white or black circle. Each time the pushbutton switch S1 (see circuit diagram) is pressed, a new random combination occurs. Thus it could happen that one player is on the point of winning when he is forced to take 'two steps backwards'!

The actual circuit is straightforward. Two flip-flops form a two-bit binary counter, which is clocked by an oscillator built round NANDs N1...N4. The oscillator is only enabled when S1 is closed. The output state of the counter is displayed via transistors on the four LEDs.

H.J. Walter

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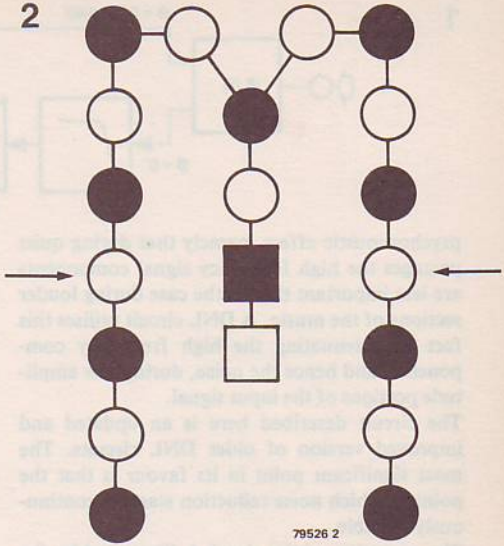
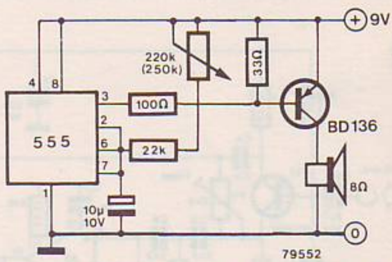


116 Metronome

Although not exactly revolutionary, the circuit shown here is both very cheap and reliable. The well-known 555 timer IC is connected as an astable multivibrator, and delivers a regular train of pulses which are rendered audible via the transistor and loudspeaker. The frequency of the metronome can be varied with potentiometer P1. A 9 V supply voltage means that the circuit can easily be powered by batteries.

If a loudspeaker with an impedance of less than 8 Ω is used, it should be preceded by a series resistor (1 W) which will compensate for the difference in impedance (and – due to the lower current consumption – ensure that the batteries last longer).



W. Kluifhout

117 Improved DNL

Dynamic Noise Limiting (DNL) is a noise reduction system patented by Philips, which is particularly useful for the reproduction of (cassette) tape recordings. As the name suggests, the system is dynamic, i.e. the noise is only suppressed at the moments when it is most intrusive which, in the case of a music signal, is during the quieter passages. The system also exploits an interesting

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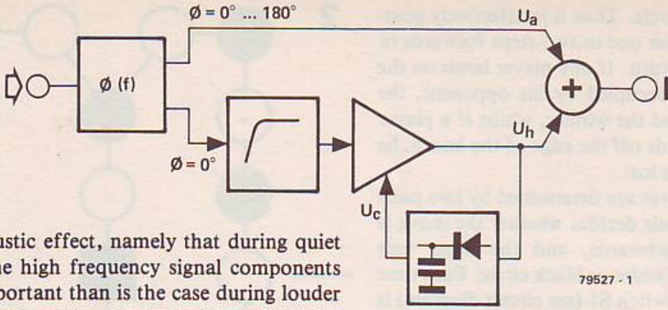
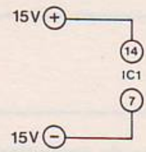


psychoacoustic effect, namely that during quiet passages the high frequency signal components are less important than is the case during louder sections of the music. A DNL circuit utilises this fact by attenuating the high frequency components, and hence the noise, during low amplitude portions of the input signal.

The circuit described here is an updated and improved version of older DNL circuits. The most significant point in its favour is that the point at which noise reduction starts is continuously variable.

The operation of the circuit is illustrated by the block diagram of figure 1. The input signal is fed to a phase shifter, which provides two output signals. One of these signals, u_a, is equal to the input signal, but is subjected to a frequency-dependent phase shift varying from 0° for low frequency signals to 180° for high frequency signals. The second output signal is identical to the input signal in all respects, including phase, and is fed to a highpass filter and then to an amplifier. The gain of the amplifier is determined by the feedback signal, u_c, which is obtained by peak rectifying the amplifier output. The result is dynamic compression/limiting of the high frequency signal components, i.e. the latter are amplified to a constant level, regardless of input signal level. The amplifier output, u_h, is summed with the phase-shifted version of the input signal. Since the phase shift was frequency-dependent, the high frequencies present in the two signals will tend to cancel. However due to the limiting effect of the amplifier stage, the



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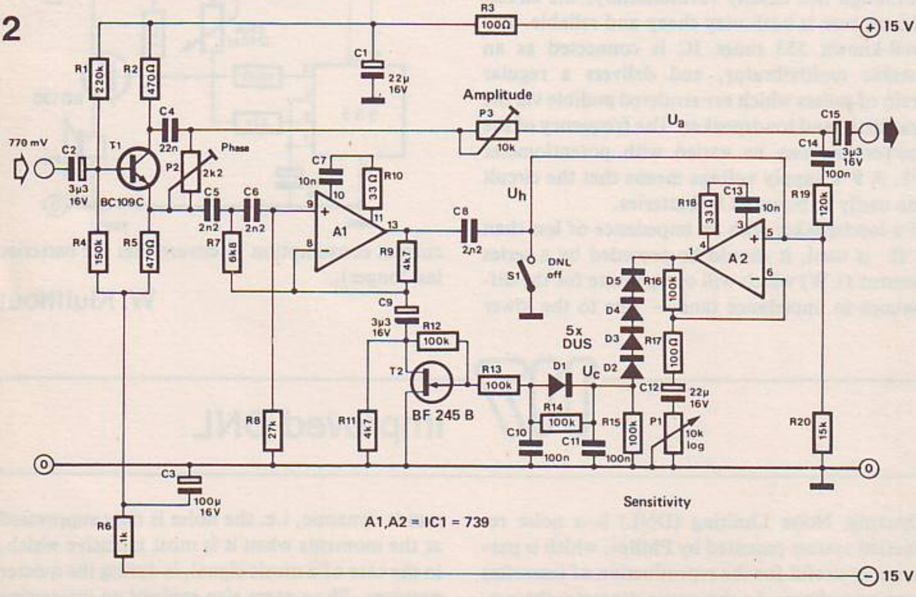


greater the amplitude of the input signal, the less the cancellation, and the smaller the attenuation of the higher frequencies. The noise reduction is therefore severest at low input signal levels, i.e. during the quieter passages of music.

The complete circuit diagram of the DNL circuit is shown in figure 2. The phase shifter is formed by T1, the frequency dependence of the shift being obtained by combining the collector (Φ = 180°) and emitter (Φ = 0°) signals via P2 and C4. The highpass filter is realised by the circuit round op-amp A1. This filter has a third-order Butterworth response with a turnover frequency of 5.5 kHz. The filter output is amplified/limited by A2. The gain of A2, and with it the sensitivity of the circuit, can be varied by means of potentiometer P1.

The peak detector consists of 4 series-connected diodes, which ensures that the control signal, u_c, is only present when the input signal rises above a certain level. A FET, T2, is used to form the voltage controlled attenuator in the feedback loop of A2. The two signals u_a and u_h are summed via preset potentiometer P3 and the series connection of R19 and C14.

The DNL function of the circuit can be rendered inoperative by means of switch S1, which simply shorts the signal u_h to earth.

During construction care should be taken to ensure that the output signal of op-amp A2 is kept at least several centimetres from the signal-carrying leads, so as to prevent the possibility of crosstalk.

The circuit can be set up by driving it with a pure noise signal, such as that from an off station FM tuner, and varying P2 and P3 for maximum attenuation.

The circuit as shown is optimised for standard level audio signal levels, i.e. 0 dB = 770 mV RMS, but can also be used for other signal levels.

R.E.M. van den Brink

118 Resistance bridge

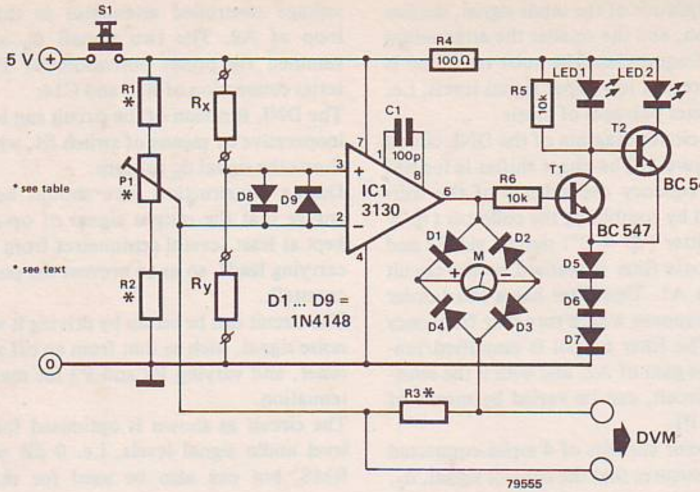
Generally speaking, resistors with a 5% tolerance are more than adequate for most circuits. From time to time there may be occasions when 1% resistors are required, or when the value of two resistors must be matched to within 1%. This is the case with for example digital meters, where it is worth the extra expense of using very accurate attenuator resistors in order to fully exploit the accuracy offered by a digital display.

The circuit described here allows two resistors, R_x and R_y, of the same nominal value to be compared with one another, and the difference to be expressed directly in per cent. The accuracy and stability of the circuit are better than 0.1%, and resistors from 10 Ω to 10 M Ω can be measured, providing the maximum permissible dissipation is not exceeded, i.e. ¼ W types for example should be greater than 27 Ω .

The operation of the circuit is based upon the
The operation of the circuit is based upon the
resistance bridge formed by R_x, R_y and the voltage divider R1, P1 and R2. If R1 and R2 are exactly the same value, the bridge current will be proportional to the extent to which R_x and R_y deviate from the mean value of these two resistors. For small differences between R_x and R_y the current is, to all intents and purposes, proportional to the difference between the two resistors. The percentage difference between the two 'unknown' resistances is expressed directly on the scale regardless of which resistor is the greater. However with the aid of the simple comparator formed by T1 and T2, which of the two resistors is the greater can be displayed on LEDs D1 and D2.

The circuit can be adapted to suit a variety of different meters. If a centre-zero reading meter or a DVM (with a floating input) are available these would be ideal in which case components D1...D7, R4...R6, T1, T2 and the two LEDs can

Table:	scale	meter M	R1 = R2	P1	R3	DVM
	0- 3%	0- 60 μA	1k2	100 Ω	5 k	−0.3... +0.3 V
	0-10%	0-200 μA	1k2	100 Ω	5 k	−1 ... +1 V
	0-10%	0-500 μA	475 Ω	50 Ω	2 k	−1 ... +1 V
	0-10%	0-200 μA	1k2	100 Ω	500 Ω	−0.1... +0.1 V
	0- 1%	0- 50 μA	475 Ω	50 Ω	2 k	−0.1... +0.1 V

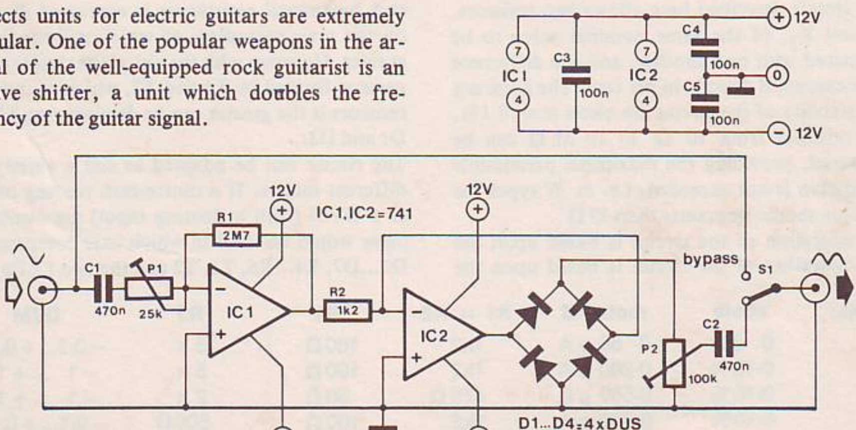


be omitted. A universal meter with a 0-10 or 0-30 scale would also be suitable.
The table lists other examples of possible meters and indicates the component changes required as well as the range scale obtained.
High stability metal oxide or 1% precision wire-wound resistors should be used for R1, R2 and R3.
Calibrating the circuit is quite straightforward. P1, which should preferably be a multi-turn type, is provisionally set to the mid-position and two resistors of the same nominal value are connected in circuit. The meter reading is noted, and then the resistors changed over. If the new reading is the same as the first, no further adjustment is required. If that is not the case, P1 is adjusted until the average of the two readings is obtained. If desired the procedure can be repeated once more for an extra check.

J. Borgman

119 Octave shifter for electric guitars

Effects units for electric guitars are extremely popular. One of the popular weapons in the arsenal of the well-equipped rock guitarist is an octave shifter, a unit which doubles the frequency of the guitar signal.



One of the ways of achieving frequency doubling – and the approach adopted here – is full-wave rectification, as commonly carried out in power supply circuits. As can be seen from the accompanying circuit diagram, the rectification is performed by a diode bridge. By including the diode bridge in the feedback loop of IC2, the non-linear voltage characteristic of the diodes has no effect upon the signal.

Pre-amplification of the guitar pickup signal is provided by IC1. The gain of this stage is set (by P1) such that the signal is just on the point of clipping. Preset potentiometer P2 can be adjusted so that the output signal level is the same as that of the input signal. A bypass switch, S1, is included allowing the unit to be switched in and out.

The signal is not only doubled in frequency, but is also distorted. The sound becomes considerably harsher, as well as being shifted up an octave. This feature would probably be considered an asset to the contemporary rock musician.

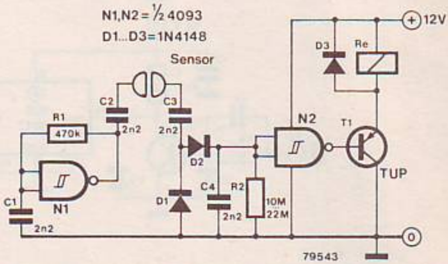
H. Schmidt

120 Liquid level sensor

An annoying drawback of many liquid level sensors is the effect of electrolytic reaction between the liquid and the sensors. Metal electrodes are prone to corrosion and consequent loss of effectiveness (reduced conductivity), with the result that they have to be replaced at frequent intervals.

One solution to this problem is to ensure that there is an AC, rather than DC potential between the sensor electrodes. The constant reversal of electrode polarity drastically inhibits the electrolytic process, so that corrosion is considerably reduced.

The actual circuit of the level sensor is extremely simple. The circuit around N1 forms an oscillator. If the two sensors are immersed in a conducting solution, C4 will be charged up via the AC coupling capacitors (C2 and C3) and the diodes, so that after a short time, the output of N2 is taken low and the relay is pulled in. The



relay can be used to start a pump, for example, which in turn controls the level of the liquid. When a conductive path between the two sensors no longer exists, C4 discharges via R2, with the result that the output of N2 goes high and the relay drops out.

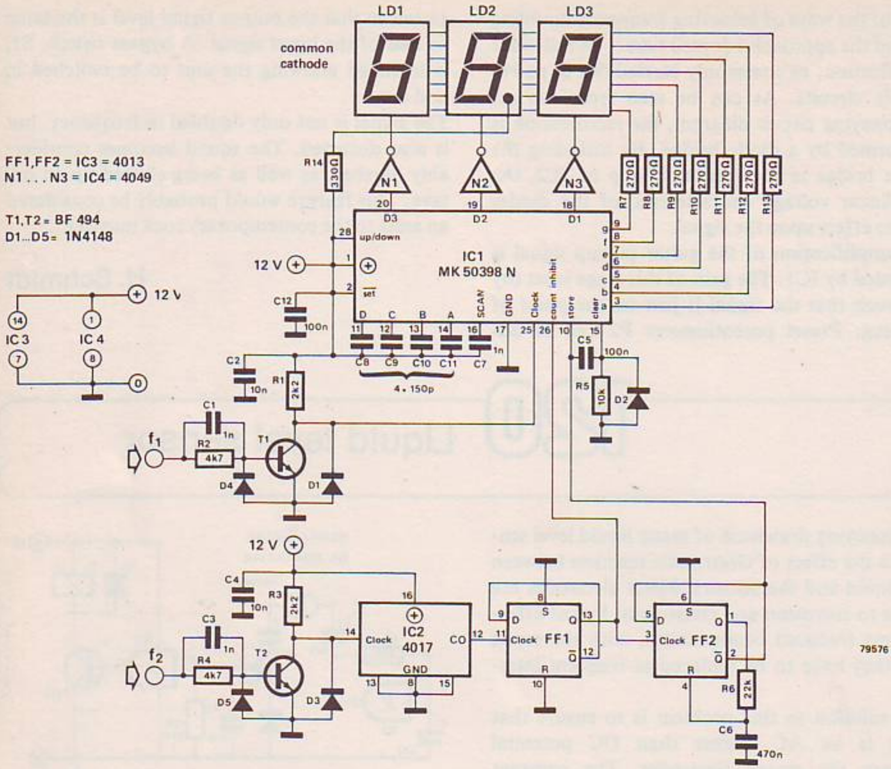
E. Scholz

121 Frequency ratio meter

There are certain situations, e.g. when checking frequency multiplier or divider circuits, PLL circuits, certain music circuits etc., where it is more important to measure the frequency ratio of two signals, rather than simple measurement of frequency itself. With the aid of the circuit shown here, the ratio between the frequency of two signals, f_1 and f_2, can be measured and displayed directly on three seven-segment displays. The circuit will measure ratios up to 99.9 with an accuracy of 0.1, providing f_1 is larger than f_2.

The heart of the circuit is the counter/display driver IC, MK 50398N, from Mostek.

The higher frequency, f_1, is fed via the input stage around T1 to the clock input (pin 25) of the counter. Pulses will be counted at this input pro-



vided pin 26 (count inhibit) is held low. Decade divider IC2 and flip-flop FF1 ensure that this pin is in fact held low for exactly ten cycles of the lower frequency signal, f_2. Thus a number appears on the displays which is ten times the ratio between f_1 and f_2. By arranging for the decimal point to light between the second and third digits, the resulting figure is thus exactly equal to the ratio f_1/f_2. Flip-flop FF2 is connected as a monostable, and is used to provide the counter IC with the correct 'store' and 'clear' pulses on pins 10 and 15 respectively.

W. Dick

122 LED lamps

When it comes to mains indicator lamps, there are basically three main options: neon lamps, incandescent lamps, and LEDs. Neon lamps have the advantage that they can be connected direct to the mains supply, and also that they consume very little power. Incandescent lamps, on the other hand, must be connected to a much lower voltage (e.g. to the secondary side of the transformer), and therefore provide only indirect indication of whether the mains supply is present, whilst as a rule dissipating a relatively large amount of power.

LEDs would represent an ideal alternative to both the above approaches, since they have a longer operating life than either neon or incandescent lamps, and dissipate no more than 20 to 30 mW. Unfortunately it is necessary to protect the LED from excessive currents by employing a series resistor, which, with a mains voltage of 240 V, will itself dissipate something over 3.5 W. The circuit shown here offers a better solution. The current through the LED is limited to a safe value not by a dropper resistor, but by the reactance of a capacitor. The advantage of this method is that no power is dissipated in the capacitor, since the current through the latter is 90° out of phase with the voltage dropped across it. The formula for calculating power dissipation for DC voltages is only valid for AC voltages provided the current and voltage are in phase i.e.

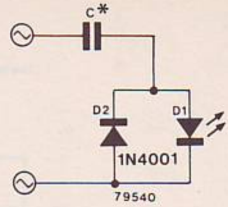
$$P_C = u_C \cdot i \cdot \cos \varphi$$

With a phase shift of 90°, which is the case with capacitors, P_C is therefore 0 W (cos 90° = 0). What little power is consumed by the circuit is entirely converted into light and heat by the LED.

The value of capacitor C, can be calculated for any given voltage, frequency and current with the aid of the following equation:

$$C \approx \frac{i}{6.28 \cdot u \cdot f} \text{ where:}$$

C is the capacitance in Farads



u is the RMS value of the mains voltage
f is the mains frequency in Hz
i is the current through the LED in Amps

With a mains voltage of 240 V, a frequency of 50 Hz and a current of 20 mA, the nearest suitable value of capacitor is therefore 330 nF. The working voltage of the capacitor should be at least twice the mains voltage.

Diode D2 is included to protect the LED from excessive reverse voltages.

U. Hartig

123 Linear thermometer

The circuit described here employs a forward-biased diode as temperature sensor. The forward voltage drop of a diode falls by approximately 2 mV for an increase in temperature of 1° C. Since this negative temperature coefficient remains the same regardless of actual ambient temperature, the scale of the thermometer will be linear.

The temperature coefficient of a diode is not of an NTC (negative temperature coefficient) resistor. However it is not possible to obtain a linear scale over a wide range of temperatures using an NTC resistor. Thus the use of a diode is justified by the wide measurement range obtained and by the ease of calibration.

The sensor diode – D1 in the circuit diagram – is a common-or-garden 1N4148, which can easily be mounted apart from the rest of the circuit.

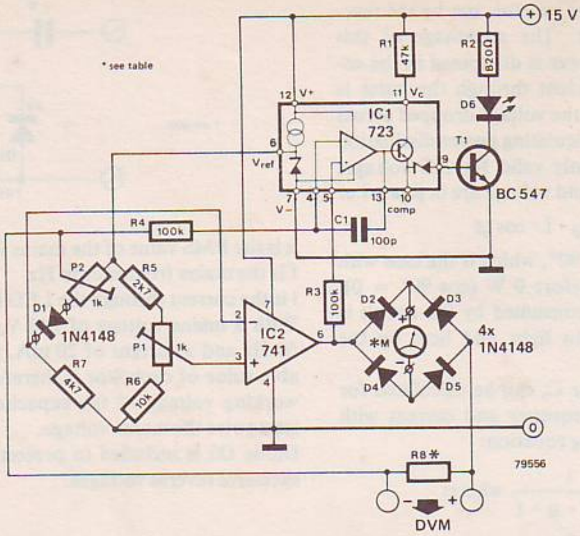
The diode forms part of a resistance bridge, comprising P1, P2, R5, R6 and R7. A reference voltage is provided by a 723. Thus the voltage on the non-inverting input of IC2 is held to a (variable) reference value via R5 and P1. Assuming the circuit is initially nulled by adjusting P1 and P2, variations in the forward voltage drop of the diode as a result of temperature fluctuations will cause the output of IC2 to swing either high or low depending upon whether the temperature rises above or falls below zero.

By using a diode bridge, D2...D5, the meter will show a positive deflection regardless of the polarity of the temperature. To provide an indication of whether the temperature is in fact above or below 0°, the output of IC2 and the reference voltage are effectively connected to the non-inverting and inverting inputs respectively

Table 1

scale	meter M	temperature	R8	DVM
0 - 30	0 - 300 μA	–30... +30° C	1 k	–0.3... +0.3 V
0 - 30	0 - 100 μA	–30... +30° C	3 k	–0.3... +0.3 V
0 - 50	0 - 300 μA	–50... +50° C	1.67 k*	–0.5... +0.5 V
0 - 50	0 - 500 μA	–50... +50° C	1 k	–0.5... +0.5 V
0 - 100	0 - 1 mA	–100... +100° C	1 k	– 1... + 1 V

* 2 x 3k3 parallel



of the 723, which thus functions as a comparator. Assuming the circuit is calibrated for zero deflection at 0° C, as the temperature falls, the voltage drop across the diode increases, therefore the voltage on the inverting input of IC2 falls, the output of IC2 goes high, taking the non-inverting input of IC1 high and with it the output of IC1. Transistor T1 therefore turns on, lighting the LED. When the temperature rises above 0° C, the reverse process occurs, resulting in the LED being extinguished.

Resistor R8 is included to allow the use of a DVM (with floating input) as a means of display. The accompanying table lists a number of alternative values for R8 along with the measurement ranges obtained for various (moving coil) meter scales. Of course, if a DVM is used, then the moving coil meter as well as D2...D5, R1... R4, T1 and the LED can be omitted.

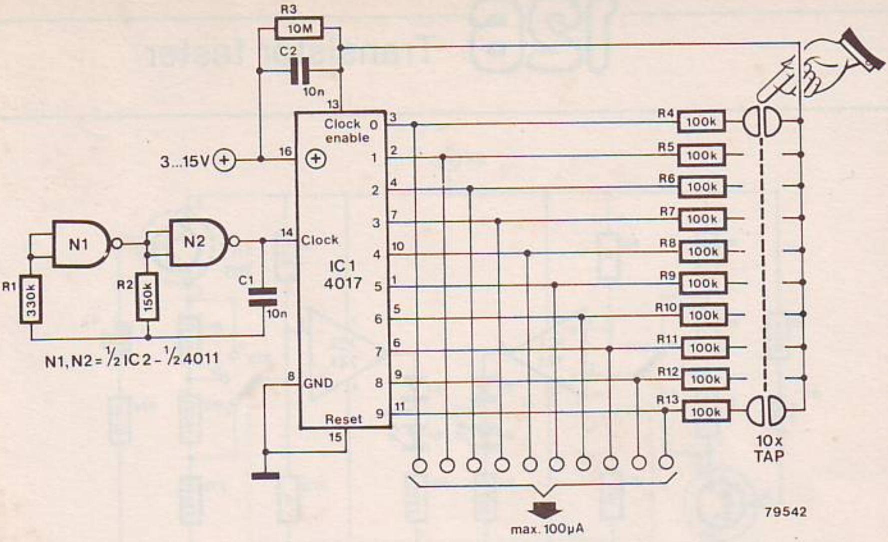
The circuit can be calibrated by suspending the sensor diode (together with a suitable length of connecting wire!) in crushed ice which is beginning to melt. With P2 provisionally set to the mid-position, P1 is then adjusted so that zero deflection on the meter (or zero voltage across R8) is obtained. The diode is then dipped into boiling water, whereupon P2 is adjusted until a voltage of 1 V over R8 is obtained. The above procedure can then be repeated. It is best to use distilled or demineralised water for both steps of the calibration procedure.

J. Borgman

124 Ten channel TAP

TAP (read: touch) switches come in all shapes and sizes, mainly as momentary action or simple on/off (latched) switches. Using only a handful of components, it is possible to construct a 'ten-channel' TAP, i.e. a ten-pole touch switch. When one of the ten sets of contacts is touched, the corresponding output will be taken high.

The heart of the circuit is formed by a CMOS decade counter/decoder, 4017, which is clocked by a simple CMOS oscillator. However when the contacts are open, the counter is inhibited, since the clock enable input is held high. The same is true if the contacts are bridged, but the corresponding output is already high, since in that case the additional skin resistance will have no effect. However, if the corresponding output is low when a set of contacts is touched, the skin resistance (which is negligibly small compared to the other resistances) forms part of a voltage divider, thereby pulling the clock enable input low.



The counter is started and increments until the output in question is taken high, whereupon the clock enable input is once more taken high and the count is stopped.

Capacitor C2 is included to suppress mains transients etc., whilst R4...R14 prevent the possibility of a shock in the event of a short between the contacts.

It must be emphasised that when the counter is started, each output in turn will go high (for a very short period) until the selected channel is reached. This should not prove to be a problem with most applications, however provision must be made for this when used with flip-flops and other edge triggered devices.

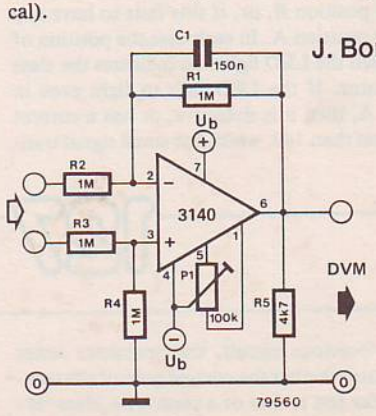
C. Horevoorts

125 Floating input for DVM

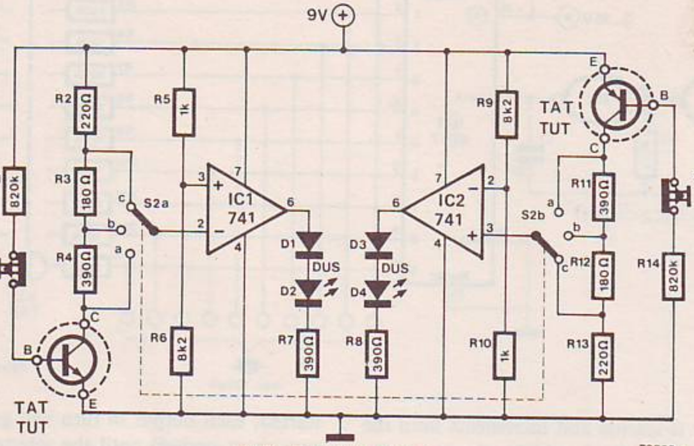
Digital voltmeters are now in widespread use and growing ever more popular. Many of the cheaper types of DVM however suffer from a slight drawback in that they have an earthed input (i.e. one of the input terminals is connected to earth or to a fixed voltage level). In many cases this is not particularly important, however there are situations (if the DVM is used in conjunction with an add-on unit such as an AC millivoltmeter, for example) where it can be something of a nuisance. With the aid of the following circuit, formed around a differential amplifier, any DVM can be provided with a floating input.

It is recommended that 1% (metal film) types are used for the 1 M resistors (R1...R4). The output voltage of the circuit is adjusted to 0 V by means of P1 (with the input short-circuited). The supply voltages $+U_b$ and $-U_b$ can be anywhere between 3 and 20 V (provided they are symmetrical).

J. Borgman



126 Transistor tester



TAT = Transistor A Tester
TUT = Transistor Under Test

Although not a precision instrument, this transistor tester should nonetheless prove a useful aid for checking the quality of 'job lots' of transistors. The circuit will determine whether or not a transitor is defective, and whether the current gain of the transistor puts it in the class of 'A'-type transistors (current gain 140...270), 'B'-type transistors (270...500), or 'C'-type transistors (greater than 500).

To test for example an NPN transistor, the device is inserted in the appropriate socket (TUT = transistor under test) and S2 switched to position C. If LED D2 lights up, the transistor is type C, if the LED remains out then S2 should be set to position B, or, if this fails to have any effect, to position A. In each case the position of S2 in which the LED lights up indicates the class of transistor. If the LED fails to light even in position A, then it is defective, or has a current gain of less than 140, which for small signal transistors means that they are basically unusable. The base current to the transistor under test can be interrupted by means of pushbutton switch S1. If the LED does not go out, it means a short exists between collector and emitter of the transistor.

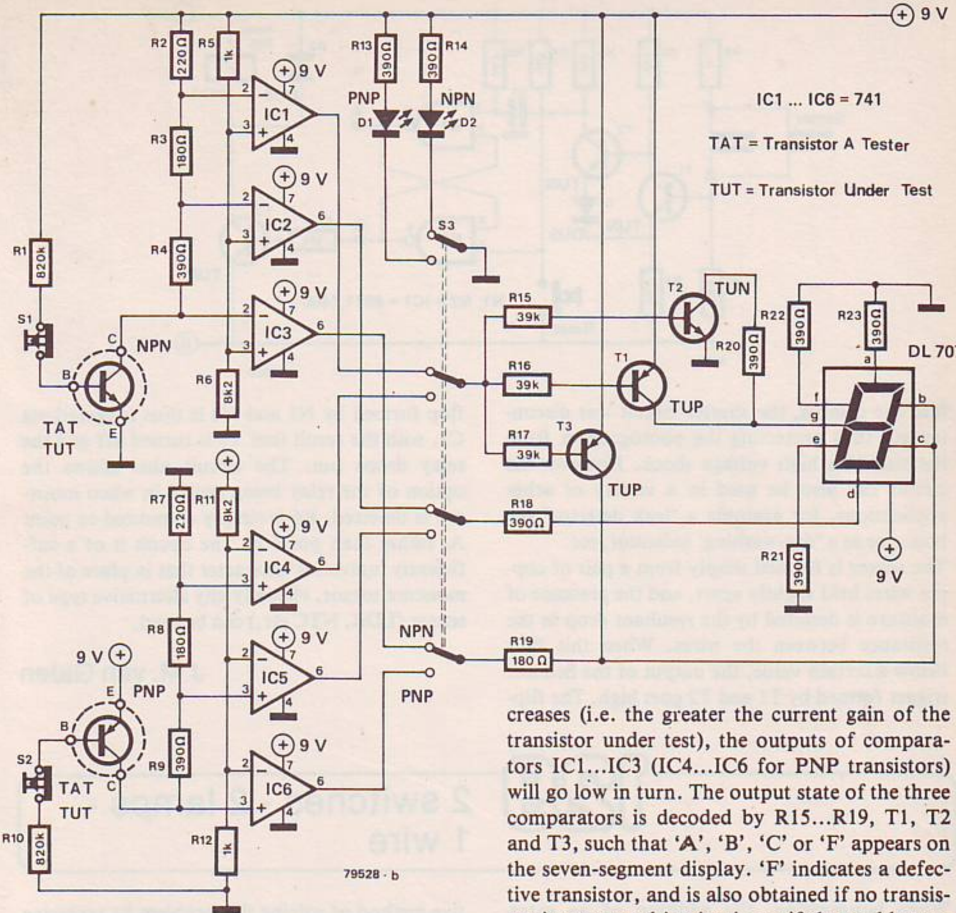
The operation of the circuit is quite simple: The transistor under test receives a base current of 10 A via R1. Assuming the transistor is not defective, this results in a voltage drop across R2... R4, and depending upon the position of S2, a portion of this voltage is compared with a fixed reference voltage by IC1. The operation of the right hand side of the circuit is virtually identical, except that it is arranged for PNP transistors.

The circuit can be powered by battery.

R. Storn

127 'De luxe' transistor tester

Like the previous circuit, this transistor tester will indicate whether the current gain of the transistor under test is that of a class 'A'-, class 'B'- or class 'C' type. The circuit will also determine whether or not the transistor is defective. The advantage of this design, however, is that the



class of transistor is automatically determined and shown directly on a seven-segment display.

The operation of the circuit is in many respects similar to its predecessor. Depending upon the current gain of the device under test, a certain DC voltage is dropped across resistors R2...R4 in the case of NPN transistors, or across R7...R9 in the case of PNP transistors. As this voltage increases (i.e. the greater the current gain of the transistor under test), the outputs of comparators IC1...IC3 (IC4...IC6 for PNP transistors) will go low in turn. The output state of the three comparators is decoded by R15...R19, T1, T2 and T3, such that 'A', 'B', 'C' or 'F' appears on the seven-segment display. 'F' indicates a defective transistor, and is also obtained if no transistor is connected in circuit, or if the pushbutton switch in the base lead of the transistor is pressed (opened). If that is not the case, the transistor has an emitter-collector short.

S3 is used to switch between NPN and PNP types.

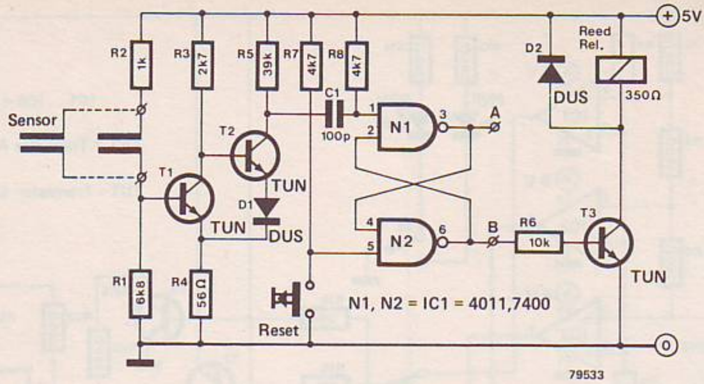
The display is a common-anode type.

R. Storn

128 Moisture sensor

When the circuit shown here detects the presence of moisture, it causes a reed relay to drop out. The relay can be used to disconnect a piece of equipment from its voltage supply, thereby eliminating the possibility of electrical shock.

The original application for the circuit was in an underwater camera which employed an electronic shutter. In the event of ingress of water



into the camera, the shutter circuit was disconnected, thus protecting the photographer from the risk of a high voltage shock. However the circuit can also be used in a variety of other applications, for example a 'leak detector' for boats, or as a 'dry-washing' indicator, etc.

The sensor is formed simply from a pair of copper wires held slightly apart, and the presence of moisture is detected by the resultant drop in the resistance between the wires. When this falls below a certain value, the output of the Schmitt trigger formed by T1 and T2 goes high. The flip-flop formed by N1 and N2 is thus triggered via C1, with the result that T3 is turned off and the relay drops out. The circuit also allows the option of the relay being pulled in when moisture is detected. R6 is simply connected to point A, rather than point B. The circuit is of a sufficiently 'universal' character that in place of the moisture sensor, virtually any alternative type of sensor (LDR, NTC etc.) can be used.

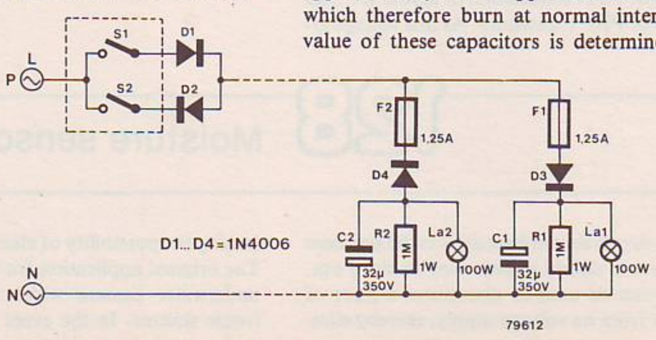
J.M. van Galen

129 2 switches - 2 lamps - 1 wire

When housewiring, the addition of an extra switch and light to an existing circuit using the same power supply point would not normally cause any problems. However, the situation can arise where it is not possible to 'run' an extra cable between the additional switch and light thereby making it impractical to fit them.

The circuit described here is a simple but effective method of solving this problem by replacing the missing wire with a little ingenuity.

It will be seen from figure 1 that diodes D1 and D2 ensure that switch S1 controls lamp La1, whilst S2 controls lamp La2. The half-wave rectified mains voltage is partially smoothed by capacitors C1 and C2, so that an RMS voltage of approximately 240 V appears across the lamps, which therefore burn at normal intensity. The value of these capacitors is determined by the



power rating of the lamps used. The appropriate value can be calculated by using the following equation:

$$C_X = 32 \cdot \frac{P_X}{100}$$

where C_X is the new value of the capacitor (in μF) and P_X the power rating (in W) of the corresponding lamp.

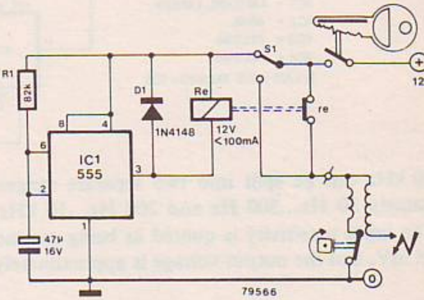
W. Richter

130 Car anti-theft protection

The circuit described here is based on an unusual method of deterring a possible car thief. Shortly after it is started, an engine fault is simulated. Restarting gives the same result suggesting that the car may be more trouble than it is worth.

The actual circuit is extremely simple. A 555 timer provides a delay of roughly 5 seconds. The normally closed contact of the relay is connected in the lead from the ignition switch to the ignition coil. Switch S1, which is used to arm the circuit, should of course be hidden.

When power is supplied to the circuit (via the ignition switch), the relay contact is initially closed and the engine will start. After the delay period provided by the 555 has elapsed, the relay contact is opened and the ignition coil is switched



out of circuit. The delay period can be altered as desired by selecting different values for R1 or C1.

B.H.J. Bennink

131 Frequency multiplier

There may be occasions when it is required to measure low frequencies with a high degree of resolution. The circuit presented here is intended as a frequency multiplier for just this purpose which offers a resolution of 0.1 Hz with a fast measuring time.

A block diagram of the frequency multiplier is shown in figure 1. As can be seen, this configuration bears more than a passing resemblance to the (by now) fairly common PLL frequency synthesiser. However, in this instance it is the division ratio which is fixed and not the input (or reference) frequency. The VCO frequency is divided by 100 and then compared with the input frequency in a phase comparator. The resulting phase difference creates a DC signal which is used to correct the VCO frequency. This means that the VCO output frequency will be exactly 100 times that of the input.

In the circuit diagram of figure 2, the input fre-

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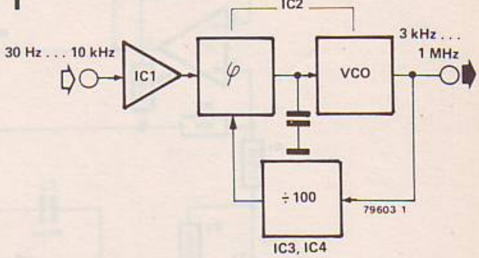
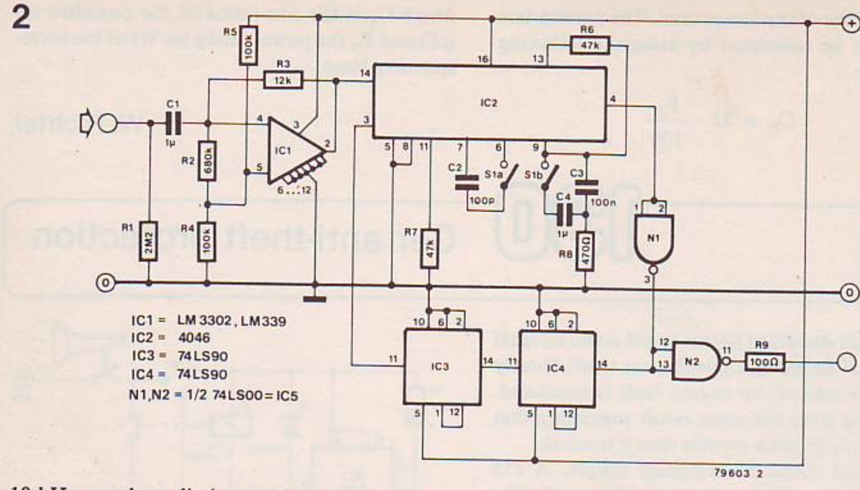


quency is first amplified by IC1 before being fed to the phase locked loop, IC2. The VCO output is divided by 100 by the two decade counters IC3 and IC4 whereupon its phase is compared with that of the input signal in the PLL itself. The VCO output frequency is fed to the meter via the inverter formed by N2. Switch S1 is included so that the overall frequency range of 30 Hz...

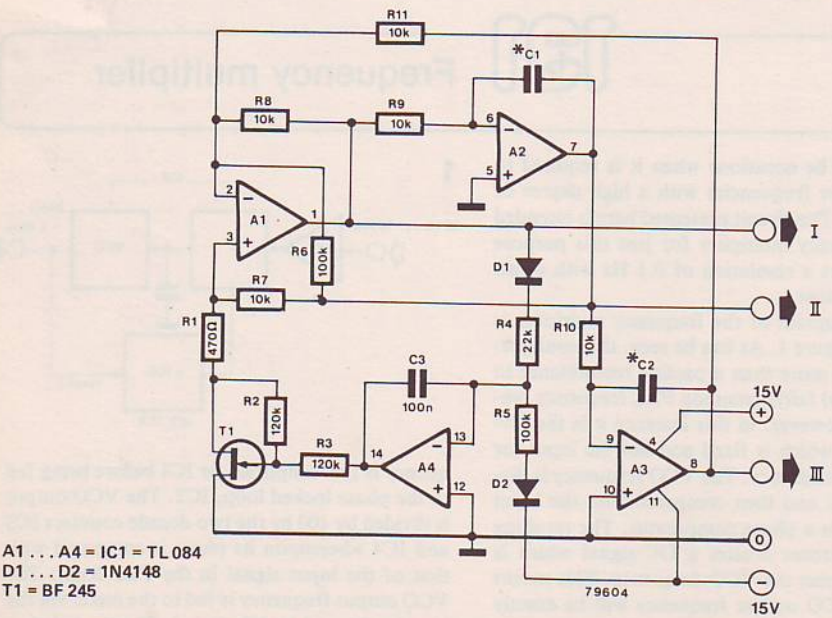
2



10 kHz can be split into two separate ranges, namely 30 Hz...300 Hz and 200 Hz...10 kHz. The input sensitivity is quoted as being around 25 mV, and the output voltage is approximately 4.5 Vp-p. Power supply requirements are 7 - 18 V at around 30 mA.

H. Rol

132 Sinewave oscillator



Under certain conditions if the output of a selective filter is fed back to the input a sinewave oscillator is produced. In itself, the idea is not new, but the way in which it is realised in the circuit shown here is original.
The output of the state variable filter formed by A1...A3, R7...R11, C1 and C2 is fed back (from the output of A2) to the input (left hand side of R7). The amplitude of the output signal is stabilised by the action of FET T1, which in conjunction with R1 forms a voltage-controlled attenuator. The control voltage is derived from the output of A1 via a diode-resistor network and the integrator round A4.
The sinewave signal is available at the outputs of A1, A2 and A3. Since A2 and A3 are connected as integrators, i.e. as lowpass filters, the distortion at output III will be lower than that at output I.
The integrators have unity gain at the resonant frequency of the circuit.
The desired value of C1 and C2 can be calculated by:

$$C1 = C2 = \frac{16}{f}$$

where f is in kilohertz and C is in nanofarad.

G. Schmidt

133 Digital frequency synthesiser

A frequency range of 0.1 Hz to 999.9 kHz, a choice of CMOS or TTL output levels, and an accuracy/stability which is limited only by that of the crystal oscillator – these are the main features of the digital frequency synthesiser shown here.
As can be seen from the block diagram in figure 1, the heart of the circuit is formed by a phase locked loop (PLL). In principle such a PLL circuit can be likened to an op-amp connected with feedback, such that the output voltage of the op-amp varies to keep the voltage at both inputs the same; the PLL circuit varies the *frequency* of the output signal, so that the frequency of both input signals remain the same. If the output frequency is divided by a factor N, and then fed back to one of the PLL inputs, the frequency of the PLL output signal will be exactly N times that of the *other* input signal. Thus all we have to do is ensure that the latter is a stable reference signal, and we have an output whose frequency is equally stable but is N times the reference frequency.
The next step is to provide for the division factor, N, to be made variable, with the result that the frequency of the output signal can also be varied. By including a divide-by-1000 counter, which can be switched in or out of circuit, the frequency range of the output signal can be extended down to as low as 0.1 kHz. Finally, output buffers which amplify the output to both TTL and CMOS levels give the circuit a more

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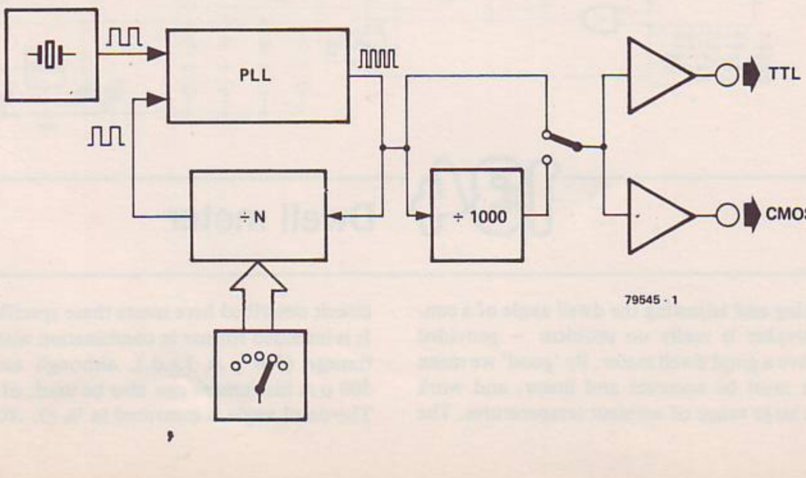

'universal' character.

The complete circuit diagram of the digital frequency synthesiser is shown in figure 2. The reference signal is provided by a 3.2768 MHz crystal which is divided by a factor of 2^{15} (= 32768) by IC5 and IC6, so that a signal whose frequency is exactly 100 Hz is fed to one input of the PLL IC (IC7). The frequency divider for the PLL output is formed by IC8...IC11. The desired division ratio (N), and hence the output frequency, is set up on the decade switches, S3...S6. The output of AND gate N10 provides the other input signal to the PLL, and due to the action of the PLL, the frequency of this signal remains constant at 100 Hz.

The operation of the phase locked loop is dependent upon the value of the capacitor connected between pins 6 and 7 of the IC. Since the output frequency of the PLL can be varied over a fairly wide range, it is necessary to ensure that the capacitor value can also be varied with frequency. This is done via electronic switches ES2 and ES3, which connect either one or two extra capacitors in parallel with C4. Control signals for these switches are derived, via suitable logic gating, from the decade switches, S3...S6.

The divide-by-1000 counter is formed by decade counters IC12...IC14. Depending upon the position of the range switch S1, the figure set up on S3...S6 will be in either Hz or kHz.

The output buffers are formed by means of inverters and a pair of balanced emitter followers. The outputs are short-circuit-proof. An additional electronic switch, ES1, is included to ensure that there is no output signal when the decade switches are set to 000.0. LED D1 lights up when the PLL is locked on, and thus provides a visual indication that the output frequency is correct.

The circuit requires two supply voltages: 15 V unstabilised, and 5 V stabilised. The unstabilised supply can safely be increased slightly. For example, two nine volt batteries connected in series will prove quite suitable.

R. Dürr and D. Hackspiel

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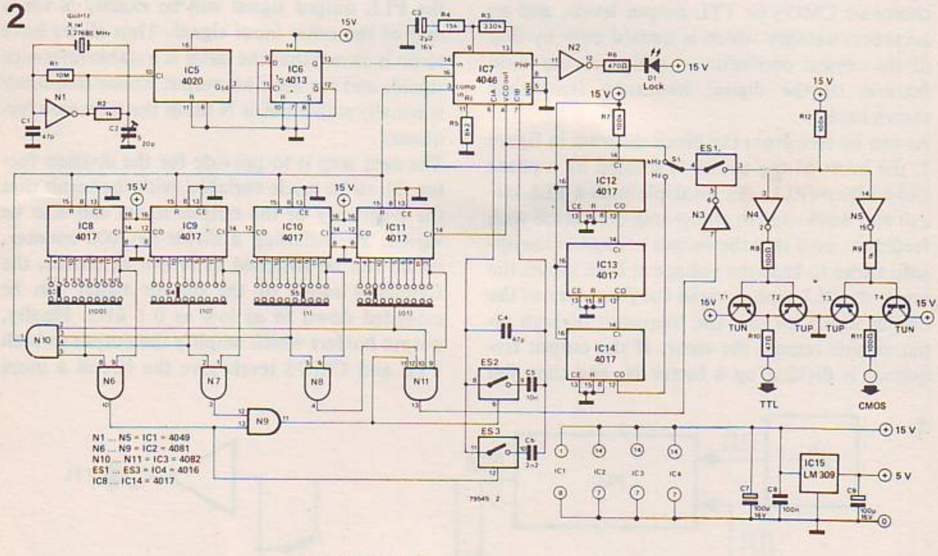
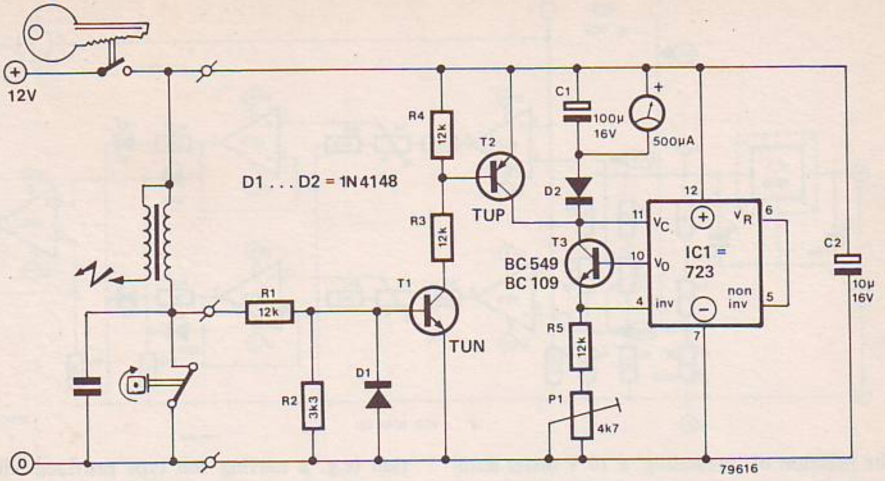


134 Dwell meter

Checking and adjusting the dwell angle of a contact breaker is really no problem – provided you have a good dwell meter. By 'good' we mean that it must be accurate and linear, and work over a large range of ambient temperatures. The circuit described here meets these specifications. It is intended for use in combination with a multimeter (500 μA f.s.d.), although any other 500 μA instrument can also be used, of course. The dwell angle is measured in % (0...100%). If



a reading in degrees is required, the actual reading must be multiplied by 3.6 and divided by the number of cylinders in the engine.

The circuit could hardly be simpler. The most important section is a constant-current source, consisting of T3 and a voltage-regulator IC (IC1). The extremely stable (and temperature-independent) reference voltage provided by the IC at pin 6 is connected to the non-inverting input of a differential amplifier inside the chip; the inverting input is connected to the emitter of T3. The IC will now adjust its output voltage (V_O) so that the emitter of T3 is maintained equal to the reference voltage. The result will be obvious: a constant voltage, independent of temperature, across a fixed resistance (R5 + P1) must produce an equally constant current. Since the base current of T3 is negligible, the collector current is equal to the (highly constant) emitter current. So far, so good.

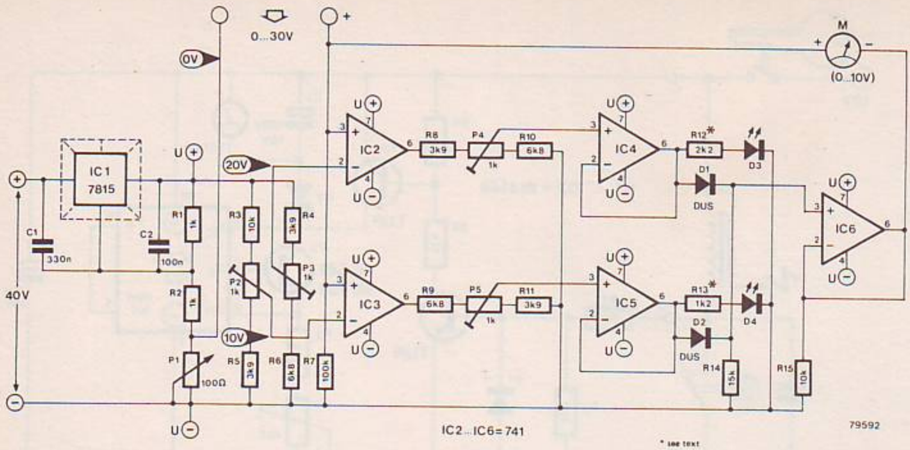
The rest of the circuit either allows this constant current to pass through the meter, or else it doesn't... When the contact breaker is open T1 and T2 will conduct, shorting out the meter circuit. As soon as the points close, T1 and T2 will turn off. The constant current determined by T3 now flows through the meter, charging C1 at the same time. As the points open and close at rapid intervals, an average voltage is developed across C1 and the meter. This voltage is proportional to the 'duty-cycle' of the breaker points: the longer the points remain closed (the larger the dwell angle, in other words), the larger the voltage across C1 will be – giving a correspondingly higher reading on the meter.

The calibration procedure is like the circuit: simplicity itself. After connecting the supply and shorting the input (R1 to supply common), P1 is adjusted for full scale deflection of the meter (100%). After all, a shorted input corresponds to a dwell angle of 100%.

J. Becela

135 Automatic voltage prescaler

If one wants to measure a voltage which is greater than the range (full-scale deflection) of a meter, there are two things which can be done. On the one hand, the input voltage can be reduced to an acceptable value by employing a voltage divider. This is tantamount to 'compressing' the entire range of voltages to be measured. Alternatively we can arrange for the meter scale to cover only a certain portion of the total range of input voltages, depending upon the amplitude of the input signal. For example, with a voltage of 26 V, a 10 V meter will 'look at' the 20 V - 30 V range, and a reading of 6 V will be obtained. The circuit described here performs



the function of 'prescaling' a 10 V meter automatically, and can be used to measure input voltages between 0 and 30 V.

With the aid of IC2 and IC3, the input voltage is compared with a reference voltage of 10 and 20 V respectively. Depending upon which comparator outputs go high, further reference voltages are fed via buffers IC4 and IC5 to diodes D1 and D2. The result is that a voltage which is equal to the greater of the two reference voltages minus the forward voltage drop of the diode, appears on the non-inverting input of IC6. The other diode remains reverse-biased. Since IC6 is connected as a voltage follower, the meter will thus show the difference between the original input voltage and the offset (reference) voltage of either 0, 10 or 20 V. LEDs D3 and D4 provide a visual indication of which scale (0...10 V, 10...20 V or 20...30 V) the meter is switched to. The brightness of the LEDs can be varied as desired by altering the values of R12 and R13.

Any type of meter with a 10 V fullscale deflection (e.g. a moving coil type provided with a suitable series resistor) can be used. However one should bear in mind that the current flowing through the meter forms a load to the remainder of the circuit. Thus the higher the impedance of the meter the better.

P1 is included to compensate for the fact that the op-amps cannot swing fully negative. This potentiometer is best adjusted by shorting the input of the circuit and adjusting the meter for zero deflection. To adjust the remaining potentiometers a 10 and 20 V reference voltage is required. The procedure is as follows: with an input voltage of 10 V, P2 is adjusted such that D4 is just on the point of lighting up. P5 is adjusted such that a zero deflection reading is obtained on the meter when D4 lights up. With a 20 V input, P3 and P4 are then adjusted in a similar fashion.

P. Sieben and J.P. Stevens

136 Bio-control

The growing awareness of the contributory role which stress plays in causing illness has led to increased interest in various forms of 'autogenic' training as a means of promoting relaxation. In particular, different types of 'bio-feedback' circuits have become popular, the idea being that certain physiological functions (heartbeat, body temperature, brain activity) can be monitored and brought under the conscious control of the subject.

The circuit described here operates on the principle of monitoring skin resistance as a measure of how tense the subject is. The same approach is used in so-called lie detectors, however in that case it is the skilled interpretation of the subject's responses to a variety of both innocuous and pointed questions which is important.

The description of the circuit is as follows: vari-

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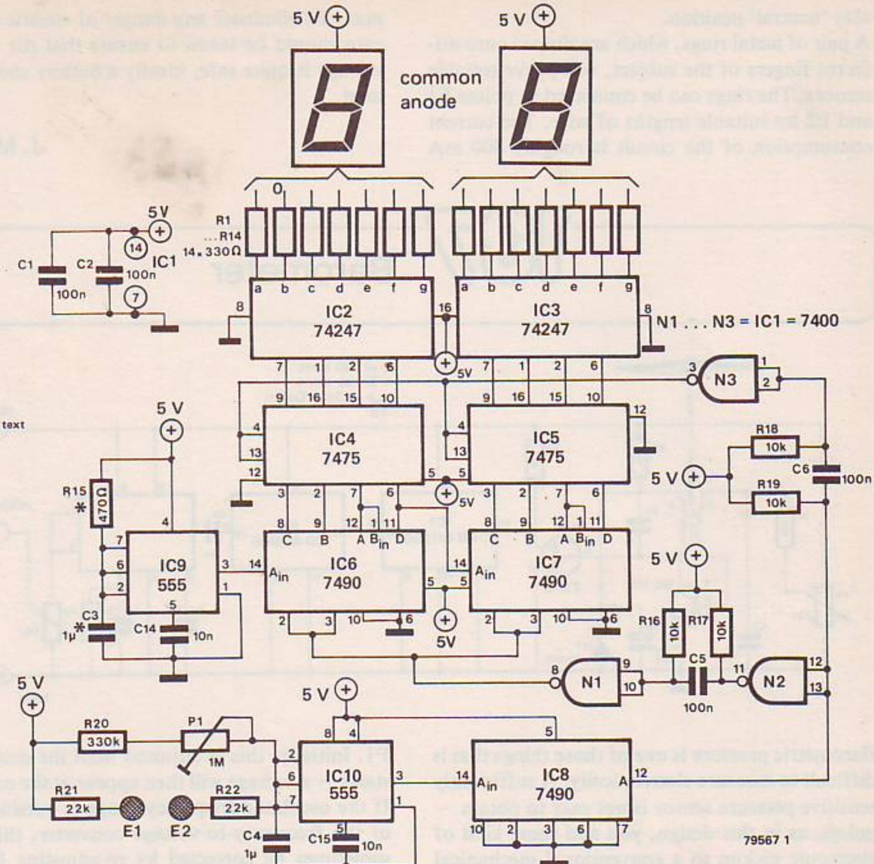

ations in skin resistance (between electrodes E1 and E2) vary the frequency of the oscillator built round a 555 timer (IC10). The output of the oscillator is fed to a 7490 divider (IC8), which in turn controls the reset inputs of the counter formed by IC6 and IC7. The result is that the period between successive pulses from IC10 determines the number of clock pulses fed to this counter from a second oscillator (IC9). The outputs of the counter are decoded and displayed on a pair of 7-segment displays, thereby providing a numerical indication of the subject's relative tenseness.

The frequency of the second oscillator, which is also formed by a 555 timer, is determined by C3 and R15. By incorporating the circuit shown in figure 2, several different clock rates can be chosen, thereby allowing the sensitivity of the circuit to be varied to suit different circumstances. Initially P1 should be adjusted to a suit-

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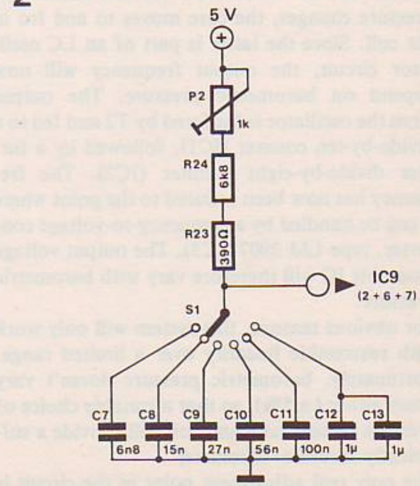
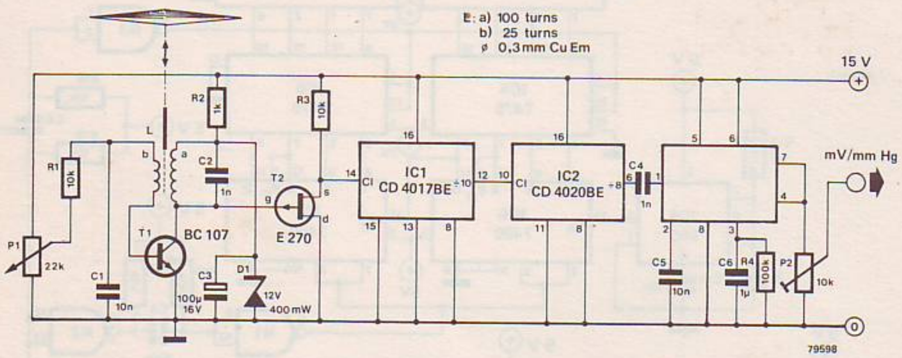


ably 'neutral' position.
A pair of metal rings, which are slipped onto different fingers of the subject, will prove suitable sensors. The rings can be connected to points E1 and E2 by suitable lengths of wire. The current consumption of the circuit is roughly 400 mA max. To eliminate any danger of electric shock, care should be taken to ensure that the supply voltage is quite safe, ideally a battery should be used.

J. Mulke

137 Barometer



Barometric pressure is one of those things that is difficult to measure electronically. A sufficiently sensitive pressure sensor is not easy to obtain – unless, as in this design, you add some kind of electronic pickup to a conventional mechanical barometer.
The ferrite core of a coil is attached to the 'drum' in the barometer. As the barometric pressure changes, the core moves to and fro in the coil. Since the latter is part of an LC oscillator circuit, the output frequency will now depend on barometric pressure. The output from the oscillator is buffered by T2 and fed to a divide-by-ten counter (IC1), followed by a further divide-by-eight counter (IC2). The frequency has now been reduced to the point where it can be handled by a frequency-to-voltage converter, type LM 2907 (IC3). The output voltage from this IC will therefore vary with barometric pressure.
For obvious reasons, this system will only work with reasonable linearity over a limited range. Fortunately, barometric pressure doesn't vary much either (± 5%), so that a suitable choice of pressure sensor, core and coil will provide a sufficiently accurate 'barometer'.
The only real adjustment point in the circuit is P1. Initially, this is adjusted until the oscillator starts – a voltage will then appear at the output. If the oscillator frequency range is outside that of the frequency-to-voltage converter, this can sometimes be corrected by re-adjusting P1. If the frequency is too far off, however, the value of C2 will have to be changed.
The preset at the output (P2) is used to adjust the output level as required. A digital or analogue millivoltmeter can be connected at this point.

Y. Nijssen

Editorial note:
At first sight, there doesn't seem to be much point in stripping an existing barometer in order to connect an electrical pointer instrument instead of the mechanical pointer. However, having an electrical voltage available that is proportional to barometric pressure opens a whole range of possibilities. Just to name one: What about designing a home weather-forecasting computer?

138 Electronic weathercock

The disadvantage of most wind direction meters is the need for complicated mechanical drive systems which are necessarily prone to wear. The unit described here is intended to offer a solution to this problem. A disc which contains a number of slots is attached to the spindle of the weather vane. A light source is mounted above the disc, and a row of 3 light dependent resistors (LDRs) is situated below the disc – as shown in figure 1. Which LDRs are illuminated will depend upon the position of the disc, and therefore upon the direction of the wind. If the slots in the disc are correctly positioned (see figure 2), the information from the LDRs can be coded into BCD format, such that each of the eight main compass points will correspond to a particular BCD code. By means of a BCD-decimal converter, the resultant information, and hence the direction of the wind, can then be displayed on a circle of LEDs. The 'electronics' of the unit are shown in the circuit diagram of figure 3. If no light falls upon an LDR, the associated transistor is turned off and the input of IC1 is pulled down to 0 V via the 470 Ω resistor. As soon as sufficient

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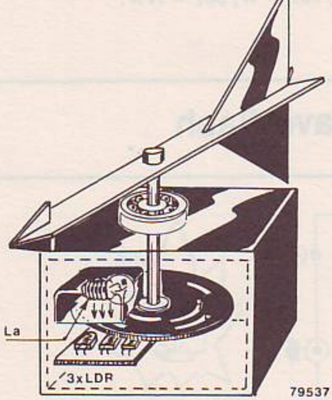

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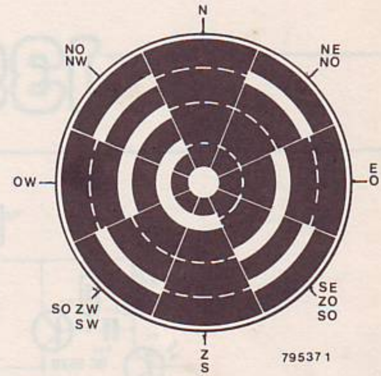

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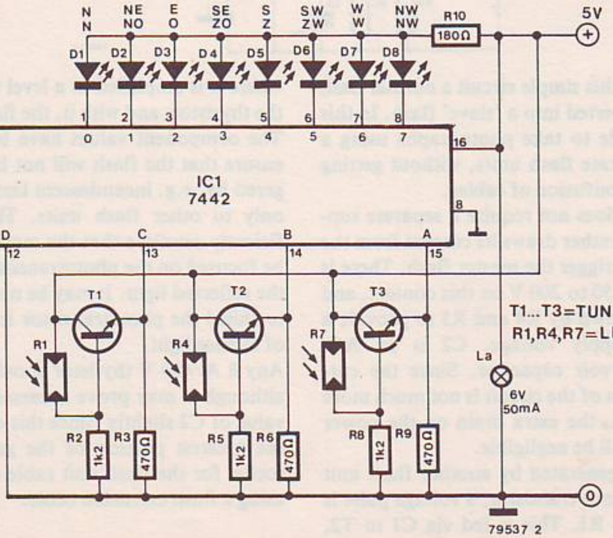


Table 1.

A	B	C	D	wind direction	LED
0	0	0	0	north	D1
1	0	0	0	north-east	D2
0	1	0	0	east	D3
1	1	0	0	south-east	D4
0	0	1	0	south	D5
1	0	1	0	south-west	D6
0	1	1	0	west	D7
1	1	1	0	north-west	D8

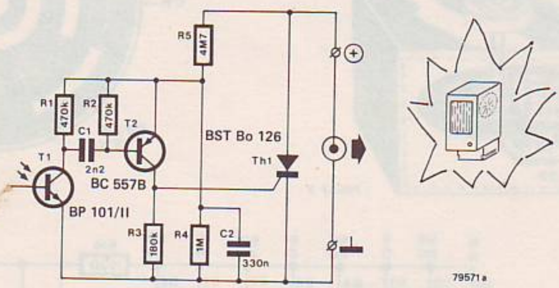
light falls on an LDR to turn the transistor on, however, the corresponding input of IC1 is taken high. Thus the state of the three LDRs is translated into a BCD code applied to the inputs of the 7442. Depending upon the combination of logic '0' s and '1' s at the four inputs, the 7442 takes one of its outputs low, with the result that the corresponding LED lights. The BCD code for each of the eight compass points is listed in table 1.

D. Maurer

Editorial note:

If the disc should come to rest exactly between two compass points, e.g. between south and south-east, then the effect of stray light may cause the wrong code to be presented to IC1, and the NW LED will light up. Therefore, it would be better to use the 'Gray code', i.e. 000 = N; 100 = NE; 110 = E; 010 = SE; 011 = S; 111 = SW; 101 = W; 001 = NW.

139 Slave flash



With the aid of this simple circuit a normal flash unit can be converted into a 'slave' flash. In this way it is possible to take photographs using a number of separate flash units, without getting tangled up in a confusion of cables.

The slave flash does not require a separate supply voltage, but rather draws its current from the contact used to trigger the master flash. There is normally some 150 to 200 V on this contact, and this is divided down by R4 and R5 to provide a suitable low supply voltage. C2 is an AC-decoupling/reservoir capacitor. Since the current consumption of the circuit is not much more than several μ A, the extra drain on the power supply battery will be negligible.

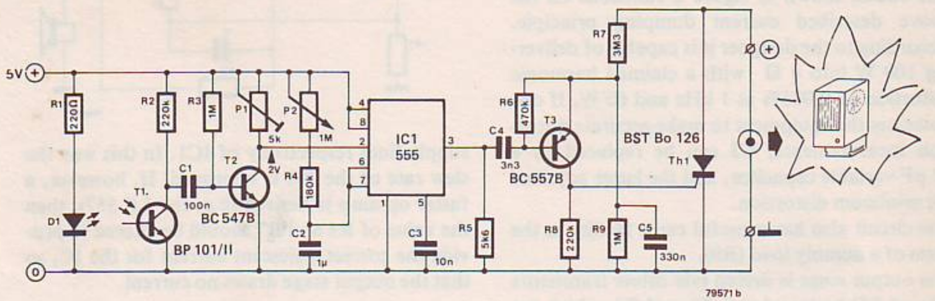
When the light generated by another flash unit falls upon the photo-transistor, a voltage pulse is generated across R1. This is fed via C1 to T2, where it is amplified to a level suitable to trigger the thyristor, and with it, the flash.

The component values have been calculated to ensure that the flash will not be spuriously triggered by, e.g. incandescent lamps, but will react only to other flash units. The circuit is sufficiently sensitive that the master flash need not be focused on the phototransistor; it will react to the reflected light. It may be necessary, however, to shield the phototransistor from other sources of intense light.

Any 8 A/400 V thyristor should prove suitable, although it may prove necessary to increase the value of C2 slightly (since this capacitor supplies the greatest portion of the gate current). The socket for the flash unit cable can best be made using a flash extension cable.

F. Schäffler

140 Photo-flash delay



One of the more specialised areas of photography is the use of ultra-short exposure times to capture events occuring at high speed. Everyone will have seen the results of this technique at one time or another: a light bulb in the process of disintegrating under the impact of a hammer, or, as in the picture shown here, a splash of water. Photographs of this type can be taken fairly simply by employing an 'open-lens' approach, i.e. the photo is taken in a darkened room and the lens of the camera is opened before the subject is illuminated. The lighting is provided by a high-speed (electronic) flash unit capable of providing extremely short exposure times.

One problem with this method is determining the exact moment at which the flash gun should be triggered. Because of the extremely short time intervals involved, this can really only be done electronically. In the case of the picture shown here, the drop of water was sensed by a photoelectric cell, which, with the aid of the following circuit, provided a predetermined delay before triggering the flash.

An LED (D1) and phototransistor T1 are used to form the light gate. When the light from the LED is interrupted, there is a sharp rise in voltage across R2. This is fed via T2 to the trigger input of the 555 timer (IC1). When the delay period provided by the timer has elapsed, a negative-going pulse appears at the output of this IC (i.e. pin 3), with the result that T3 and the thyristor are turned on, and the flash is triggered.

Any 0.8 A/400 V thyristor will prove suitable, however it may be necessary to increase the value of C5 slightly. The DC bias voltage on the collector of T2 should be adjusted to 2 V by means of P1.

With the aid of P2, the delay provided by the circuit can be varied between approximately 0.25 and 1.3 s. By altering several component values the range of possible delays can also be varied. The delay time is given by 1.1 x R x C2, where R is the series connection of P2 and R4. The minimum permissible value for R2 is 1 k. As one might expect, the light gate is the section of the circuit which will present the most difficulty when it comes to construction. Whatever arrangement is chosen will depend largely on individual circumstances, however the sensitivity of the circuit is greatest when the LED and phototransistor are mounted as close together as possible. Care should be taken to ensure that light from the LED cannot reach the lens of the camera.

F. Schäffler

141 Current dumping amplifier

The circuit exploits the fact that, due to the effect of the four passive components, R2, R3, L and C shown in figure 1, the non-linear characteristic of the output stage becomes unimportant. Thus it is possible to use a Class-B output stage (i.e. the output transistors are biased to their cut-off points so that there is no quiescent output current) with all the advantages and none of the disadvantages (crossover distortion) of that configuration.

The circuit shown in figure 2 functions on the above described current dumping principle. According to the designer it is capable of delivering 100 W into 4 Ω with a claimed harmonic distortion of 0.006% at 1 kHz and 60 W. If one possesses the equipment to make accurate distortion measurements, C3 can be replaced by a 22 pF variable capacitor, and the latter adjusted for minimum distortion.

The circuit also has a useful extra facility in the form of a dummy load (R9).

The output stage is driven (via driver transistors T2 and T5) by transistors T1 and T4, which are connected in series with the positive and negative supply lines respectively of IC1. In this way the slew rate of the 741 is improved. If, however, a faster op-amp is desired (e.g. the LF 357), then the value of R4 and R7 should be altered to provide the correct quiescent current for the IC, so that the output stage draws no current.

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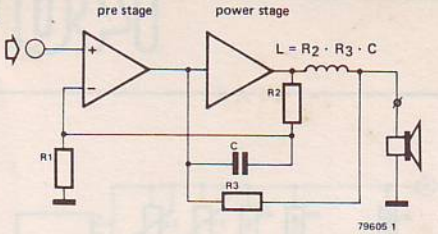

G. Schmidt

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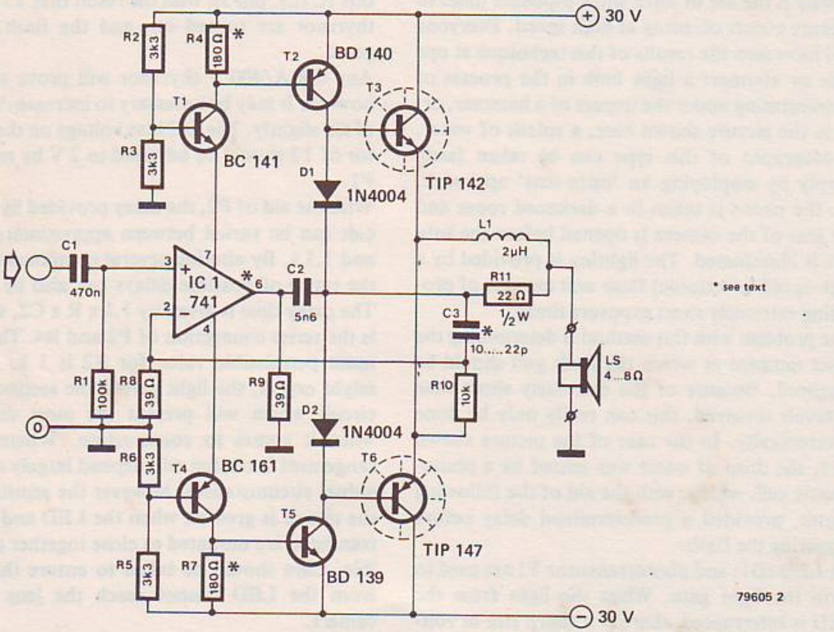

142 Doorbell drone

There seems to be no end to the variety of different sounding doorbells which people are prepared to design. Everything from the Hallelujah Chorus to the chimes of Big Ben have been simulated for the entertainment of visiting door-to-door salesmen.

The circuit presented here produces a sound which is somewhat akin so that of bagpipes, and while not exactly signalling the death knell of original bagpipes, should prove popular north of the border.

The circuit is also intended to foil ill-mannered

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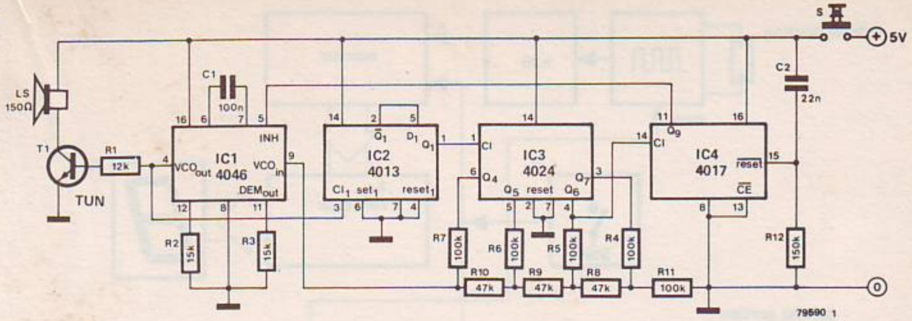

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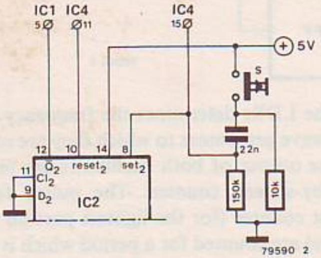

visitors who insist on pressing the doorbell for an annoying long time, since the bell automatically cuts out after approximately two seconds.
As can be seen from the circuit diagram in figure 1, very little in the way of components is required to build this 'exclusive' doorbell. A 4046 phase locked loop IC (IC1) is used as a voltage controlled oscillator with a nominal frequency of around 800 Hz determined by the values of R3 and C1. The actual frequency of the oscillator is controlled by feeding the output signal to one half of a dual flip-flop (IC2) which is connected as a divide-by-two counter, and then to a binary ripple counter. The ladder network of resistors R4...R11 provides a staircase voltage, which is fed back to the control voltage input (pin 9) of IC1, thereby producing the 'bagpipe' effect. At the end of the count cycle IC4 takes the inhibit input of IC1 high, thus ensuring that the 'bagpipes' do not continue to sound if the bellpush is held down. R12 and C2 automatically reset IC4 the next time the bellpush is depressed.

S. Halom

Editorial Note:
Although the original circuit as shown in figure 1 will prove an effective remedy against over enthusiastic bell-pushers, unfortunately it does not take into account what will happen if the push-button switch (S) is only depressed for a brief moment. Since releasing the switch interrupts the supply voltage to the circuit, the bagpipes will be cut off in their prime! To forestall a flood of letters from incensed Scotsmen we include the following possible modifications. As CMOS ICs draw very little current they can be provided with a continuous supply voltage. By using the other flip-flop in IC2, the circuit can be modified to ensure that the entire 'melody' will be heard even if the bellpush is only depressed momentarily.
The circuit of figure 1 should be altered as follows:
- *switch S is replaced by a link*
- *C2 and R12 are omitted*
- *the connection between pin 11 of IC4 and pin 5 of IC1 is broken*

The circuit should then be connected as shown in figure 2.

143 Digital contrast meter

When enlarging photographs, two factors are of prime importance, the required exposure time, which is determined by the density of the negative, and the contrast of the negative. The latter determines which grade of paper should be used in order to obtain a print with good overall tonal contrast. The contrast of the negative is basically the difference between the lightest and darkest portions of the exposed film. If we take the second log of this difference, we obtain the contrast

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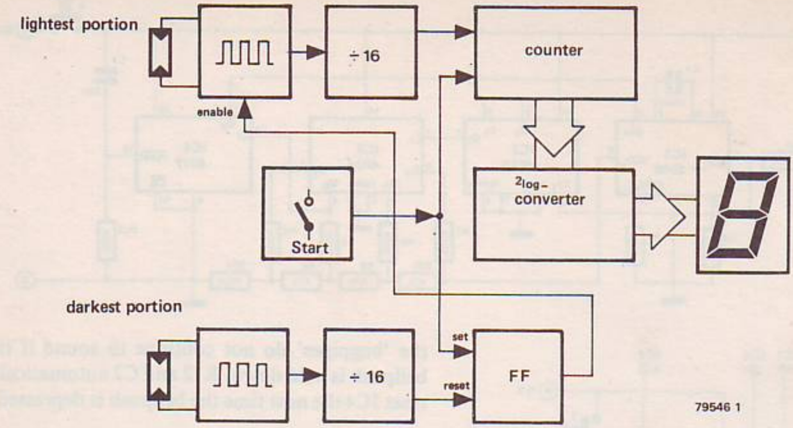

ratio of the negative. Thus, for example, if the lightest part of the negative lets through 8 times as much light as the darkest part, the contrast ratio of the negative will be 3 ($2^3 = 8$). The circuit employs two light dependent resistors as sensors, and displays the contrast of the negative directly on a seven-segment display. The operation of the circuit is illustrated in the block diagram of figure 1. The amount of light falling upon the LDRs determines the frequency of the squarewave generators to which they are connected. The output of both oscillators are fed to a divide-by-sixteen counter. The pulses for the topmost counter (for the lightest portion of the negative) are counted for a period which is determined by the frequency of the signal from the lower counter. The result is that the value stored in the subsequent binary counter represents the

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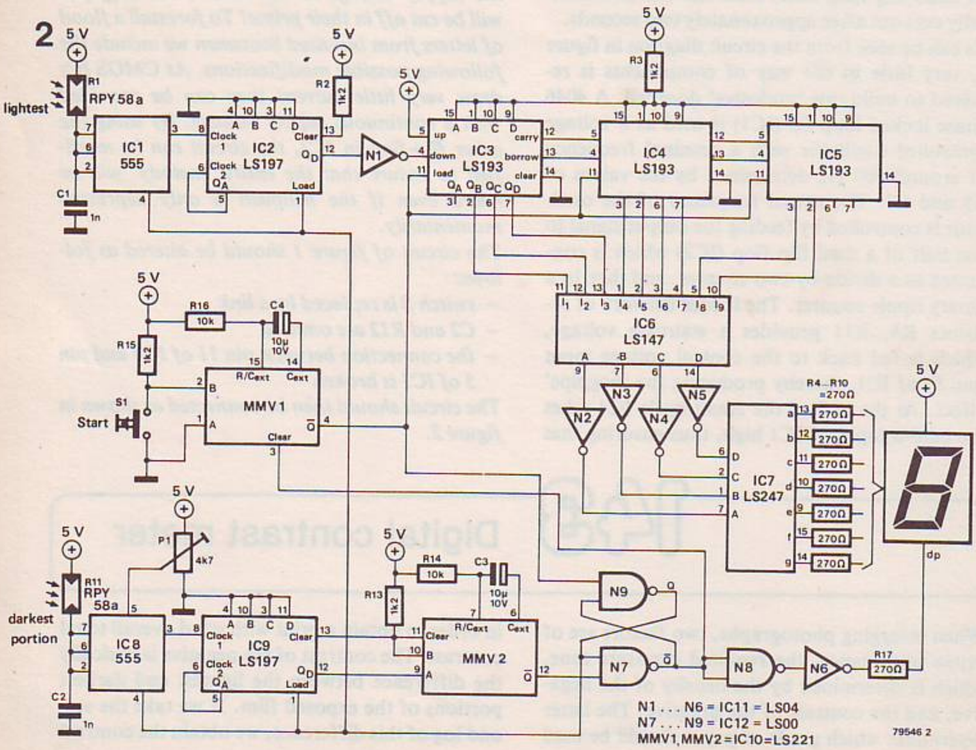


ratio of the two clock generator frequencies, and hence the ratio of the lightest and darkest portions of the negative. The contents of the 'ratio' counter, are then fed to the $\log_2$ converter, the output of which is decoded and displayed. A measurement cycle is initiated by closing the start switch, which sets the flip-flop and resets the counter.

The complete circuit diagram of the contrast meter is shown in figure 2. With the exception of the two clock oscillators, in which 555 timers are used, the circuit is low power Schottky TTL. IC2 and IC9 are the divide-by-16 counters, whilst the binary 'ratio' counter consists of IC3, IC4 and IC5. This counter uses negative logic, i.e. it begins with all outputs high, and then counts down. Thus at the start of each measurement cycle, a 'load' pulse, and not as one might expect, a 'reset' pulse, is applied to the counter. The paralleled data inputs of the counter are all held high, and the pulses to be counted are fed to the 'down' input. The reason for adopting this arrangement is the $\log_2$ converter, which also uses negative logic. This converter is formed by IC6, a decimal-BCD priority encoder. This IC recognises the highest order bit in the input signal which is active, i.e. logic 0, and outputs the BCD equivalent of that bit's 'weight'. For example, assume the counter holds the binary code for the number 8 (base 10). All bits will be logic 1, with the exception of I3 (remember, we are working with negative logic). IC6 recognises that the highest order bit which is logic 0, is the third bit, therefore it outputs the BCD code for 3. As we have already established, 3 is $\log_2$ of 8, thus the conversion is complete.

The divide-by-sixteen counter, IC9, is followed by a monostable, which provides a reset pulse to the flip-flop formed by N6 and N7 at the end of each measurement cycle. The set pulse is provided by a second monostable, which is triggered by the start switch, S1. The decimal point on the seven-segment display is lit during each measurement cycle.

Since IC6 uses negative logic, its output signals must be inverted (by N2...N5) before they can be fed to the BCD-seven-segment decoder, IC7. The display is a common-anode type, e.g. HP 5082-7750, FND 557. Any type of LDR which is intended for measurement purposes, as opposed to switching applications, can be used. The type named in the circuit diagram is particularly suitable.

The circuit should be adjusted such that, as far as possible, the frequency of the two 555 oscillators is the same when identical amounts of light are falling on both LDRs.

The LDRs should therefore be laid upon a surface which is evenly illuminated. Coarse adjustment is performed by varying the values of C1 and C2, whilst fine adjustment – which in many cases will be all that is required – is carried out with the aid of P1.

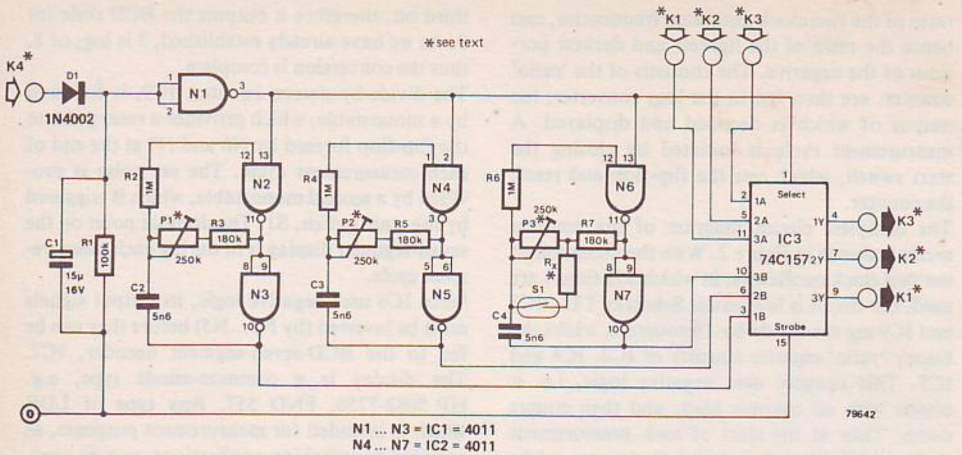
J. van Dijk

144 Emergency flight controller

When flying radio controlled model aeroplanes there is always the chance that a fault will occur in either the transmitter or the receiver, and that the plane will no longer obey the control signals. If one is fortunate, the model will fall near the operator, however it may equally well happen that the plane will remain in flight for a considerable distance, and that the last the unfortunate owner will see of his model is it disappearing over the horizon! The circuit described here is designed to prevent the latter possibility, and also attempts to lessen the severity of the crash, by ensuring that the model will assume a glide trajectory.

The circuit reacts to a loss of output from the receiver. When both transmitter and receiver are functioning normally, the position of the servos is determined by the transmitted control pulses. Depending upon the make of servo, a pulse width of 1.5 ms corresponds to the neutral position, whilst pulse widths of 1 and 2 ms correspond to the extreme positions of the servo. When the stream of control pulses is interrupted, the three multivibrators in the circuit set the servos to a predetermined position.

Input K4 is connected to the output of the receiver. Inputs K1, K2 and K3 are connected to the receiver servo control outputs for the elevator, rudder and engine throttle respectively, whilst outputs K1, K2 and K3 are connected to



the corresponding servos. As long as control pulses are received via K4, the multiplexer, IC3, ensures that inputs K1, K2 and K3 are connected to the corresponding outputs (and servos). However when the control pulses are interrupted, the multiplexer switches to the outputs of the three oscillators. The position of P1, P2 and P3 then determine the position of the servo control horns. A mercury switch is connected across P3 (elevator control). The switch should be mounted such that it will close when the angle of descent is greater than 10°, whereupon the position of the elevator servo will be determined by the value of R_x (10...200 k).

W. van Staeyen

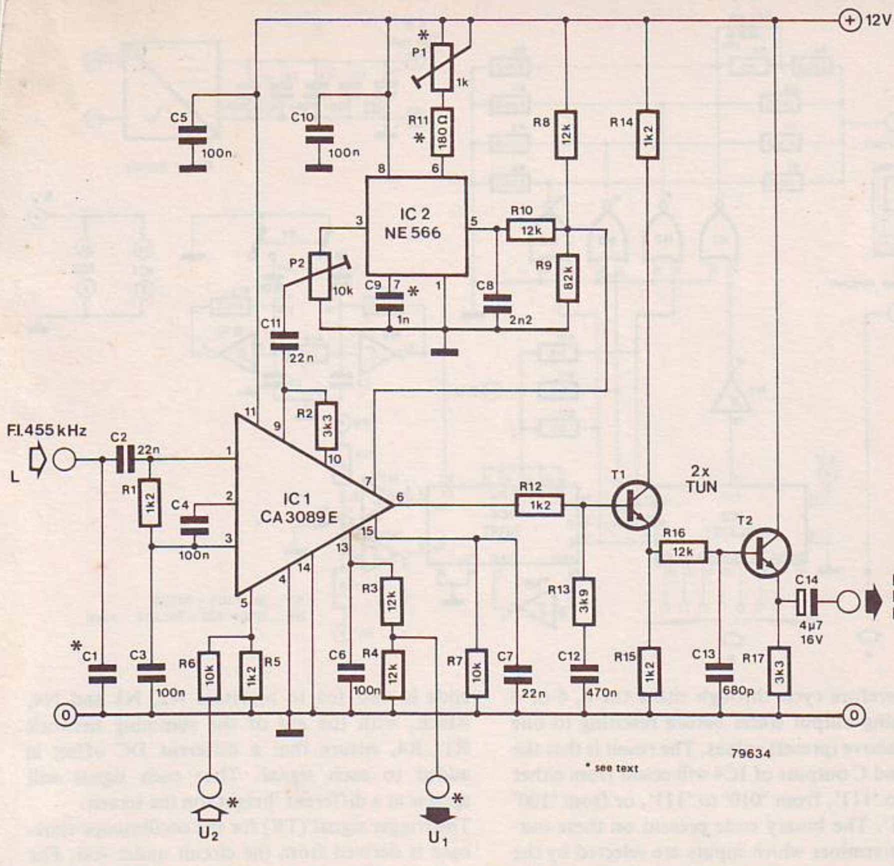
145 FM PLL using CA 3089

The following circuit should prove particularly interesting to those readers considering building their own FM tuner. The novel feature of the circuit is that the well-known CA 3089 IC is not used as a conventional IF amplifier/demodulator, but as part of a phase locked loop. The resulting circuit is slightly more expensive and complicated than the 'standard' amplifier/demodulator circuits, however the results obtained are a significant improvement on those of a 'classical' CA 3089 IF strip.

The circuit is intended as an IF amplifier/demodulator for a double conversion tuner operating at an intermediate frequency of 455 kHz. When using a PLL circuit for FM demodulation the S/N ratio of the demodulated signal is proportional to the ratio of frequency deviation/IF frequency, hence the PLL demodulator should operate at the lower IF of 455 kHz.

Briefly, the circuit functions as follows: The IF input signal is first fed to C1, which removes any high frequency signal components which might affect the operation of the PLL. The exact value of C1 will depend upon the mixer circuit which converts the 10.7 MHz output of the front end to the desired IF frequency of 455 kHz. If the input signal is sufficiently 'clean' then lowpass filtering can be omitted. The CA 3089 amplifies and limits the IF signal to approximately 300 mV, whereupon it is fed to the on-chip quadrature detector. Output voltage U_1 can be used to drive a signal strength meter. The control voltage for the PLL VCO is taken from the AFC output (pin 7) of the IC. The voltage divider formed by R8/R9 is required to set the correct DC bias on pin 5 of IC2. Lowpass filtering of the control signal is provided by R10 and C8.

Largely for reasons of good linearity and high stability, a modern integrated circuit VCO, the NE566, was chosen. However the stability of the VCO is also significantly influenced by the associated frequency-determining components. P1 should preferably be a cermet trimmer, whilst a metal oxide resistor should be used for R11



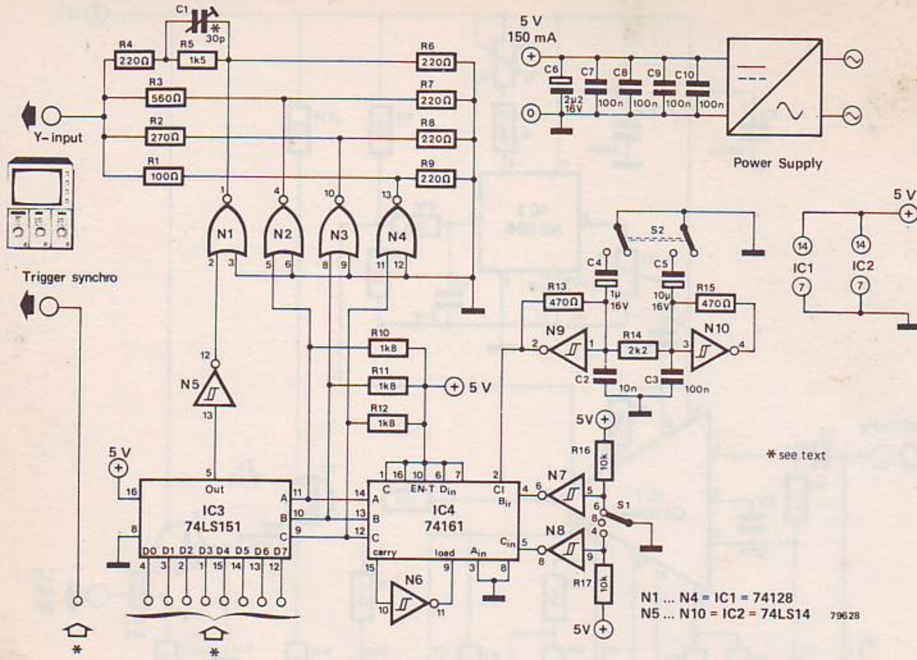
and a ceramic disc capacitor with extremely low temperature coefficient should be used for C9. With the aid of a variable voltage divider (P2), the squarewave output voltage at pin 3 of IC2 (approximately 5.4 V) is reduced to roughly 0.3 V and then fed back via C11, to the input of the quadrature detector of IC1.
The demodulated output signal is available at pin 6 of the CA 3089. If a squelch (muting) facility is required, this can be realised by feeding a positive control voltage to pin 5 of IC1, thereby suppressing the audio output. Finally, the audio signal is lowpass filtered and this output can be used with virtually any stereo decoder.

J. Deboy

146 Logic analyser

Although the name 'logic analyser' is generally understood to denote a quite different type of circuit, there are good reasons for borrowing the title to describe the electronic switch presented here. The switch is intended to simultaneously display the logic state of a number of test points in a digital circuit on an oscilloscope screen.

The circuit functions as follows: The oscillator formed by N9/N10 generates a clock signal with a frequency of either 1 kHz or 100 kHz, depending upon the position of switch S2. This signal is fed to a counter, IC4, which, depending upon the position of switch S1, can be preset to either '1000' (8), '1010' (10) or '1100' (12). The counter



will therefore cycle through either the 8, 6 or 4 remaining output states before resetting to one of the above (preset) values. The result is that the A, B and C outputs of IC4 will count from either '000' to '111', from '010' to '111', or from '100' to '111'. The binary code present on these outputs determines which inputs are selected by the multiplexer, IC3. The multiplexer scans the inputs in turn, and transfers the input signal to the output. Depending upon the position of S1, therefore, 4, 6 or all 8 inputs will be scanned. To ensure that each input signal is displayed 'separately' on the screen, the corresponding binary code is also fed to inverters N2, N3 and N4, which, with the aid of the summing network R1...R4, ensure that a different DC offset is added to each signal. Thus each signal will appear at a different 'height' on the screen.

The trigger signal (TR) for the oscilloscope timebase is derived from the circuit under test. For slower scopes in particular, variable capacitor C1 can be adjusted to obtain optimal picture quality.

P.C. Demmer

147 Battery monitor

Only three LEDs are used to give an indication of the car (or boat) battery condition. The LEDs light as follows:

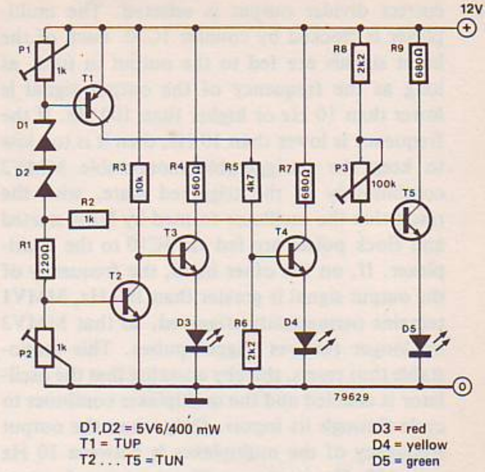
D3	< 12 V
D3 + D4	12...13 V
D4	13...14 V
D4 + D5	> 14 V

Preset P2 sets the voltage above which D3 goes out (13 V); P1 sets the point at which D4 lights

(12 V); finally, P1 sets the voltage above which D5 lights (14 V). The calibration procedure is rather critical, and will have to be repeated several times since the various adjustments affect each other.

The photo shows the author's prototype. All components are mounted in a small plastic tube, with the LEDs at one end and a 'cigarette lighter' plug at the other. The unit can then easily be plugged into the corresponding socket on the dashboard for a quick check of battery condition. If the suggested colours are used for the various LEDs, red will correspond to 'battery low'; yellow (with or without red) indicates 'battery normal'; and green will normally light when the battery is 'on charge'.

S. Jacobsson

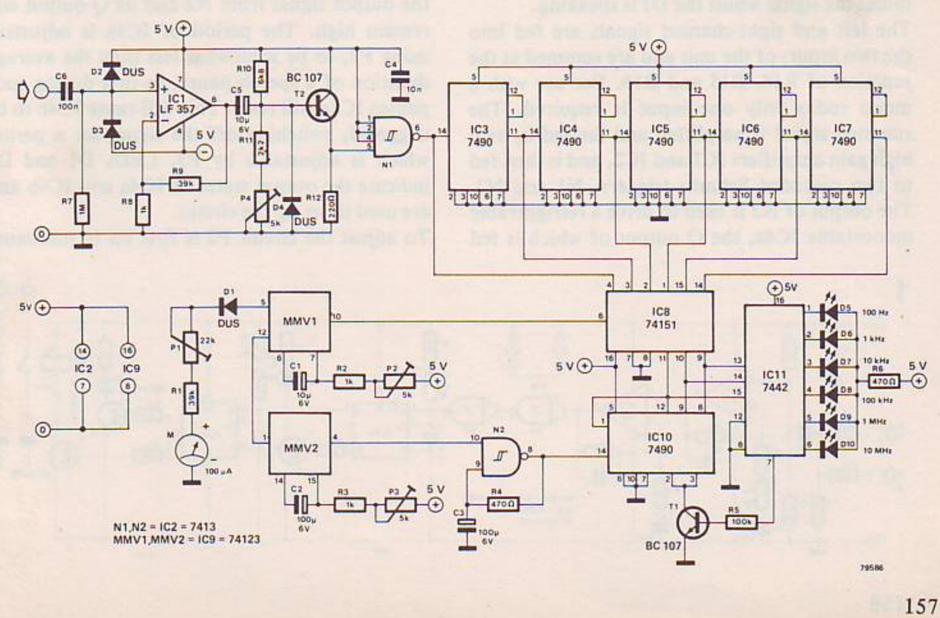


148 Analogue frequency meter

This circuit for a frequency meter offers six ranges: 100 Hz, 1 kHz, 10 kHz, 100 kHz, 1 MHz and 10 MHz. Switching between ranges is performed automatically, and the display is analogue.

The input signal is first amplified to TTL level by means of IC1, T2 and N1, whereupon it is fed to a series of a decade dividers (IC3...IC7). Thus a signal with a frequency between 10 Hz and 100 Hz can be obtained either at the output of N1, or at the output of one of the decade counters. The analogue section of the circuit (MMV1, the moving coil meter and associated components) is designed to produce a full-scale meter deflection for an input signal of 100 Hz.

A multiplexer, IC8, is used to ensure that the



correct divider output is selected. The multiplexer is clocked by counter IC10. Each of the input signals are fed to the output in turn, as long as the frequency of the output signal is lower than 10 Hz or higher than 100 Hz. If the frequency is lower than 10 Hz, then it is too low to keep the retriggerable monostable MMV2 continuously in the triggered state, with the result that the oscillator formed by N2 is started and clock pulses are fed via IC10 to the multiplexer. If, on the other hand, the frequency of the output signal is greater than 100 Hz, MMV1 remains permanently triggered, so that MMV2 no longer receives trigger pulses. This monostable thus resets, thereby ensuring that the oscillator is enabled and the multiplexer continues to cycle through its inputs. Only when the output frequency of the multiplexer is between 10 Hz and 100 Hz is the oscillator stopped, since MMV1 is not triggered sufficiently often to keep MMV2 in the triggered state. The result of stopping the oscillator is that the multiplexer in turn stops at the input which provided the signal of the appropriate frequency. LEDs D5...D10 provide an indication of the range selected.

To calibrate the meter, P3 and P4 should initially be set to the midposition, whilst P1 and P2 are adjusted for maximum and minimum resistance respectively. A 100 Hz signal (with an amplitude of greater than 1 V) is fed to the input of the circuit and P3 adjusted such that the multiplexer begins to cycle through its inputs. This can be verified by checking that the LEDs light up in turn. P2 is now adjusted until the 100 Hz range LED (D5) lights up. P1 is then adjusted for full-scale deflection on the meter. Finally, the circuit can be adjusted for maximum input sensitivity (approximately 10 mV) by means of P4.

H. Bichler

149 D.J. killer

It is possible to distinguish speech from music by virtue of the fact that distinct pauses occur in speech, whereas music is more or less continuous. The DJ killer detects these pauses and mutes the signal whilst the DJ is speaking.

The left and right-channel signals are fed into the two inputs of the unit and are summed at the junction of R14, R15 and R16. For use with a mono radio only one input is required. The summed signal is amplified and limited by two high gain amplifiers IC1 and IC2, and is then fed to two cascaded Schmitt triggers, N1 and N2. The output of N2 is used to drive a retriggerable monostable IC4a, the Q output of which is fed to the input of a second retriggerable monostable IC4b.

So long as a continuous signal is present at the input IC4a will be continuously retriggered by the output signal from N2 and its Q output will remain high. The period of IC4a is adjusted, using P2, to be somewhat less than the average duration of a speech pause, so that during such pauses IC4a will reset. This will cause IC4b to be triggered, switching off the signal for a period which is adjustable by P3. LEDs D1 and D2 indicate the output states of IC4a and IC4b and are used to set up the circuit.

To adjust the circuit P2 is first set to minimum

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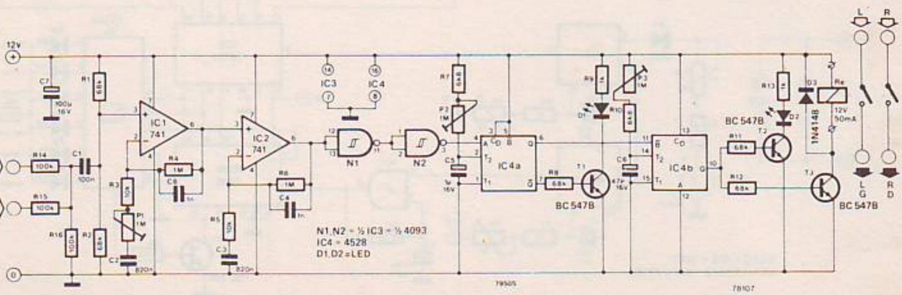

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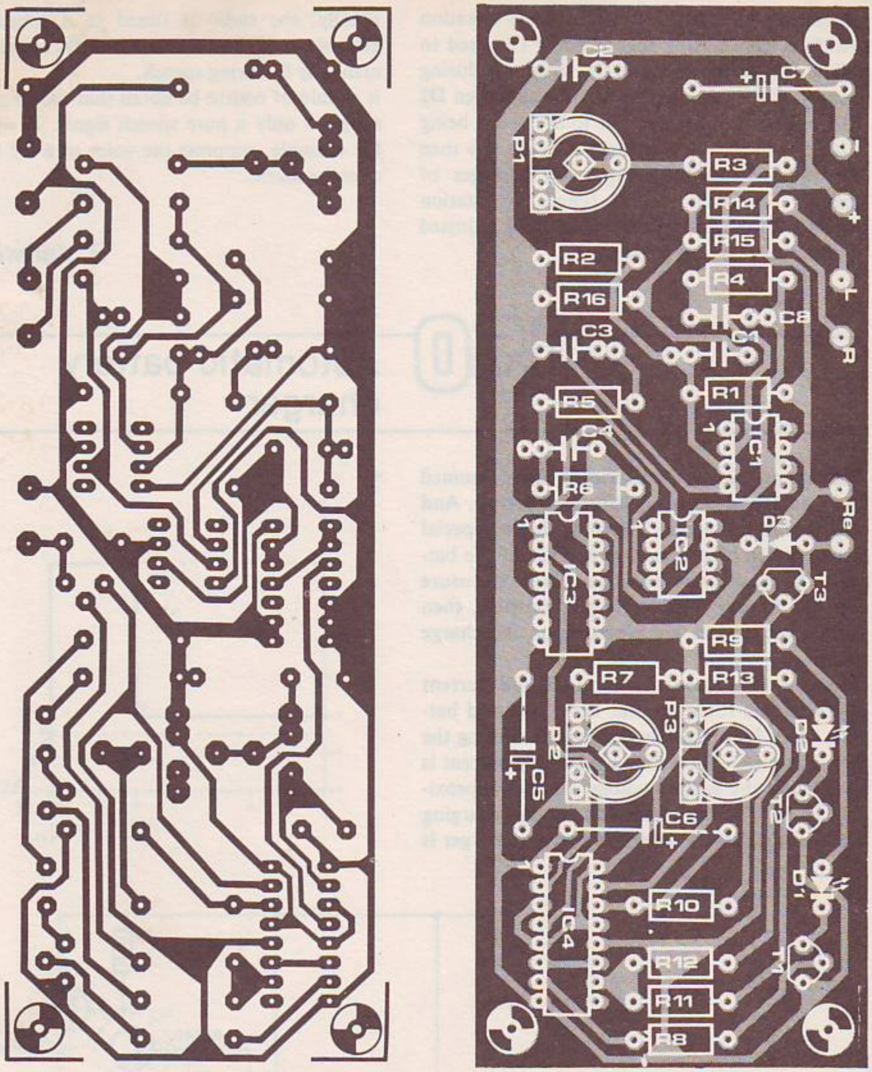

Parts list:

Resistors:
R1, R2, R8, R11, R12 = 68 k
R3, R5 = 10 k
R4, R6 = 1 M
R7, R10 = 6k8
R9, R13 = 1 k
R14, R15, R16 = 100 k
P1, P2, P3 = 1 M

Capacitors:
C1 = 100 n
C2, C3 = 820 n
C4, C8 = 1 n
C5 = 1 µ /16 V
C6 = 47 µ /16 V
C7 = 100 µ /16 V

Semiconductors:
D1, D2 = LED
D3 = 1N4148
IC1, IC2 = 741
IC3 = N1, N2... = 4093
IC4 = 4528
T1, T2, T3 = BC 547B

Miscellaneous:
relay 12 V/50 mA

resistance. The radio is then tuned to a station which is transmitting speech and P1 is used to adjust the sensitivity until D1 goes out during pauses. If the sensitivity is set too high then D1 will stay on continuously due to the circuit being triggered by noise, whereas if it is too low then D1 will extinguish during quiet passages of speech. The radio is then tuned to a station which is broadcasting music and P2 is adjusted until D1 stays on continuously.

Finally, the radio is tuned to a speech programme and P3 is adjusted until D2 remains permanently lit during speech.

It should of course be noted that the circuit will suppress only a pure speech signal. It will not, for example, suppress the voice of a DJ talking over the music.

R. Vanwersch

150 Automatic battery charger

Recharging lead-acid batteries is often assumed to be an extremely straightforward matter. And that is indeed the case, assuming that no special demands are being made on the life of the battery. On the other hand, if one wishes to ensure that the battery lasts as long as possible, then certain constraints are placed upon the charge cycle.

Figure 1 illustrates the ideal charge current characteristic for a normal 12 V lead-acid battery which is completely discharged. During the first phase (A – B), a limited charging current is used, until the battery voltage reaches approximately 10 V. This restriction on the charging current is necessary to ensure that the charger is

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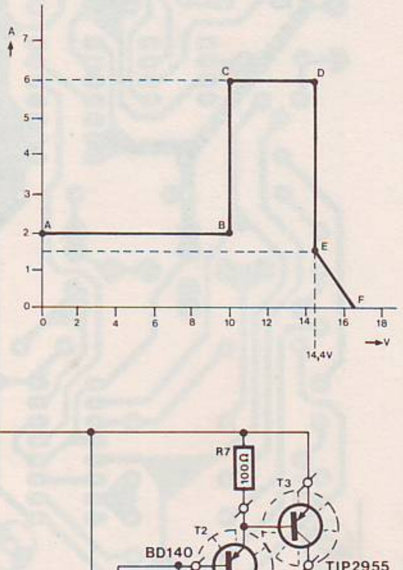

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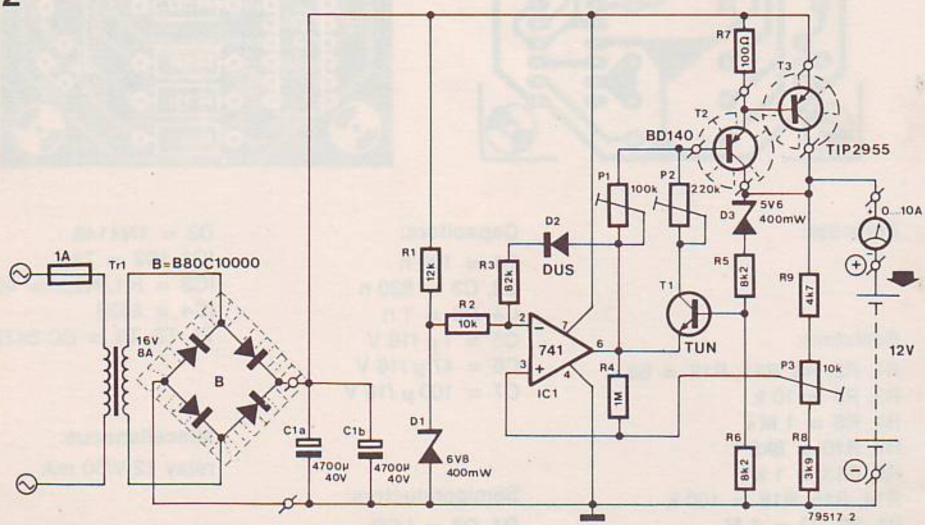

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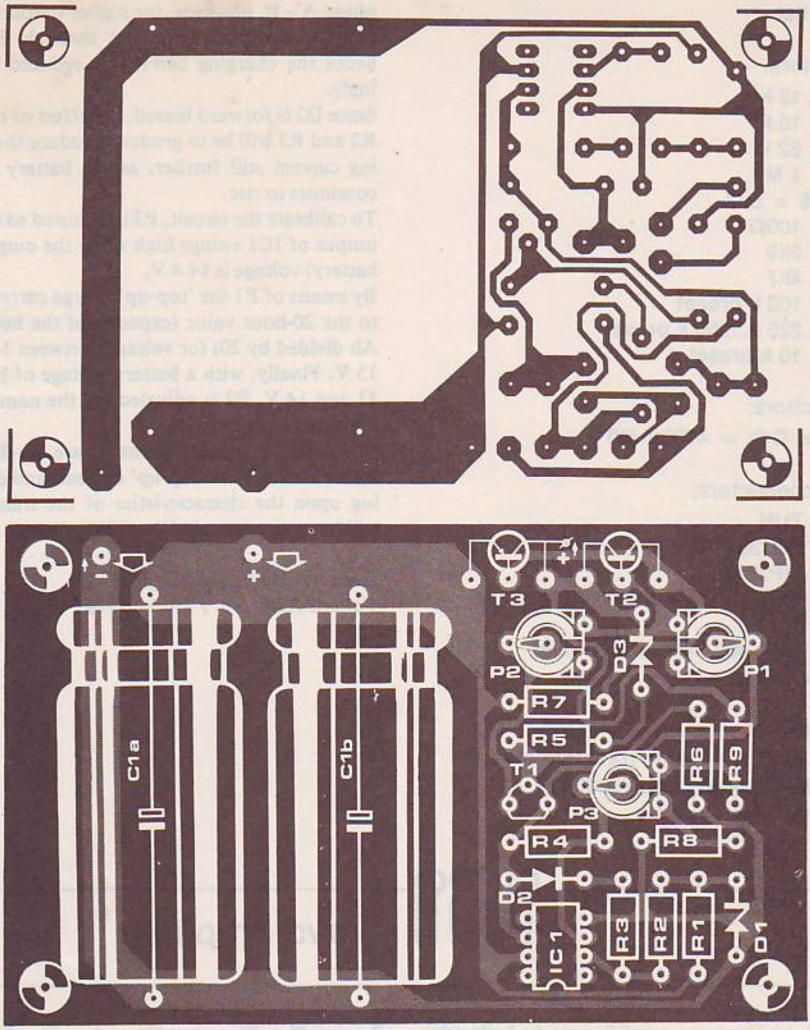

not overloaded (excessive dissipation). For the next phase (C – D), the battery is charged with the '5-hour charging current'. The size of this current is determined by dividing the nominal capacity of the battery in ampere-hours (Ah) by 5. At the end of this period the battery should be charged to 14.4 V, whereupon the final phase (E – F) starts. The battery is charged with a much smaller 'top-up' current, which gradually would decrease to zero if the battery voltage were to reach 16.5 V.

The circuit described here (see figure 2) is intended to provide a charge cycle which follows that described above. If the battery is completely discharged (voltage $<$ 10 V), so little current flows through D3 that T1 is turned off. The output of IC1 will be low, so that the base currents of T2 and T3, and hence the charging current, are determined solely by the position of P1.

If the battery voltage is between 10 and 14 V, D3 is forward biased and T1 is turned on. The output of IC1 still remains low, so that the charging current is now determined by both P1 and P2. If the wiper voltage of P3 exceeds the zener voltage of D1, then due to the positive feedback via R4, the output voltage of IC1 will swing up to a value determined by the zener voltage of D1 and the forward voltage drop of D2. As a result T1 is turned off and the charge current is once again determined by the position of P1. In contrast to

Parts list

Resistors:
R1 = 12 k
R2 = 10 k
R3 = 82 k
R4 = 1 M
R5, R6 = 8k2
R7 = 100Ω
R8 = 3k9
R9 = 4k7
P1 = 100 k preset
P2 = 220 k...250 k preset
P3 = 10 k preset

Capacitors:
C1a = C1b = 4700µ /40 V

Semiconductors:
T1 = TUN
T2 = BD138, BD140
T3 = TIP2955
D1 = 6V8, 400 mW zener diode
D2 = DUS
D3 = 5V6, 400 mW zener diode
IC1 = 741

Miscellaneous:
Tr = 16 V, 8 A mains transformer
B = B80C10000 bridge rectifier
fuse = 0.5 A slo-blo

phase A – B, however, the higher output voltage of IC1 means that current through P1, and hence the charging current, is reduced accordingly.
Since D2 is forward biased, the effect of resistors R2 and R3 will be to gradually reduce the charging current still further, as the battery voltage continues to rise.
To calibrate the circuit, P3 is adjusted so that the output of IC1 swings high when the output (i.e. battery) voltage is 14.4 V.
By means of P1 the 'top-up' charge current is set to the 20-hour value (capacity of the battery in Ah divided by 20) for voltages between 14.5 and 15 V. Finally, with a battery voltage of between 11 and 14 V, P2 is adjusted for the nominal (5-hour) charging current.
The initial charging current (phase A – B) is set by the value of the 'top-up' current, and depending upon the characteristics of the transistors, will be approximately 30 to 100% greater.

Siemens Components Report
Volume XIII, No. 1 March 1978.

151 Servo amplifier

A high quality servo amplifier can be built using only one IC and a handful of passive components. The SN28654 (Texas Instruments) contains a pulse-width modulator and an output stage that is capable of driving servomotors (see figure 1).
An input pulse at pin 3 is compared to a pulse that is generated by an internal monostable multivibrator (the 'monoflop'). The resultant pulse is stretched (using RC networks and Schmitt triggers) and passed to the output stage and from there to the motor.
The complete circuit is shown in figure 2. Apart from the RC networks (R5/C4 and R8/C5) and some decoupling capacitors, the only external components are the servo-motor and the associated servo-potentiometer. This potentiometer controls the timing of the internal monoflop, so that the motor will run until the internal pulse

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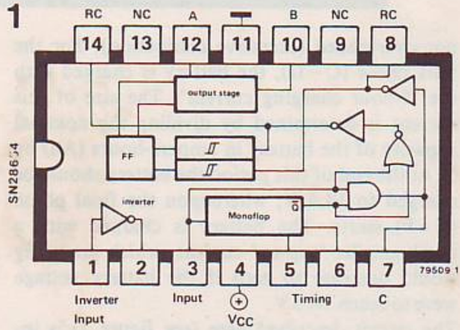

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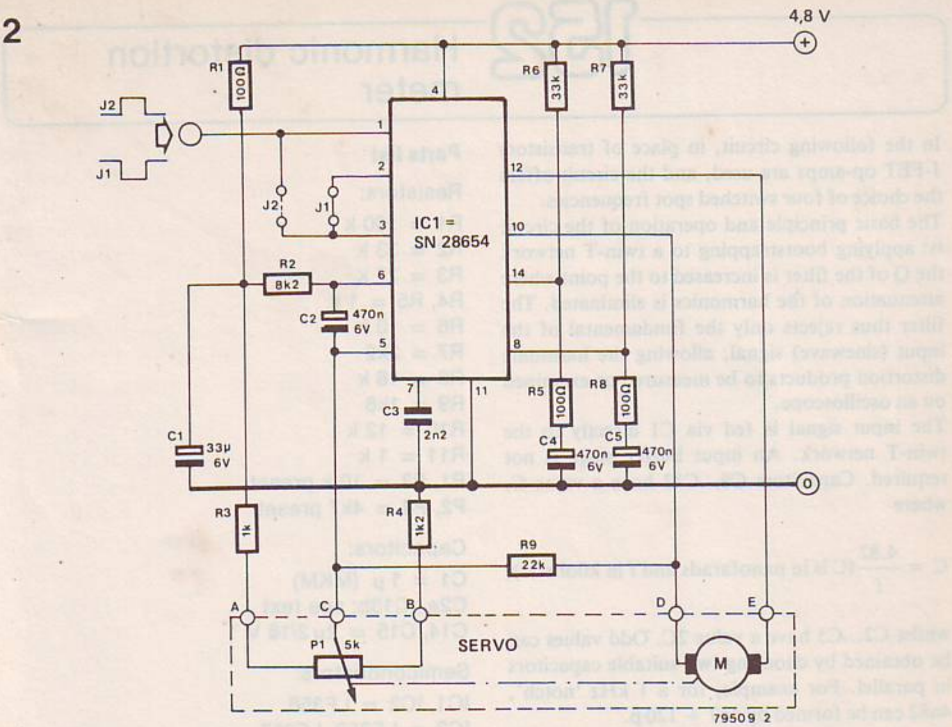

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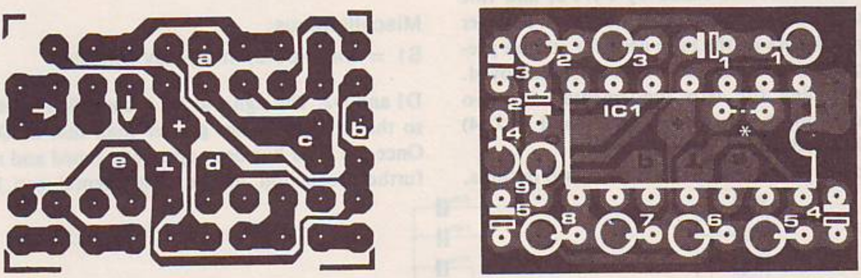

Parts list. Resistors:

R1, R5, R8 = 100 Ω
R2 = 8k2
R3 = 1 k
R4 = 1k2
R6, R7 = 33 k
R9 = 22 k

Capacitors:

C1 = 33 μ /6 V
C2, C4, C5 = 0.47 μ /6 V
C3 = 2n2

Semiconductors:

IC1 = SN 28654

length corresponds to the input pulse – provided the motor is connected the right way round, of course!

The printed circuit board (figure 3) offers the option of including the inverter (between pins 1 and 2) in the circuit if required. This means that either negative or positive input control pulses can be used.

The advantages of this servo amplifier are:

- high output current: 400 mA without external transistors;
- motor control in both directions with a single supply voltage;
- adjustable 'dead zone' (determined by C3);
- power consumption less than 800 mW.

152 Harmonic distortion meter

In the following circuit, in place of transistors J-FET op-amps are used, and the circuit offers the choice of four switched spot frequencies.

The basic principle and operation of the circuit is: applying bootstrapping to a twin-T network the Q of the filter is increased to the point where attenuation of the harmonics is eliminated. The filter thus rejects only the fundamental of the input (sinewave) signal, allowing the harmonic distortion products to be measured or examined on an oscilloscope.

The input signal is fed via C1 directly to the twin-T network. An input buffer stage is not required. Capacitors C6...C13 have a value C, where

$$C = \frac{4.82}{f} \text{(C is in nanofarads and f in kilohertz),}$$

whilst C2...C5 have a value 2C. Odd values can be obtained by choosing two suitable capacitors in parallel. For example, for a 1 kHz 'notch', 4n82 can be formed by 4n7 + 120 p.

The filter is coarse tuned by P1/P3, and fine tuned by P2/P4. Inexpensive multi-turn trimmer potentiometers of the type used for station pre-set controls in radios and TV's can be employed. When tuning for zero fundamental, the two branches of the network (P1/P2 and P3/P4) should be adjusted alternately.

The distortion signal is available at two outputs, D1 and D2. The signal at D2 is amplified by IC3 so that it is ten times greater than that at D1. Once the filter has been optimally tuned and no further reduction in the fundamental can be

Parts list

Resistors:
R1 = 100 k
R2 = 33 k
R3 = 27 k
R4, R5 = 1 k
R6 = 10 k
R7 = 2k2
R8 = 18 k
R9 = 1k8
R10 = 12 k
R11 = 1 k
P1, P3 = 10 k preset
P2, P4 = 4k7 preset

Capacitors:
C1 = 1 μ (MKM)
C2a...C13b: see text
C14, C15 = 2μ2/16 V

Semiconductors:
IC1, IC2 = LF356
IC3 = LF356, LF357

Miscellaneous:
S1 = three-pole, multi-way switch

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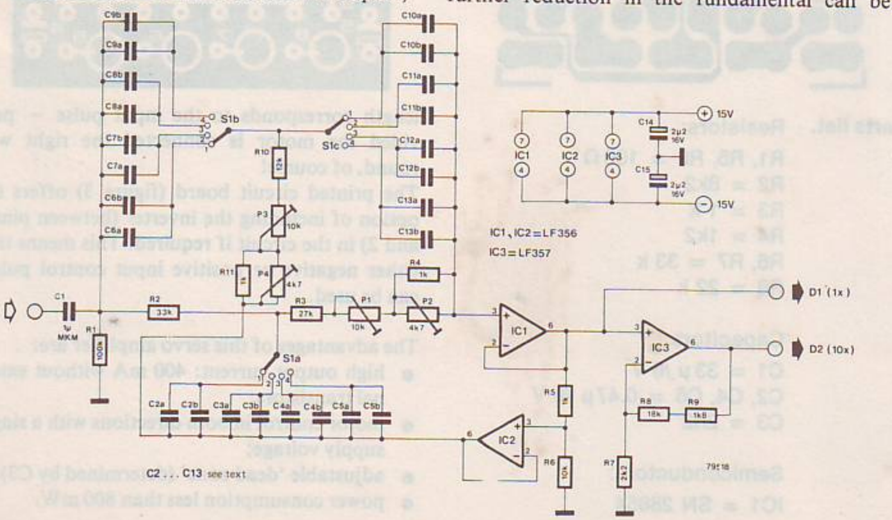

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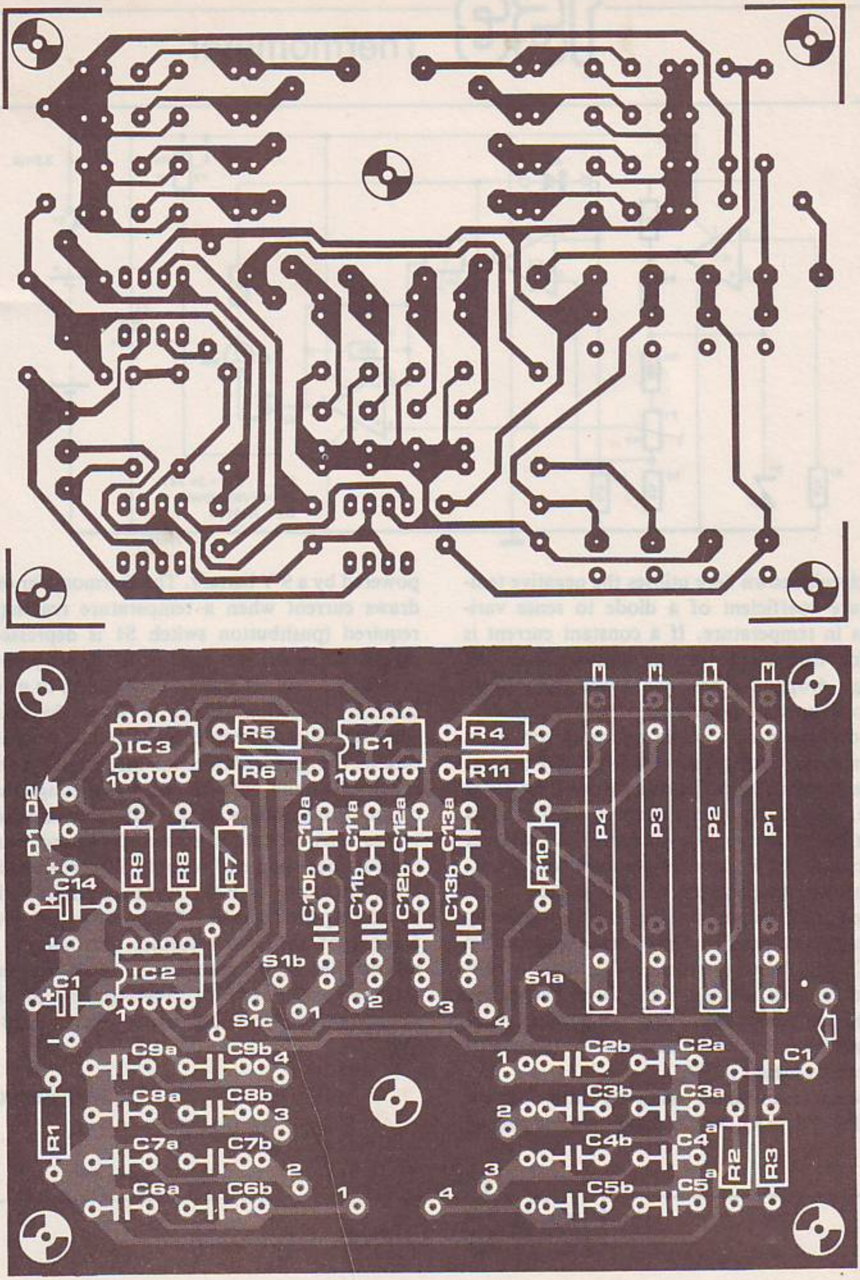
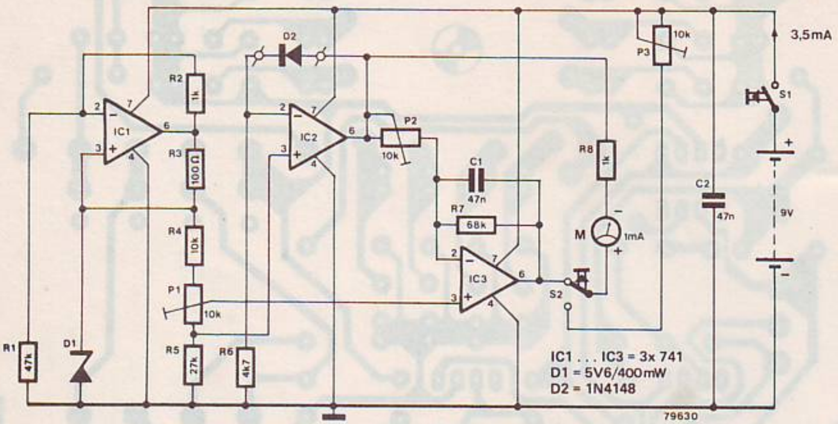


obtained, the peak-peak value of the distortion signal (D_{pp}) and the peak-peak value of the input signal U_{ipp} should be measured. The percentage distortion can then be calculated as follows:

$$\%d_{pp} = \frac{U_{Dpp} \cdot 100}{U_{ipp}} \text{ for D1, and}$$

$$\%d_{pp} = \frac{U_{Dpp} \cdot 10}{U_{ipp}} \text{ for D2.}$$

153 Thermometer



The circuit shown here utilises the negative temperature coefficient of a diode to sense variations in temperature. If a constant current is flowing through a forward-biased diode, the voltage dropped across the diode is inversely proportional to temperature.

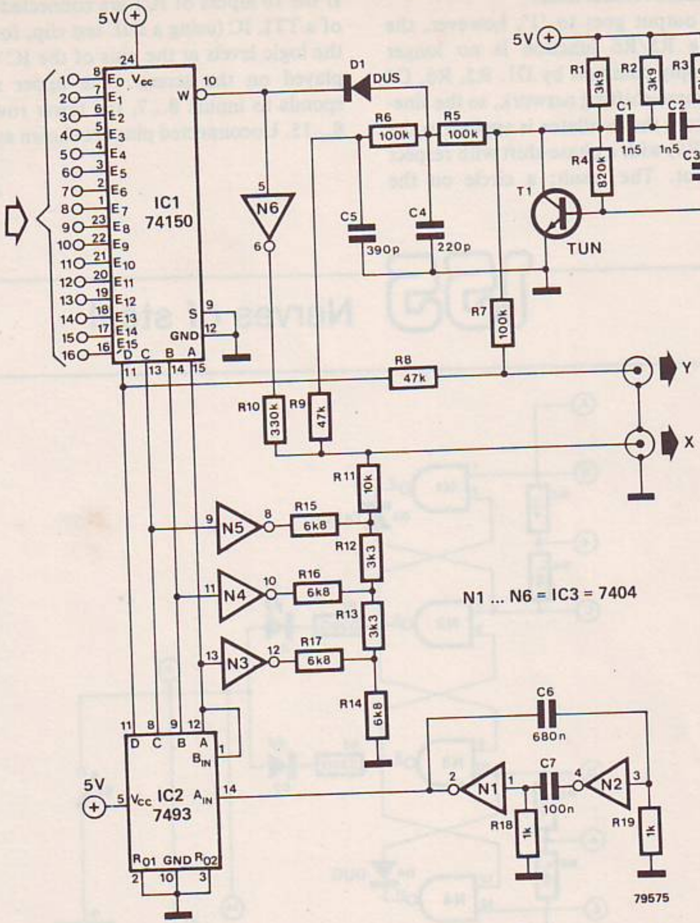
In order to obtain a stable reference voltage, a 'super zener' configuration is used. IC1 ensures that a constant current flows through the zener diode, so that the zener voltage is unaffected by variations in the supply voltage. A 5.6 V zener was chosen for its low temperature coefficient. When the temperature of the sensor diode changes, the output voltage of IC2 varies by roughly 2 mV per °C. This voltage is amplified by IC3 and fed to the meter. The meter is calibrated for zero reading at the lower end of the desired temperature scale (e.g. 0°C) by means of P1, and for full-scale deflection at the top end of the scale by means of P2.

The circuit consumes relatively little current (roughly 3.5 mA), which means that it can be powered by a 9 V battery. The thermometer only draws current when a temperature reading is required (pushbutton switch S1 is depressed). Switch S2 allows the state of the battery to be monitored, and P3 should be adjusted to give a suitable deflection. However since the meter reading will also be influenced by the temperature of the sensor diode, the measurement thus obtained should only be taken as a rough indication of the battery state. With the component values shown in the diagram, the circuit has a measurement range of roughly 50°C (depending upon the setting of P2). The range can be varied by altering the value of R7 (e.g. R7 = 33 k gives a range of 100°C). A further possibility is to reverse the meter connections, i.e. if the scale was previously 0 to 50°C, reversing the connections to the meter would give a scale of −50 to 0°C.

S. Jacobsson

154 Digisplay

The idea itself isn't new, but, the simplicity of the circuit described here lends it a special charm... Only three ICs and a handful of other components are required to recognise the logic levels of sixteen different signals and display them on an oscilloscope screen. The display consists of two rows of noughts and ones. This is achieved as follows. If a sine-wave is applied to the Y-



input of an oscilloscope, the display depends on the signal applied to the X-input. If a sawtooth is applied, the sinewave is traced on the screen; if there is no signal on the X-input, a vertical line will be displayed; and, finally, if a sinewave of the same frequency as the first but with different phase is applied, a circle or ellipse can be obtained. The vertical line or circle can be positioned at any point on the screen by adding a suitable DC offset to the X- and/or Y-input signals. In the circuit described here, two rows of eight lines or circles are displayed.

The circuit is shown in figure 1. Up to sixteen input signals are fed to the inputs of IC1. IC2 is a four-bit binary counter, and it applies binary numbers from 0 to 15 to the A, B, C and D inputs of IC1. When the number 'ØØØØ' is applied, the signal at input 1 (E_0, pin 8) of IC1 is passed (in inverted form) to its output, W. As the count at the A...D inputs proceeds, the rest of the inputs 2...16 are also scanned in sequence and passed to the output. When a '1' is present at the selected input, the output signal from IC1 is at logic zero. The voltage at the R5/R6 junction is clamped to supply common via D1 and the output of N6 is 'high', so the X-output signal is determined by the output of IC2 and the resistor network R11...R17. This signal is the 'DC component' that is required to step the display along the eight positions in one horizontal row.

The Y-output signal consists of two components. A 'DC shift' signal is taken from the D-output of IC2, to switch the display from the upper to the lower row and back, as required. Superimposed on this signal is the output from a simple RC oscillator (T1). If all 16 inputs to IC1 are at logic one (so that the W output is always 'Ø'), the display will therefore consist of two

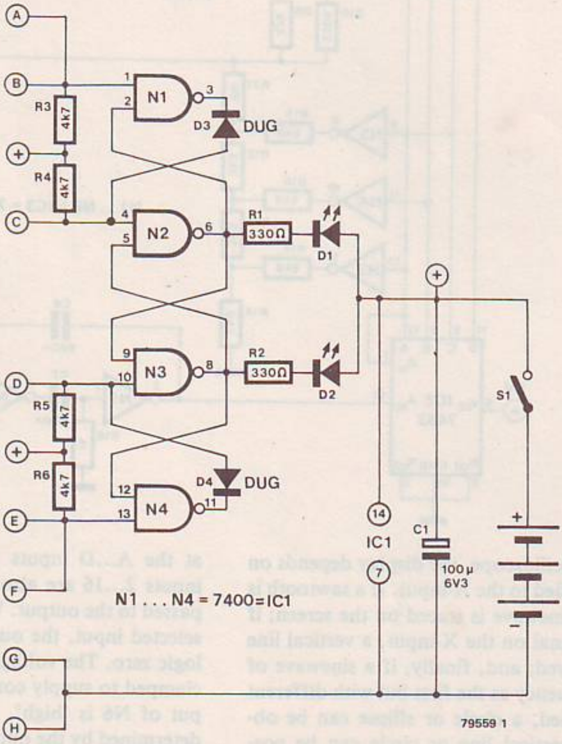
rows of eight short vertical lines.
When the W output goes to '1', however, the voltage at the R5/R6 junction is no longer clamped to supply common by D1. R5, R6, C4 and C5 are a phase-shifting network, so the sine-wave output from the oscillator is applied to the X-output (via R9) with a phase-shift with respect to the Y-output. The result: a circle on the screen.

If the 16 inputs of IC1 are connected to the pins of a TTL IC (using a DIL test clip, for instance), the logic levels at the pins of the IC will be displayed on the screen. The upper row corresponds to inputs Ø...7, the lower row to inputs 8...15. Unconnected pins are shown as 'ones'.

A. Kraut

155 Nerves of steel

1



Behind the above title lies a wellknown type of dexterity game, in which two players each attempt to pass a ring along a length of wire without touching it. The first player to reach the end of the wire is the winner. If however a player's ring should happen to brush the wire, an LED lights, indicating that he must go back to the start and begin again. The circuit incorporates an additional refinement in that, whilst one player's 'go-back-to-start' LED is lit, the other player can touch his own wire without incurring a penalty (i.e. without his own LED lighting up), thereby enabling him to speed up. However the second player must be careful, since the moment the first player reaches the start again, his LED will go out, simultaneously enabling the LED of the second player.
The actual circuit is straightforward, being based on the operation of two flip-flops formed by N1...N4. At the start of the game, once both

player's rings have touched the start electrodes (C and D), the outputs of N2 and N3 are high (and LEDs D1 and D2 are extinguished), whilst the outputs of N1 and N4 are low. The free inputs of N2 and N3 are also low, i.e. at a potential just above the forward voltage drop of a germanium diode (roughly 0.2 V). Assume now that player 1 touches his wire (B). The input of N1 is momentarily taken low, which takes the output of N1 high and the output of N2 low. The 'go-back-to-start' LED of player 1 thus lights up, whilst the outputs of N3 and N4 remain unchanged.

What happens now if player 2 touches the wire (with D1 still lit)? The input of N4 (E) is momentarily taken low, thus taking the free input of N3 high. Since the other input of N3 is low, the output of N3 will remain high, so that LED D2 cannot light up. This situation will only change when the first player once more touches the start electrode, taking the output of N2 high again.

Figure 2 shows a sketch of a possible layout for the game. Ordinary fairly stiff copper wire can be used, and obviously the 'difficulty factor' can be varied depending upon the shape into which the wire is bent and upon the diameter of the rings.

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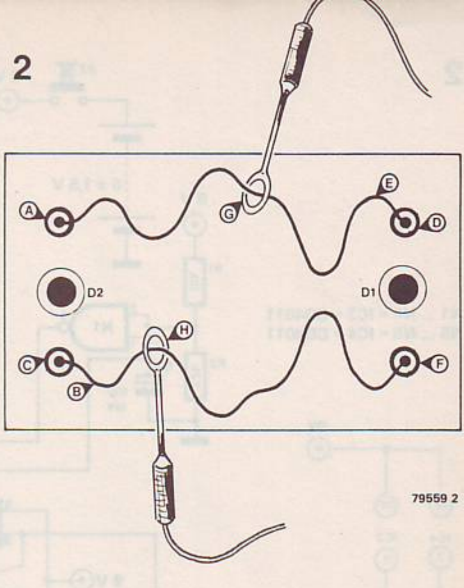

R.J. Horst

156 Bicycle speedometer

Circuits for bicycle speedometers have been fairly common, the difference in this particular design being the digital readout. The speed sensing is carried out by a number of magnets attached to the spokes or rim of the wheel which operate a pair of reed switches. The principle is illustrated in the drawing in figure 1, where the reed switches are shown fitted on the bicycle front forks. The main advantage that a digital display has over a moving coil meter is that of robustness in a situation where the younger generation can create a very harsh environment. Current consumption is kept to a minimum by arranging for the power supply to be switched on only when a readout is required. This switch (S2) should ideally be mounted on the handlebars (i.e. using an electric bicycle horn button or similar).

The circuit diagram for the digital speedometer is shown in figure 2. The principle behind the circuit is uncomplicated: The pulses from the reed switches are fed to a counter (IC1, IC2) for a predetermined length of time. The counter is

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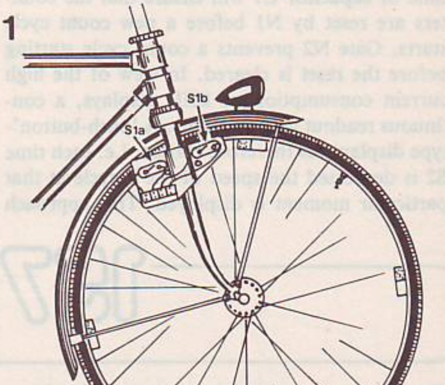

then inhibited and the count decoded and displayed. Decoding and display drive is performed by the counter itself. N3 and N4 serve to eliminate contact bounce from the reed switches, S1a and S1b, whilst the count pulses are fed to IC1

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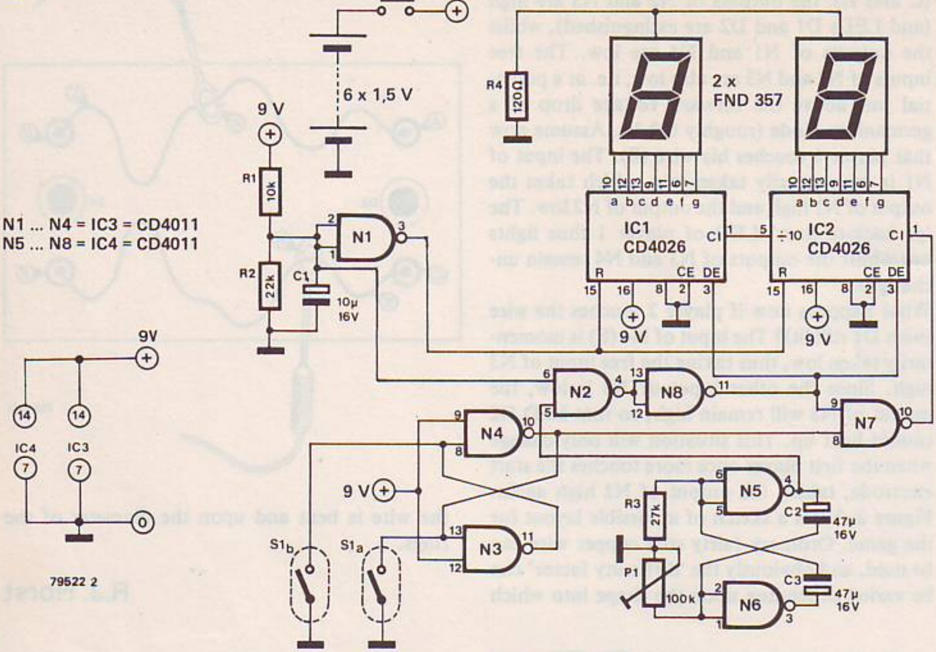

via N7. The measurement period is determined by the circuit round N5, N6, and can be varied by adjusting P1. The meter can therefore be calibrated with the aid of this preset. The charge time of capacitor C1 will ensure that the counters are reset by N1 before a new count cycle starts. Gate N2 prevents a count cycle starting before the reset is cleared. In view of the high current consumption of LED displays, a continuous readout is not feasible. A 'push-button'-type display was therefore chosen, i.e. each time S2 is depressed the speed of the bicycle at that particular moment is displayed. This approach also means that the components which would have been required to ensure that the counter is automatically reset after each count can be dispensed with.

In principle any number of magnets can be employed, however in order to avoid excessively long count periods, a minimum of three is recommended. The circuit should be calibrated (i.e. P1 adjusted for the desired count period) with the aid of an existing speedometer.

P. de Jong

157 Noise level meter

There are many potential applications nowadays to justify the use of a noise level meter – for instance, monitoring the sound output at dances, discos etc. The unit described here was designed primarily to establish the noise level produced by model engines. It has five switched ranges from 70 dB to 120 dB in 10 dB steps and is readable to ½ dB. The prototype was found to be accurate to ± 1 dB.

The circuit for the noise level meter is shown in figure 1. The sound signal is picked up by the microphone M1 and filtered by the network C1, C2, R1 and R2. These components, together with the capacitance of the microphone and the input impedance of the amplifier, ensure that the frequency response of the system is corrected to suit the internationally standardised 'A' weighting curve shown in figure 2. This 'weighted' sig-

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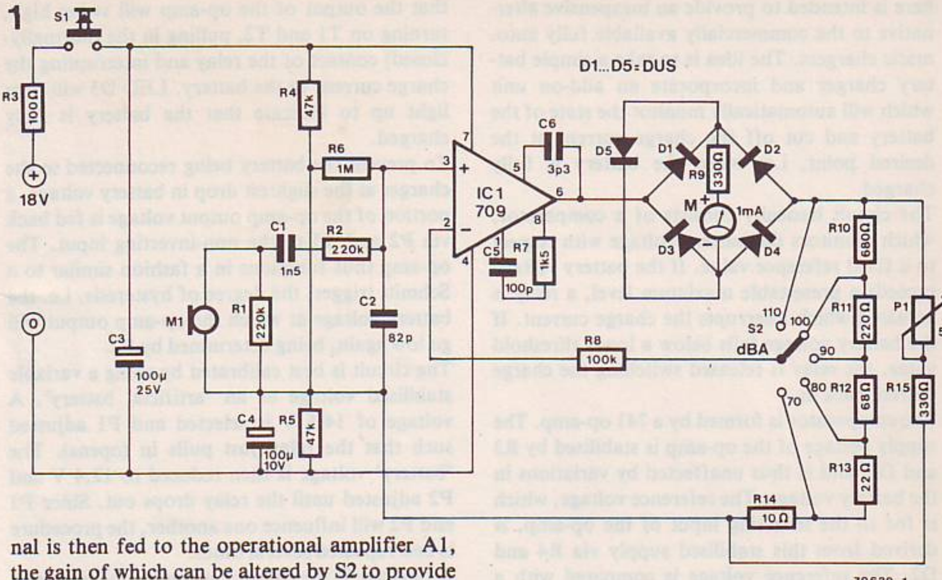


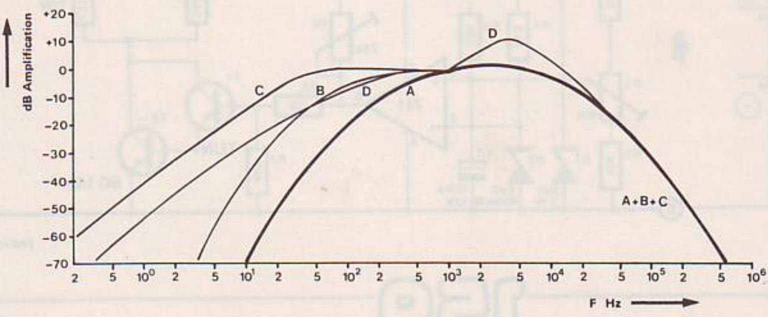
nal is then fed to the operational amplifier A1, the gain of which can be altered by S2 to provide five noise ranges.

The AC output of the op-amp is then rectified by diodes D1...D4 and fed to the meter via resistor R9. As this rectifier is included in the feedback loop the meter reading remains linear over the entire scale. Diode D5 is included to limit the current through the meter to a safe value, thereby reducing the risk of damage if a 'loud' noise is measured on a 'quiet' range. Components C5, C6 and R7 are included to provide frequency compensation and to prevent instability.

Under normal operation the circuit will only draw about 2 mA, so it can be powered by two PP3 (or similar) batteries. The push-button switch S1 ensures that the circuit is not inadvertently left on. The meter should be calibrated in dBs and should have a full scale deflection of + 10 (normal log scale).

P. Barnes

2



158 Automatic battery charger

Automatic battery chargers are not particularly cheap, however the protection they afford against overcharging and possible battery damage is highly desirable. The circuit shown here is intended to provide an inexpensive alternative to the commercially available fully automatic chargers. The idea is to take a simple battery charger and incorporate an add-on unit which will automatically monitor the state of the battery and cut off the charge current at the desired point, i.e. when the battery is fully charged.

The circuit basically consists of a comparator, which monitors the battery voltage with respect to a fixed reference value. If the battery voltage exceeds a presettable maximum level, a relay is actuated which interrupts the charge current. If the battery voltage falls below a lower threshold value, the relay is released switching the charge current back in.

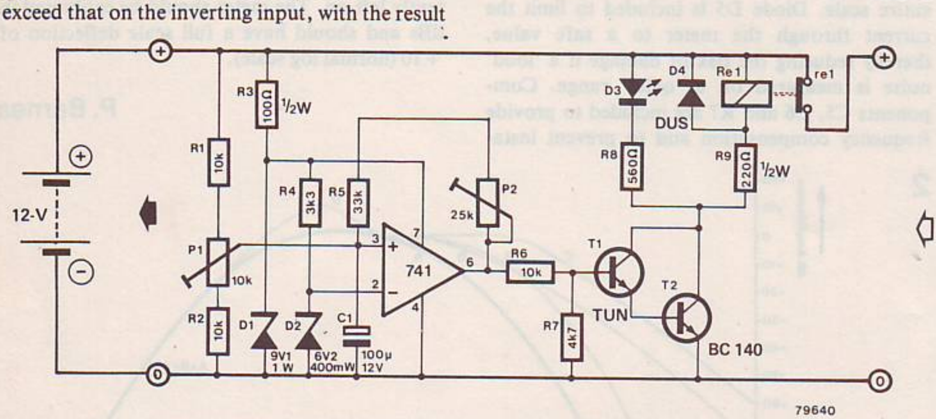
The comparator is formed by a 741 op-amp. The supply voltage of the op-amp is stabilised by R3 and D1, and is thus unaffected by variations in the battery voltage. The reference voltage, which is fed to the inverting input of the op-amp, is derived from this stabilised supply via R4 and D2. The reference voltage is compared with a portion of the battery voltage, which is taken from the voltage divider, R1/P1/R2. As the battery voltage rises, at a certain point (determined by the setting of P1) the voltage on the non-inverting input of the op-amp will eventually exceed that on the inverting input, with the result that the output of the op-amp will swing high, turning on T1 and T2, pulling in the (normally-closed) contact of the relay and interrupting the charge current to the battery. LED D3 will then light up to indicate that the battery is fully charged.

To prevent the battery being reconnected to the charger at the slightest drop in battery voltage, a portion of the op-amp output voltage is fed back via P2 and R5 to the non-inverting input. The op-amp thus functions in a fashion similar to a Schmitt trigger, the degree of hysteresis, i.e. the battery voltage at which the op-amp output will go low again, being determined by P2.

The circuit is best calibrated by using a variable stabilised voltage as an 'artificial battery'. A voltage of 14.5 V is selected and P1 adjusted such that the relay just pulls in (opens). The 'battery' voltage is then reduced to 12.4 V and P2 adjusted until the relay drops out. Since P1 and P2 will influence one another, the procedure is best repeated several times.

A final tip: if the charge current is too large to be switched by the relay, the circuit can still be used by connecting the relay in the primary of the battery charger transformer.

H. Heere



159 5-minute chess clock

In games of speed chess, where each player has only 5 or 10 minutes to complete all his moves, mechanical chess clocks leave something to be desired in terms of accuracy, especially when both players have only 30 or 40 seconds left. The author of the circuit described here offers a solution to this problem by employing LEDs to provide an unequivocal display which counts off the time remaining in multiples of 10 seconds. The clock uses two counters, one for player A

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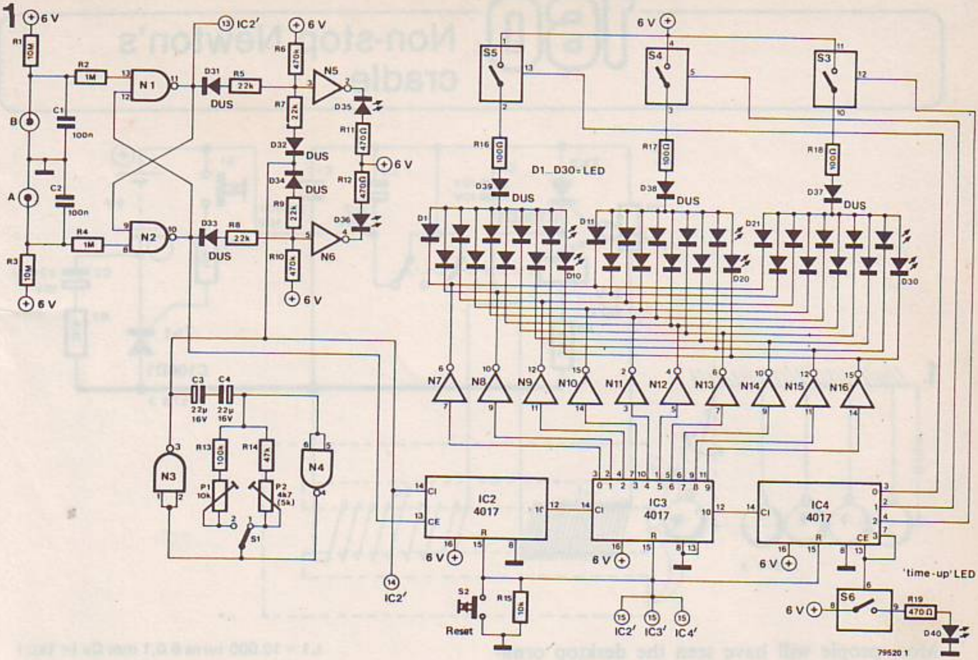

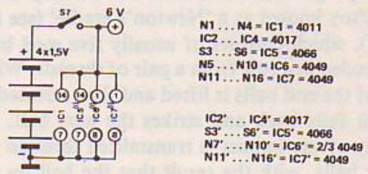

and one for player B. By bridging a set of touch contacts each player can stop his own counter and start his opponent's. The state of each counter is displayed on a circle of 30 LEDs (see figure 2). In a 5-minute game S1 is set to position 1, whereupon each LED lights up in turn for (300 seconds/30 =) 10 seconds. With S1 in position 2, the time limit is increased to 10 minutes per player, i.e. each LED lights up for 20 seconds.

If a player exceeds his time limit, then LED D40 (A) or D40' (B) lights up. The counters can be reset for the start of a new game by pressing S2. LEDs D35 and D36 provide a visual indication of who is to move.

Assuming player B has just made a move on the board, he presses TAP switch B, which takes the output of N1 (which together with N2 forms a set/reset flip-flop) high. The output of N2 goes

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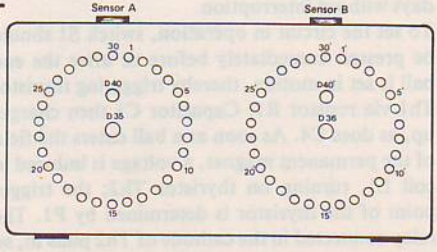
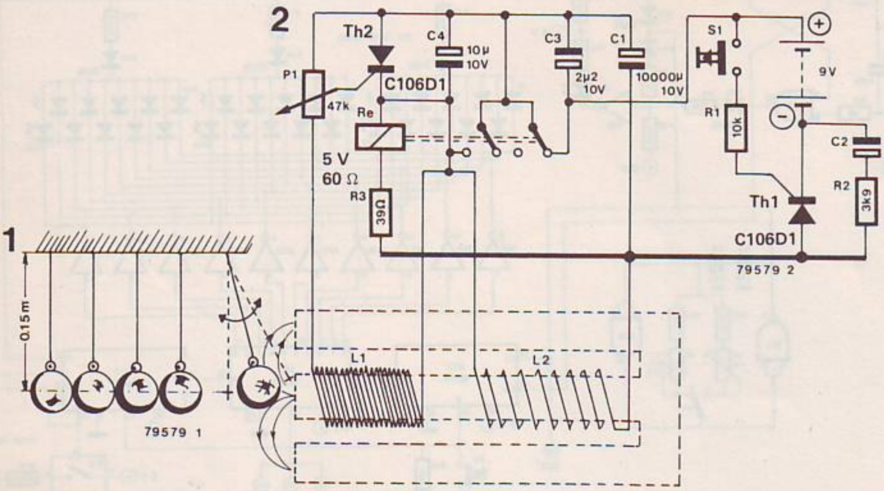


low, causing counter A (IC2, IC3, IC4) to start counting the clock pulses provided by N3 and N4; counter B is inhibited until TAP switch A is touched. Since D31 is now reverse biased, D35 will be turned on and off via D32 at a rate equal to the clock frequency. D33 is forward biased, pulling the input of N6 low, so that D36 will be extinguished.

The circuit can be powered by four 1.5 V batteries or by ni-cads. The current consumption is approximately 45 mA. P1 and P2 can be calibrated using a known accurate timebase; each LED should light up for 10 seconds with S1 in position 1 and twenty seconds with S1 in position 2.

S. Woydig

160 Non-stop Newton's cradle



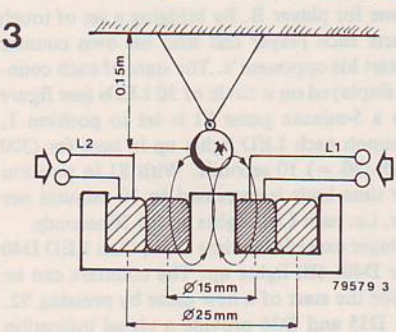
L1 = 10.000 turns Ø 0,1 mm Cu (= 1kΩ)

L2 = 2300 turns Ø 0.4 mm Cu (= 25 Ω)

Most people will have seen the desktop ornament/toy known as a 'Newton's cradle' (see figure 1), which consists of usually five steel balls suspended in a row from a pair of threads. When one of the end balls is lifted and then released so that it falls back and strikes the next ball, the energy of the impact is transmitted through the other balls, with the result that the ball on the opposite end of the row swings up. It then in turn falls back, energy is again transmitted through the row, and the first ball swings up, and so on. The energy losses of the system are fairly high, and after a number of oscillations the balls are returned to rest. The idea behind the circuit described here, is to compensate for the natural energy losses of the system, so that it continues to oscillate indefinitely (i.e. until the circuit is disconnected or the batteries run out!).
If, for the moment we ignore the energy losses, the frequency at which the system oscillates will be:

$$f = \frac{1}{2\pi \cdot \sqrt{\frac{l}{g}}}$$

where l is the length of the thread and g is the force of gravity (9.81 m/s²). Thus with a length of 0.15 m, the fundamental frequency of the system will be approximately 1.3 Hz. In order to compensate for natural energy losses, the magnetic field system shown in figure 2 has been designed. Together with the accompanying circuit, the idea is that a magnetic force is applied to one of the end balls in the cradle. If the circuit is powered by 6 1.5 V cells (manganese-alkali), the cradle should continue to oscillate for roughly 5 days without interruption.



To set the circuit in operation, switch S1 should be pressed immediately before or after the end ball is set in motion, thereby triggering thyristor Th1 via resistor R1. Capacitor C1 then charges up, as does C4. As soon as a ball enters the field of the permanent magnet, a voltage is induced in coil L1, turning on thyristor Th2; the trigger point of the thyristor is determined by P1. The relay connected in the cathode of Th2 pulls in, so

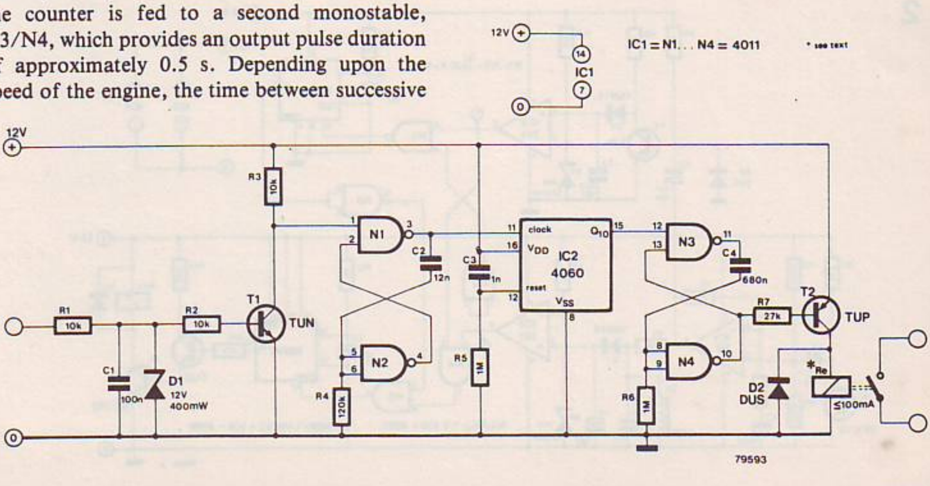
that current flows through coil L2, and an additional magnetic field is created which repels the ball. As soon as the ball leaves the magnetic field, the voltage induced in L1 collapses and Th2 is turned off. The process then repeats itself at the natural frequency of the system.
If the ball is stopped, no charge current will flow to C1, with the result that C2 will discharge. If the discharge current is smaller than the holding current of Th1, the latter turns off and circuit switches off. Figure 3 shows a cross-section of the coil and magnet system. L1 (10,000 turns of enamelled copper wire, 0.01 mm diameter, 1 k) and L2 (2300 turns enamelled copper wire, 0.4 mm diameter, 25 Ω) are wound on a permanent magnet core, and enclosed in transformer laminations. Any readily available alternative type of thyristor can be used.

K. Bartkowiak

161 Varispeed windscreen wiper delay circuit

In most windscreen wiper delay circuits the wiper speed is independent of the speed of the car. However the faster the car travels, the more rain falls on the windscreen, therefore, ideally, the shorter the delay should be. A variable delay circuit could be controlled by a sensor mounted in the speedometer cable. However this approach would be fairly complicated. The simpler solution adopted here, is to derive the control signals from the contact breaker, so that the wiper speed is varied in accordance with the engine speed.
The input of the circuit is connected to the contact breaker; when the contacts open, the full battery voltage appears across the input, with the result that T1 provides a short output pulse. The resultant pulse stream is used to trigger the monostable multivibrator formed by N1 and N2. The frequency of the multivibrator is then divided by ten by the counter, IC2. The output of the counter is fed to a second monostable, N3/N4, which provides an output pulse duration of approximately 0.5 s. Depending upon the speed of the engine, the time between successive pulses will be between roughly 10 and 40 seconds. Thus transistor T2 is regularly turned on for a short period, causing the wiper relay to pull in and the wipers to perform a single sweep. By arranging for a capacitor of roughly 2.2 μ F to be switched in parallel with C4, the wipers can be made to perform a double sweep every cycle.
The zener diode D1 is included to protect the circuit from excessively large surge voltages appearing across the contact breakers, whilst diode D2 protects T2 against the back EMF induced by the relay. Preferably, the holding current of the relay should not exceed 100 mA; if that is the case, however, a transistor with a higher output current capability should be used.

D. Laues



162 IR lock

The following circuit is intended as an infra-red lock for house doors, garage doors, etc. Since the 'key' is almost impossible to copy, it should provide an effective obstacle to unwanted visitors.

Figure 1 shows the infra red transmitter. An astable multivibrator, formed by NAND gates N1...N3, drives an output transistor, T1, which turns the infra-red emitter diode on and off at a frequency which can be varied by means of P1.

The receiver circuit is shown in figure 2. Light pulses received by the phototransistor T1 are amplified by IC1 and fed to an LC circuit tuned to roughly 23 kHz. The filtered output signal is rectified by D1 and fed to op-amp IC2, which is connected as a Schmitt trigger. The trigger threshold is set by zener diode D4 to 2.4 V. The unfiltered output of IC1 is also fed to a second Schmitt trigger, (IC3). The output of this op-amp (point 1) will remain high as long as the voltage level at its input is 2.4 volts or greater regardless of the frequency of the received signal.

Assuming point 1 is high, a positive going edge at the output of IC2 (point 2) will 'turn the lock' as follows: when point 2 goes high, the output of N1 also goes high, and with it the input of monostable multivibrator MMV1. However, since this monostable is triggered by a negative going edge, the output state of the monostable remains unchanged, i.e. the Q output remains low. The

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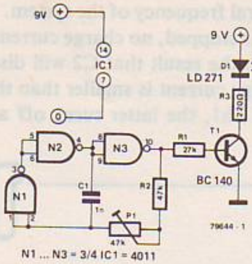
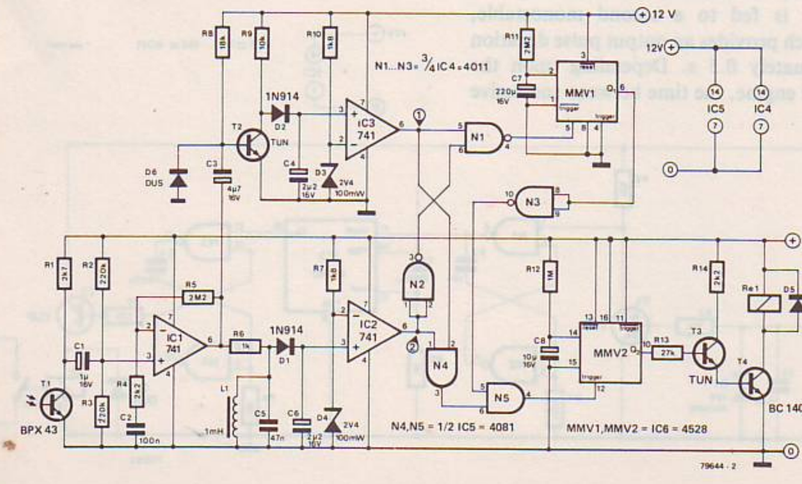


positive going edge at point 2 is also transferred to the trigger input of MMV2, which since it is triggered by positive going pulses, turns on the Darlington pair T3/T4 and pulls in the relay. Thus for the pulse duration of MMV2 the lock is 'opened'.

If the modulation frequency of the transmitter signal deviates from 23 kHz, only point 1 will be high; point 2 will go low, with the result that, via N1, the negative going edge will trigger MMV1. Thus for the pulse duration of MMV1 – which is several minutes – MMV2 cannot be triggered. Even if the modulation frequency is subsequently corrected, since one input of N5 is held low, the lock cannot be opened during this period. If a flip-flop is used in place of a relay, the circuit could, for example, be used to switch a car alarm system on and off.

H.J. Urban

2



163 Sequencer

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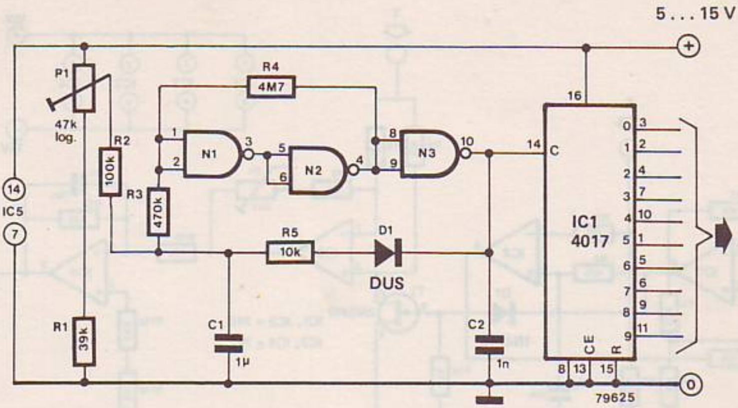

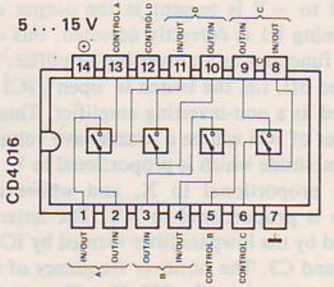

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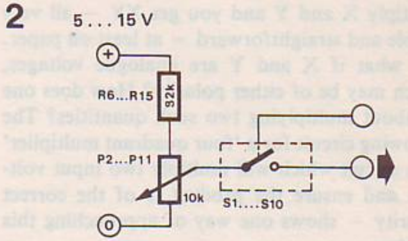
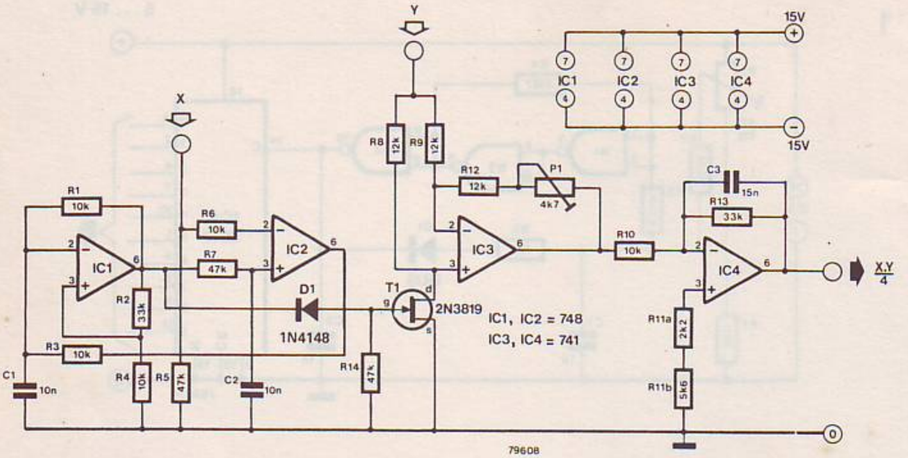


N1 ... N3 = 3/4 IC5 = 4011
S1 ... S10 = IC2,IC3,IC4 = 4016

The following design for a sequencer, which will generate a 10 note analogue waveform, is distinguished by its relative simplicity. To control a synthesiser two types of signal are required: a gate pulse to trigger the envelope shaper (ADSR), and a control voltage for the voltage controlled oscillators (VCOs).
The VCO voltages are generated as follows. An oscillator, formed by N1, N2 and N3, clocks a decade counter (IC1). Each output of the counter is connected to an analogue switch (as shown in figure 2), the input voltage of which can be varied by means of a potentiometer. The outputs of all the switches are joined together, so that an analogue waveform, composed of 10 discrete voltage levels, is generated at this point. The frequency of the resultant signal can be varied by means of P1.
The gate signal for the ADSR is derived from the clock signal, however since each synthesiser places different demands on the type of gate pulse required, no circuit is given.
Readers may wish to experiment with extending the circuit. One possibility is to include a monostable multivibrator (at the clock input of IC1), which allows one to cycle through the analogue waveform step by step. Each of the preset voltage levels on the inputs of S1...S10 are then compared with a reference voltage. If a shorter cycle (i.e. less than 10 steps) is required, the appropriate output of IC1 should be connected to the reset input (pin 15).

J.C.J. Smeets

164 Four quadrant multiplier



Multiply X and Y and you get XY – all very simple and straightforward – at least on paper. But what if X and Y are analogue voltages, which may be of either polarity? How does one go about multiplying two such quantities? The following circuit for a 'four quadrant multiplier' – a circuit which will multiply two input voltages and ensure the product is of the correct polarity – shows one way of approaching this problem.

Basically the circuit generates a squarewave signal, whose duty-cycle is proportional to one of the input signals and whose amplitude is proportional to the other. The average value of the squarewave, and hence the value of the product voltage, is obtained by lowpass filtering.

The squarewave generator is formed by IC1, R1, R2, R4 and C1. The output of IC1 is lowpass filtered by R7 and C2, then compared with the input voltage, X. The duty cycle of the squarewave is modulated via the output of IC2, R3 and C1, whilst the amplitude of the output signal of IC1 is held constant. The output of IC1 is also used to control the FET switch, T1. When this switch is 'closed' i.e. T1 is turned on, a voltage equal to −Y is present at the output of IC3; assuming P1 is correctly adjusted, this op-amp then functions as an inverting amplifier. If T1 is turned off, i.e. the switch is 'open', IC3 is connected as a non-inverting amplifier. Thus at the output of IC3 will be a squarewave voltage with an amplitude which is proportional to Y, a duty-cycle proportional to X, and whose average value is proportional to XY. The latter is obtained by the lowpass filter formed by IC4, R10, R13 and C3. The turnover frequency of this filter is approximately 330 Hz. The circuit will quite happily multiply analogue signals with frequencies which are an order of magnitude lower than the turnover point of the two lowpass filters. The author has used the circuit for correlation measurements on very low frequency EEG signals.

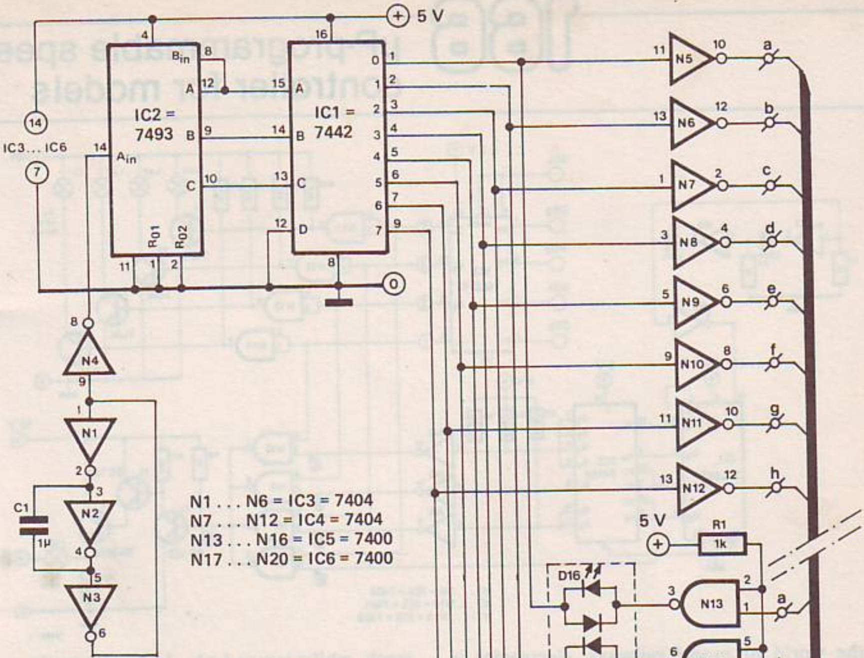
The adjustment of P1 is necessary since, when conducting, T1 has a significant resistance. With an input voltage X = 0 (input grounded) and Y = +6 or −6 V, P1 should be adjusted for minimum output voltage of IC4 (roughly ± 40 mV).

P. Creighton

165 Ribbon cable tester

For microcomputer enthusiasts and anyone working with large scale digital circuits, a ribbon cable tester can prove a useful aid.

The circuit described here will simultaneously

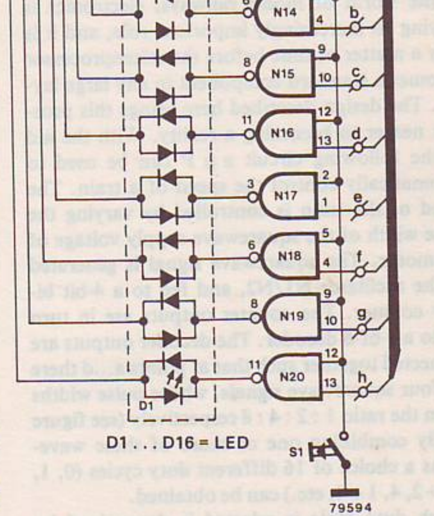


test 8 cores, with the facility for extending this to 16. A clock oscillator (N1...N3) drives a 4-bit counter (IC2). Three of the counter's outputs are used to clock a BCD-decimal decoder (IC1). The outputs of the decoder each go low in turn for the duration of a certain clock period. The outputs are connected via inverters N5...N12 to a set of terminals, to which one end of the ribbon cable is attached. The other end of the cable is connected to the inputs of gates N13...N20. Between the outputs of these gates and the outputs of IC1 8 pairs of reverse-parallel connected LEDs (D1...D16) are inserted.

The odd-numbered LEDs will light up only if the corresponding NAND output is low **and** the corresponding output of IC1 is high (87½% of the time). The even-numbered LEDs on the other hand, will light up only if the NAND outputs are high and the outputs of IC1 are low (12½% of the time).

If there is a break in one of the cores, the corresponding NAND output will be low and the associated LED will light up. If the core is intact, then the LED will be extinguished, since the logic levels on either side of the LED change state simultaneously.

The circuit will also check for shorts between cores, since in that case the anode of an even-numbered LED will be high, whilst the cathode will be low, causing the LED to light up. Note



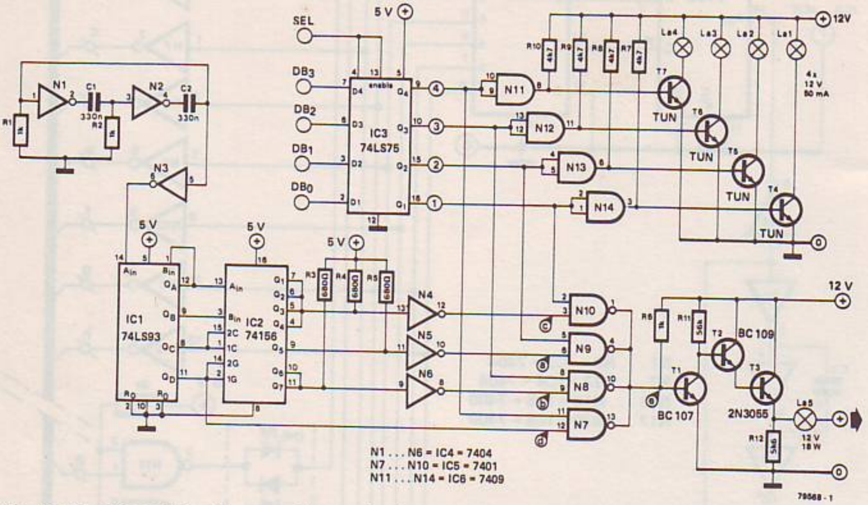
that series resistors for the LEDs are not necessary. If no cable is connected, the odd-numbered LEDs will light up. Switch S2 functions as a lamp test for the even-numbered LEDs.

For a 16-core version of the circuit a 74154 should be used in place of IC1 (the D-input is of course used), whilst the number of inverters, NAND gates and LEDs is doubled.

J.J. van der Weele

166 µP-programmable speed controller for models

1



In the world of model railways, electronics is playing an increasingly important role, and it is only a matter of time before the microprocessor becomes a standard component in any large layout. The design described here brings this prospect nearer to becoming a reality. With the aid of the following circuit a µ P can be used to automatically control the speed of a train. The speed of the train is controlled by varying the pulse width of the squarewave supply voltage of the motor. The squarewave signal is generated by the oscillator N1/N2, and fed to a 4-bit binary counter. The counter outputs are in turn fed to a 1-of-8 decoder. The decoder outputs are connected together such that at points a...d there are four squarewave signals, whose pulse widths are in the ratio 1 : 2 : 4 : 8 respectively (see figure 2). By combining one or more of these waveforms a choice of 16 different duty cycles (0, 1, 2, 1 + 2, 4, 1 + 4, etc.) can be obtained.

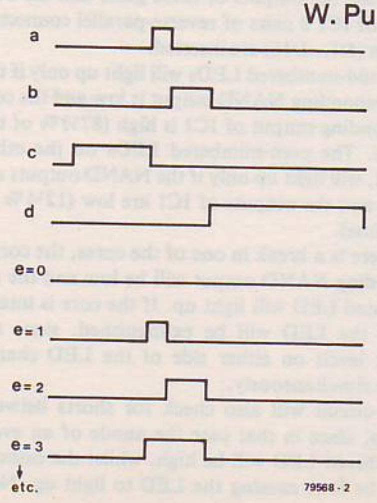
Which duty cycle is selected is determined by NAND gates N7...N10; the output state of these gates is in turn determined by the information present on the data bus (DB0...DB3) of the microprocessor system. Between the µ P data bus and the NAND gates is a 4-bit latch (IC3). Information is only transferred from the data bus to the gates when a select pulse is received.

The waveform selected by the µ P is amplified by T1/T2/T3. Lamp L5 protects the circuit from excessive current in the event of a short on the track, whilst lamps La1...La4 indicate the speed of the train in binary code.

If a µ P system is not available, the 'manually-operated' processor circuit shown in figure 3 can be used instead. The circuit performs the same basic function as a µ P, with the exception that the 'brainwork' is done by the hobbyist himself. By pressing either S1 or S2 the speed of the train can be increased or reduced in single steps. If the circuit of figure 3 is used, then IC3 in figure 1 can of course be omitted.

2

W. Pussel



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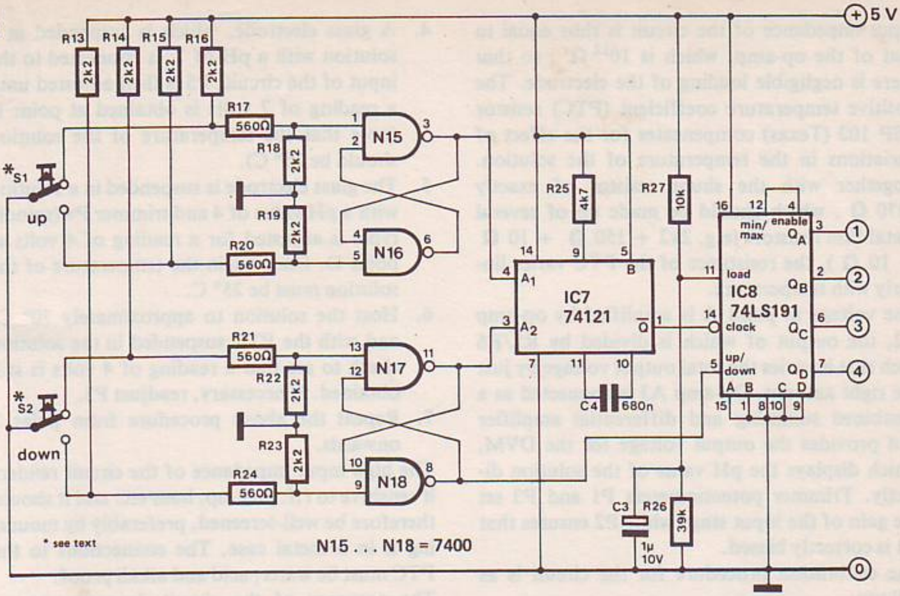

167 pH meter circuit for DVM

To accurately measure the concentration of hydrogen ions (pH value) in a solution, a 'glass electrode' is often used in chemistry laboratories. The electrode is constructed on the principle of a galvanic cell, and the output voltage of the electrode is proportional to the pH value of the solution to be measured. The temperature of the solution considerably affects the pH value, thus a pH meter is effectively a temperature compensated millivoltmeter.

The circuit shown employs an op-amp (A1) to amplify the output voltage of the electrode. The

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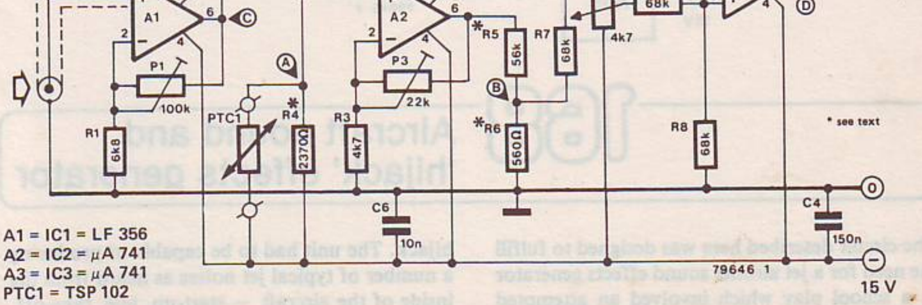


input impedance of the circuit is thus equal to that of the op-amp, which is $10^{12}\ \Omega$, so that there is negligible loading of the electrode. The positive temperature coefficient (PTC) resistor TSP 102 (Texas) compensates for the effect of variations in the temperature of the solution. Together with the shunt resistor of exactly 2370 Ω, which should be made up of several metal film resistors (e.g. 2k2 + 150 Ω + 10 Ω + 10 Ω), the resistance of the PTC varies linearly with temperature.

The voltage at point A is amplified by op-amp A2, the output of which is divided by R5/R6 such that it varies the total output voltage by just the right amount. Op-amp A3 is connected as a combined summing and differential amplifier and provides the output voltage for the DVM, which displays the pH value of the solution directly. Trimmer potentiometers P1 and P3 set the gain of the input stage while P2 ensures that A1 is correctly biased.

The calibration procedure for the circuit is as follows:

1. With the inputs short-circuited P2 is adjusted for zero volts at point C.
2. Again with the inputs shorted, potentiometer P5 (wirewound type) is adjusted such that 7 volts are present at point D.
3. Trimmer potentiometer P4 (spindle type) is adjusted such that, with the PTC at a temperature of 25° C, zero volts are present at point A.
4. A glass electrode, which is suspended in a solution with a pH of 7, is connected to the input of the circuit. P5 is then adjusted until a reading of 7 volts is obtained at point D (note that the temperature of the solution should be 25° C).
5. The glass electrode is suspended in a solution with a pH value of 4 and trimmer P4 (spindle-type) is adjusted for a reading of 4 volts at point D. Once again the temperature of the solution must be 25° C.
6. Heat the solution to approximately 70° C, and with the PTC suspended in the solution check to see that a reading of 4 volts is still obtained. If necessary, readjust P3.
7. Repeat the above procedure from point 3 onwards.

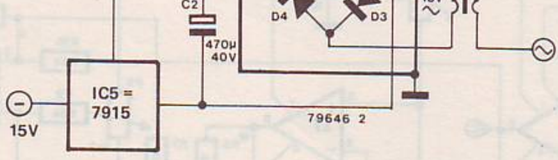
The high input impedance of the circuit renders it sensitive to r.f. pick-up, hum etc. and it should therefore be well-screened, preferably by mounting it in a metal case. The connections to the PTC must be water, acid and alkali proof.

The accuracy of the circuit depends upon a stable supply voltage (± 15 V), and upon the accuracy of the reference solution used during calibration (not to mention the accuracy of the DVM).

Glass electrodes are available commercially, and are supplied with instructions on how they should be used.

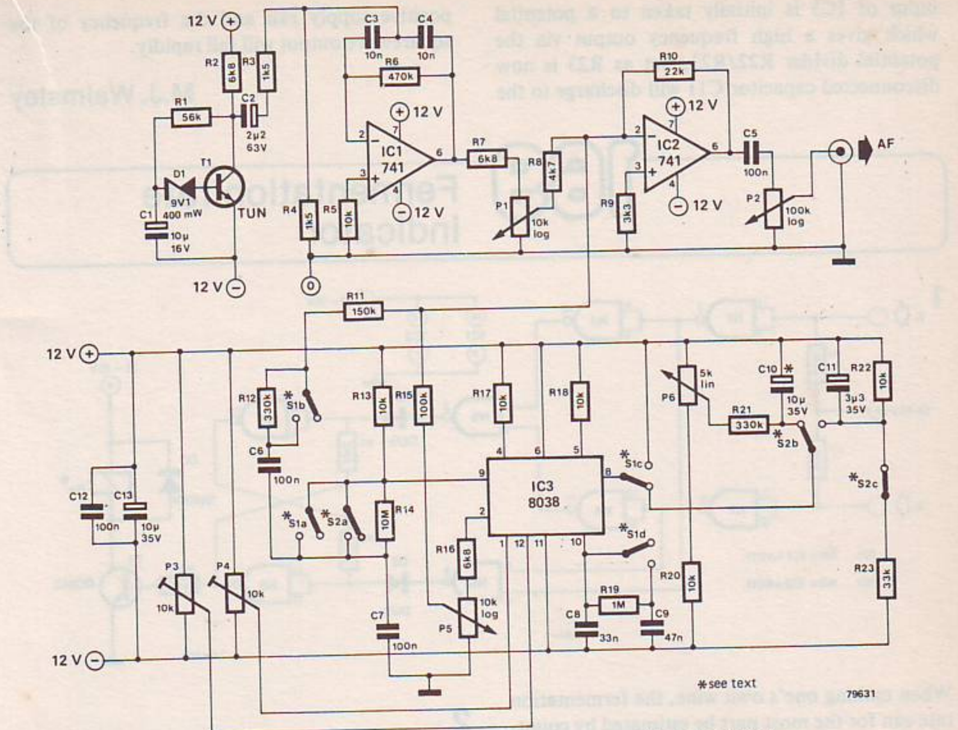
Th. Rumbach

2



168 Aircraft sound and 'hijack' effects generator

The circuit described here was designed to fulfill the need for a jet airliner sound effects generator in a school play which involved an attempted hijack. The unit had to be capable of producing a number of typical jet noises as heard from the inside of the aircraft – start-up, idle, take-off,



P1 = jet roar volume
P2 = overall volume
P5 = compressor whistle volume
P6 = throttle
S1 = gunfire
S2 = tyre squeal

in-flight, approach, landing and reverse thrust conditions – together with tyre squeal on landing and machine-gun fire.

To simulate the sound of a jet engine both the roar from the 'hot end' of the engine and the whistle from the compressor (whose pitch varies with engine speed) are required. The engine roar is obtained by feeding white noise through a band pass filter which emphasises frequencies around 800 Hz. Transistor T1 and the zener diode D1 form the white noise generator whose output is fed to IC1, the band pass filter. The volume of the roar can be altered by potentiometer P1.

The whistle is derived from the sinewave output of the 8038 waveform generator IC3 whose frequency range is set by C8 and is typically between 10 Hz and 10 kHz. The actual frequency is determined by the throttle control P6 which is connected to the FM input of IC3 via switches S1c and S2b, while the whistle volume is controlled by potentiometer P5. Engine inertia (lag in response to throttle demands) is realistically imitated by the integrating network R21/C10. C10 should be a low-leakage type – if available, a 10 μ paper capacitor would be a good choice.

Both these signals, the engine roar and compressor whistle, are then summed by IC2 and passed to the external amplifier through the overall volume control P2. By varying the settings of these controls all of the above mentioned jet engine sounds can be realised. The purity of the sinewave signal can be adjusted by potentiometers P3 and P4.

The gunfire effect is obtained from the squarewave output of IC3 when switch S1 is closed. By closing this switch the squarewave is allowed to pass through to the summing amplifier and the FM input of IC3 is taken high to give minimum frequency whilst the frequency range itself is also decreased by the addition of C9 in parallel with C8. Resistor R19 is included so that C9 is always kept charged to the average voltage across C8 to prevent a 'chirp' when S1 is first closed.

The tyre sqeal effect is also obtained from the squarewave output of IC3. When switch S2 is closed the squarewave is enabled, and the FM input of IC3 is initially taken to a potential which gives a high frequency output via the potential divider R22/R23, but as R23 is now disconnected capacitor C11 will discharge to the positive supply rail and the frequency of the squarewave output will fall rapidly.

M.J. Walmsley

169 Fermentation rate indicator

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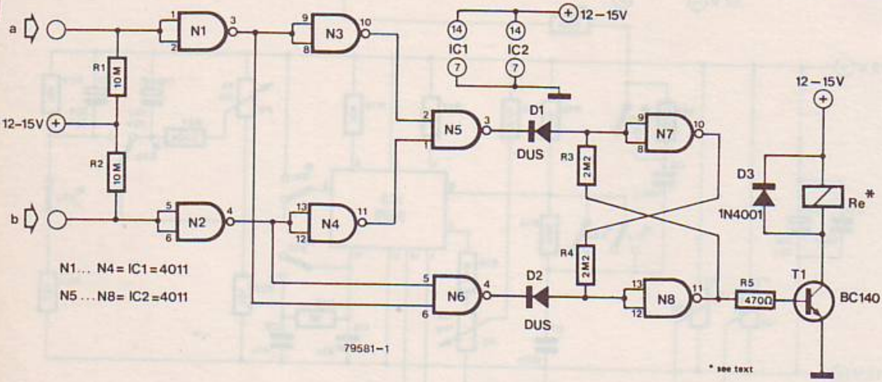

When making one's own wine, the fermentation rate can for the most part be estimated by counting the number of times the level of the (sterilising) liquid in the air-lock rises and falls as a result of the CO_2 which is produced. Towards the end of the fermentation process however, the level tends to 'jitter' somewhat, so that accurate measurements are not possible. One solution to this problem is to employ two electrodes, one of which is mounted higher than the other – see figure 2. The difference in height between the two should be greater than that by which the level of liquid fluctuates (approximately 2 mm).

The circuit shown in figure 1 is designed to produce an output pulse only if both electrodes are suspended in the liquid, after previously having been clear of the liquid. Enamelled copper wire, 0.3 mm in diameter is used for the electrodes. Insulating sleeving is pushed over the wire, whilst an earth connection is also suspended in the liquid.

As is apparent from the circuit diagram, the input of inverters N1 and N2 are held high via pull-up resistors R1 and R2 when neither of the electrodes is in contact with the liquid. The output of the OR-gate formed by N3, N4 and N5 is therefore low, as is the output of the set-reset flip-flop N7/N8. The output of NAND gate N6 is high. If the level of the liquid in the fermen-

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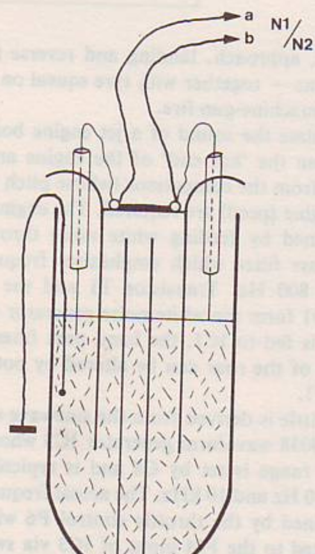

tation lock rises to cover the lower of the two electrodes, the output of the corresponding inverter will go high. This has no effect upon the output of the NAND gate, but the output of the OR-gate will be taken high also. Due to the effect of diode D1, however, the set-reset flip-flop remains in its original state. Should the liquid level fall, the only result will be that the output of the OR-gate is returned low once again. Only if the level rises still further to cover the second electrode will the output of N6 go low and the flip-flop be triggered, turning on T1 and feeding a pulse to the counter (Re). Since the flip-flop can only be triggered by '0' logic levels via D1 and D2, both electrodes must clear the liquid before the flip-flop is reset and another pulse can be counted.

Any normal 12 V impulse counter can be used.

J. Ryan

170 Frequency to voltage converter for multimeter

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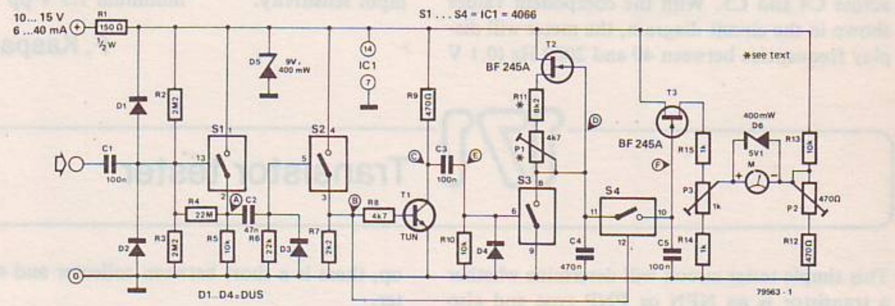

Most frequency measurements are carried out by means of a digital frequency meter or else on an oscilloscope. However both these instruments are relatively expensive, and thus not often 'standard equipment' for the hobbyist. One way to measure the frequency of a signal without investing in specialised equipment is to use a frequency-voltage converter, which can then be 'plugged in' to an ordinary multimeter. That is the function of the circuit described here. A meter with a 5 V range should be used; the conversion ratio is linear, assuming the scale is calibrated in ms (1 V = 5 ms). The circuit is built round a quad analogue switch IC, type 4066. The squarewave signal at point A is switched via S1 to the differentiating network, C2/R6. The resulting pulses are then fed via S2 on the one hand to the inverter formed by T1, and on the other hand to S4. The result is that S3 and S4 open and close alternately, i.e. S3 is open when S4 is closed, and vice versa. Assuming that S4 is closed, capacitor C4 will be charged linearly by the constant current source T2. The charge is transferred via S4 to storage capacitor C5. S4 and C5 thus function as a sample and hold circuit. If now S4 is opened, S3 will close; capacitor C4 will discharge via S3 to ground, and new measurement cycle will begin. Depending upon

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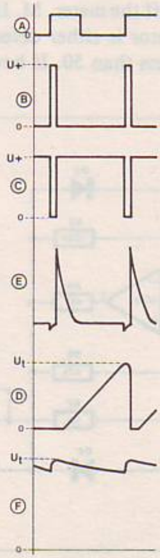


the characteristics of the FET, T3, the sample and hold circuit will increase the voltage by approximately 2 V. Thus a maximum charge voltage of around 6.5 V is possible.

The circuit is calibrated as follows: With the input unconnected, the wiper of P2 is turned to the positive end stop (junction of P2 and R13). A DC voltage of 6.5 V is applied to the gate of T3, and P2 is adjusted for full-scale deflection on the meter. Potentiometer P4 is adjusted for zero reading on the meter with 0 V on the gate of T3. A known frequency is then fed to the input of the circuit (e.g. 50 Hz mains signal from a doorbell transformer), and P2 is then fine tuned for a reading of 20 ms.

The pulse diagram in figure 2 illustrates the signals obtained at points A...E in the circuit, and across C4 and C5. With the component values shown in the circuit diagram, the meter will display frequencies between 40 and 2000 Hz (0.1 V = 0.5 ms = 2000 Hz).

Different frequency ranges can be obtained by altering the value of components R11, R12 and C4 accordingly.

The appropriate values can be calculated by using the formula:

$$U_{C4} = \frac{I_{C4}}{C4 \cdot f_{in}}$$

where $I_{C4} = \dfrac{U_{R11} + U_{P1}}{R11 + P1}$

Finally one or two specifications:

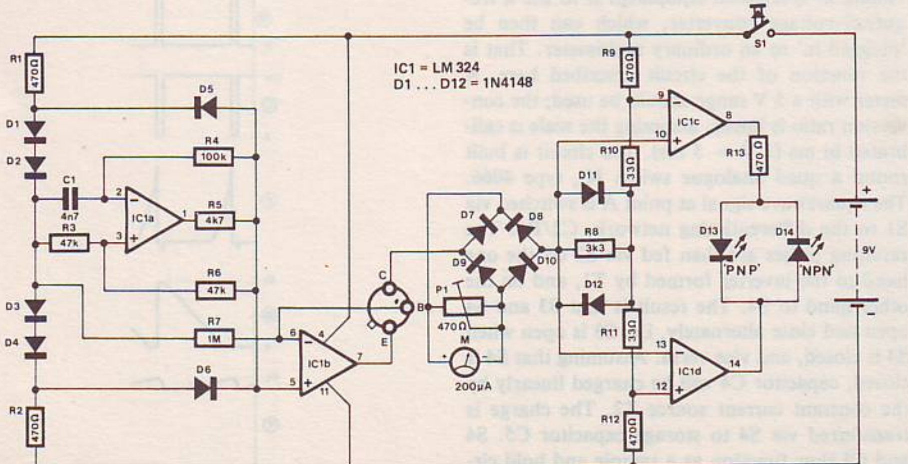
supply voltage:	10...15 V
current consumption:	5 mA
input impedance:	1 MΩ
input sensitivity:	minimum 1.5 V pp

F. Kasparec

171 Transistor tester

This simple tester circuit will determine whether a transistor is an NPN or PNP type and also measure the current gain of the unknown device. When the pushbutton switch, S, is depressed, one of the LEDs D13 or D14 will light to show the polarity of the transistor, whilst the h_{FE} can be read directly off the meter, M. If neither LED lights, the transistor is either defective or has a current gain of less than 50. If both LEDs light up, there is a short between collector and emitter.

The circuit functions as follows: IC1a forms the basis of a squarewave oscillator, the frequency of which is roughly 1 kHz. The squarewave oscillates about half supply voltage, and, with the aid of IC1b, is used to generate a base-emitter voltage which is alternately positive and negative. Thus whenever the polarity of the base bias



voltage is of correct polarity for the type of transistor under test, a base current will flow, causing a collector current to flow through R8. Depending upon the direction of the current through R8, either a positive or negative voltage is dropped across this resistor, with the result that, via IC1c or IC1d, the appropriate LED will light to signify the polarity of the transistor under test.
The collector current of the transistor also flows through the diode bridge and the meter, M. Since the base current remains more or less constant, the size of the collector current can be taken as a measure of the current gain of the transistor. Full-scale deflection of the meter corresponds to an h_{FE} of 500.
The meter can be calibrated with the aid of P1, the simplest method being to use a transistor with a known current gain.

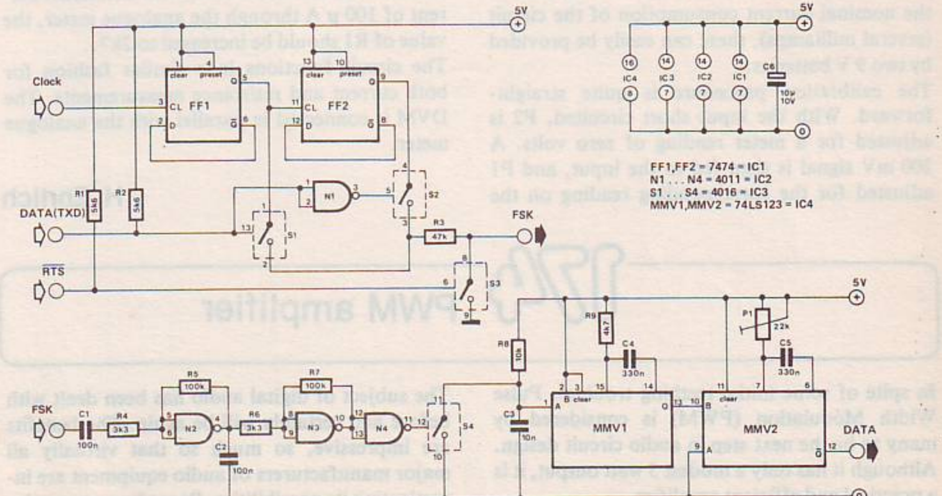
H.G. Brink

172 FSK modem

The most notable feature of this circuit for an FSK modulator/demodulator is its extreme simplicity:

- it requires only one supply voltage
- uses only 4 common ICs
- is simple to set up

The circuit conforms to Kansas City Standard (CUTS) format, i.e. logic Ø = 1200 Hz, logic 1 = 2400 Hz, and the transmission rate is 300 Baud.
The modulator circuit is quite straightforward. The clock signal is derived from the frequency of the UART, which is 16 times the baud rate (i.e. 4800 Hz). After the clock signal has been divided down by FF1 and FF2, 2400 Hz and 1200 Hz signals are available at the inputs of S1 and S2. Depending upon the logic level of the data input signal, one of these two switches is closed and a signal of the appropriate frequency is present at the modulator output. At the input of the demodulator circuit, N2, N3 and N4 form an amplifier/limiter. The actual demodulation is performed by the two re-triggerable monostable multivibrators, MMV 1 and MMV 2. The pulse duration of MMV 1 is roughly 420 μ s, whilst that of MMV 2 is approximately 850 μ s (depending upon the position of P1). With an input frequency of 2400 Hz, MMV 1 is continuously retriggered, so that its Q output is held high. No trigger pulses are fed to MMV 2, with the result that the $\overline{Q}$ (data) output remains high. However, with an input frequency of 1200 Hz, MMV 1 will

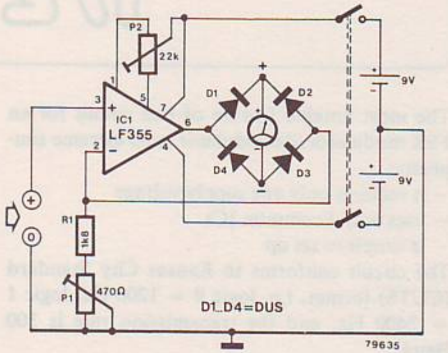


not be retriggered before the Q output goes low, so that MMV 2 is triggered and the data output also goes low. By adjusting P1 to keep the pulse duration of MMV 2 as short as possible, the delay between rising and falling edges of the data signal can also be kept short.
Users of the Elekterminal can take the clock signal for the modulator from either pin 17 or pin 40 of the UART. The Baud rate switch on the Elekterminal should be set to the 300 Baud position. The RTS input should only be used in conjunction with UARTs provided with such an output.

H. Stettmaier

173 Voltage trend meter

The advantages of a digital multimeter are sufficiently well known that they do not need to be repeated here. However there are situations where it is useful to determine whether the quantity being measured is increasing or decreasing, particularly if it is subject to sudden fluctuations. An opamp connected as an AC amplifier is particularly suited to this task.
Most simple DVMs contain an LSI chip with an input sensitivity of 200 mV and an extremely high input impedance. A suitable opamp is the LF 355 used as a voltage-current converter, which has an input impedance of $10^{12}\,\Omega$.
The circuit shown here is designed for an input voltage of 200 mV and a current through the moving coil meter of 100 μ A. For other input voltages and/or output currents the trimmer potentiometer P1 and resistor R1 should be altered accordingly.
The opamp requires two supply voltages (positive and negative) between 5 and 18 V. In view of the nominal current consumption of the circuit (several milliamps), these can easily be provided by two 9 V batteries.
The calibration procedure is quite straightforward. With the input short circuited, P2 is adjusted for a meter reading of zero volts. A 200 mV signal is then fed to the input, and P1 adjusted for the corresponding reading on the meter.



If the meter has a scale of e.g. 0...3/30, then by calibrating the moving coil meter to read '2' for a maximum reading (with e.g. 200 mV in) on the DVM, an 'overload' range up to 300 mV can be obtained. In the above case, with a constant current of 100 μ A through the analogue meter, the value of R1 should be increased to 2k7.
The circuit functions in a similar fashion for both current and resistance measurements. The DVM is connected in parallel with the analogue meter.

H. Ehrlich

174 PWM amplifier

In spite of some initial teething troubles, Pulse Width Modulation (PWM) is considered by many to be the next step in audio circuit design. Although it has only a modest 3 watt output, it is a practical and efficient amplifier.

The subject of digital audio has been dealt with before and certainly will be again. The benefits are impressive, so much so that virtually all major manufacturers of audio equipment are investigating its possibilities. Recording companies

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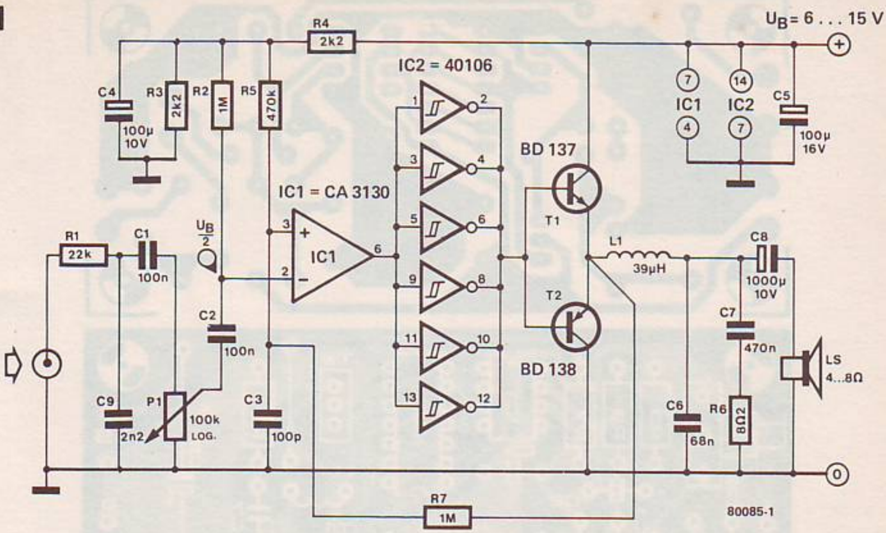


Figure 1. The self-oscillating PWM amplifier. With a 12 V supply, it will deliver 3 watts into 4 ohms.

too are aware of the potential of a digital system (digitally recorded records are already available commercially).

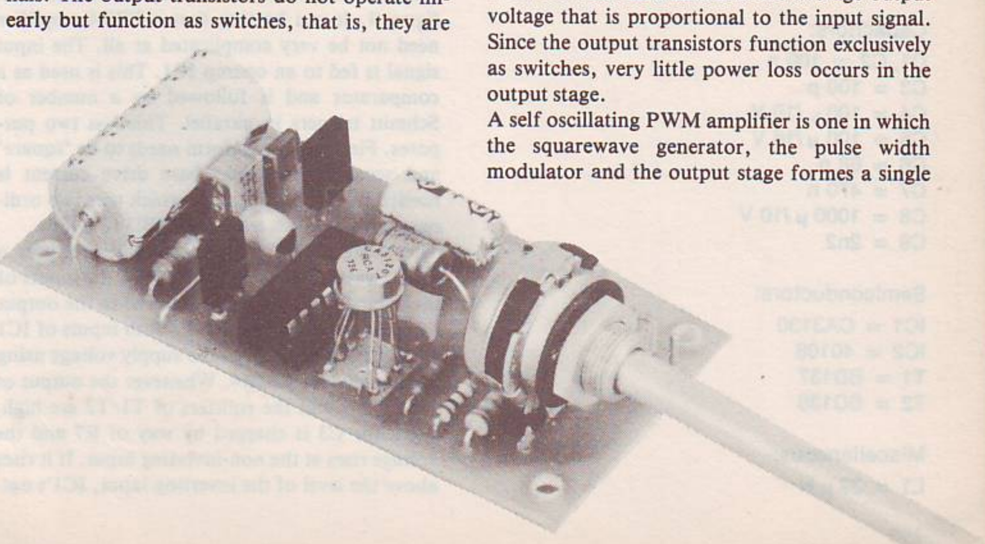
Until quite recently, the performance of PWM amplifiers was disappointing due to the poor quality semiconductors used. With the introduction of modern high speed switching transistors, PWM is now coming of age.

The PWM amplifier

It might be a good idea to recap the principles briefly. A PWM amplifier contains a symmetrical squarewave generator. The duty cycle of this squarewave is then modulated by the audio signals. The output transistors do not operate linearly but function as switches, that is, they are either full on or off. Under quiescent conditions the duty cycle of the output waveform is 50% which means that each of the output transistors is fully saturated (conducting) for an equal amount of time. The average output voltage is therefore zero. It therefore follows that if one of the output switches is closed for a longer period than the other, the average output voltage will then be either negative or positive depending on the polarity of the input signal.

It can be seen then that it is the average output voltage that is proportional to the input signal. Since the output transistors function exclusively as switches, very little power loss occurs in the output stage.

A self oscillating PWM amplifier is one in which the squarewave generator, the pulse width modulator and the output stage formes a single



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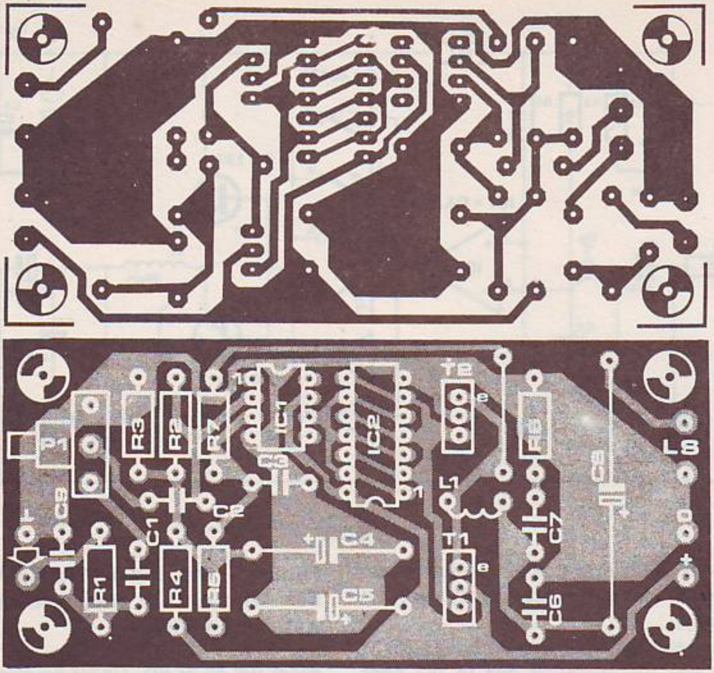


Figure 2. The printed circuit and parts layout of the PWM amplifier.

Parts list

Resistors:
R1 = 22 k
R2, R7 = 1 M
R3, R4 = 2k2
R5 = 470 k
R6 = 8 Ω 2
P1 = 100 k log. potentiometer

Capacitors:
C1, C2 = 100 n
C3 = 100 p
C4 = 100 µ /10 V
C5 = 100 µ /16 V
C6 = 68 n
C7 = 470 n
C8 = 1000 µ /10 V
C9 = 2n2

Semiconductors:
IC1 = CA3130
IC2 = 40106
T1 = BD137
T2 = BD138

Miscellaneous:
L1 = 39 µ H

unit. This produced an efficient amplifier with only a very small number of components. A version of this is described here.

The circuit diagram

The circuit of the complete amplifier is shown in figure 1. It can be seen that a PWM amplifier need not be very complicated at all. The input signal is fed to an opamp IC1. This is used as a comparator and is followed by a number of Schmitt triggers in parallel. This has two purposes. Firstly the waveform needs to be 'square' and secondly sufficient base drive current is needed for the output stage which uses two ordinary but fairly fast transistors (BD 137/138).
The entire amplifier oscillates and produces a squarewave. This is because one of the inputs of the comparator (IC1) is connected to the output by means of an RC network. Both inputs of IC1 are biased to one half of the supply voltage using voltage divider R3/R4. Whenever the output of IC1 is low and the emitters of T1/T2 are high, capacitor C3 is charged by way of R7 and the voltage rises at the non-inverting input. If it rises above the level of the inverting input, IC1's out-

put changes low to high and the emitters of T1/T2 change from high to low. As a result, C3 is now discharged through R7, the voltage at the plus input drops below that of the minus input and the output of IC1 switches back to a low state. The result is a squarewave output; the frequency of which is determined by R7 and C3. The values given result in an oscillation at 700 kHz.

Provided Murphy doesn't get in the way, we should have an oscillator. Now we have to pulse width modulate it. The level at the inverting input of IC1, which is used as a reference, does not remain constant but is determined by the audio signal. The point at which the output of the comparator changes, is also determined by the amplitude. As a result the *width* of the squarewaves is constantly changed (modulated) by the audio signal.

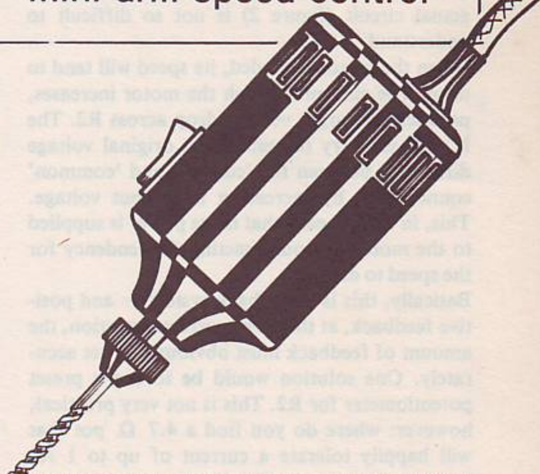
At the output of the amplifier, filtering is required: it is not supposed to act as a 700 kHz transmitter! An LC/RC network is used, consisting of L1/C6 and C7/R6.

With a load of 8 ohms and a supply voltage of 12 volts, the amplifier produced 1.6 watts. At 4 ohms, 3 watts were measured. Cooling the output transistors was not necessary. The harmonic distortion proved to be surprisingly low for such a simple design. Less than 0.32% total harmonic distortion from 20 Hz-20 kHz was measured.

Figure 2 shows the printed circuit board and parts layout for the amplifier. Its construction requires little time and money, so it offers an excellent opportunity for anyone wanting to become better acquainted with PWM.

E. Postma

175 Mini drill speed control



Miniature electric drills have been available for some time. Most of them are battery powered. For precision work, it is useful to have a speed control; if constant speed can be maintained, independent of the load, so much the better. Both of these objectives can be achieved fairly simply, using an integrated voltage regulator.

Before going into the actual circuit, it is a good idea to take a brief look at how these little DC motors work. Why does the speed drop when the motor is loaded?

Normally, a fairly constant voltage is applied to the motor. Off load, the speed increases until the power consumption is exactly sufficient to cover the electrical and mechanical losses in the motor. When the motor is loaded, the speed drops. This reduces the back EMF, so the current through the motor increases; a new equilibrium is reached when the increased power consumption equals the reduced electrical and mechanical losses plus the power delivered to the load. In other words, the motor supplies the power required by the load – but at reduced speed. Obviously, there is a limit: if the motor is loaded too heavily, it will stop.

If the speed is to remain constant, the voltage across the motor will have to be increased when the motor is loaded. In this way, the current (and the power output) can increase without affecting the speed.

In the circuit described here, the main active component is a voltage stabiliser IC, the 79G. This is a negative voltage regulator; it was chosen because its output voltage can be reduced to as low as −2.23 V. The minimum output voltage of its positive voltage counterpart, the 78G, is approximately 5 V. The extended control range at the low voltage end is important, since the motors in miniature drills are all fairly low voltage types – they are intended for battery use. This circuit can be used to power 2.5...12 V motors, with any current rating up to 1 A.

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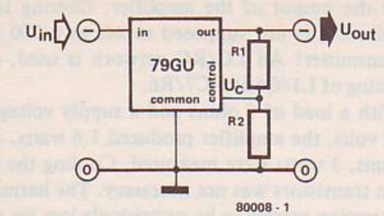

As figure 1 illustrates, the basic regulator circuit using this IC is very simple. The output voltage is determined by the ratio between the two resistors, as follows:

$$U_{out} = \frac{R1 + R2}{R2} \times U_{control}$$

For the 79G, $U_{control}$ is -2.23 V.

As can be seen, the output voltage of this regulator is determined by the voltage on the control input – i.e. that at the R1/R2 junction in figure 1. To be more precise, it is the voltage between the control input and the 'common' connection that sets the output voltage. Knowing this, the actual circuit (figure 2) is not so difficult to understand.

When the motor is loaded, its speed will tend to drop. The current through the motor increases, producing a larger voltage drop across R2. The IC will now try to restore the original voltage difference between the 'control' and 'common' connections, by increasing its output voltage. This, in turn, means that more power is supplied to the motor – counteracting the tendency for the speed to drop.

Basically, this is a feedback system – and positive feedback, at that. For correct operation, the amount of feedback must obviously be set accurately. One solution would be to use a preset potentiometer for R2. This is not very practical, however: where do you find a 4.7 Ω pot that will happily tolerate a current of up to 1 A?

Adding P2 is an infinitely better solution. With its slider turned right up, the circuit becomes identical to that given in figure 1, as far as the regulator is concerned; the voltage across the motor is held constant. As the slider of P2 is turned down, more and more positive feedback is added. With P2 set correctly, the motor speed will remain almost constant, independent of load.

Construction

A suitable printed circuit board is given in figure 3. The only components not mounted on this board are the transformer, fuse, and potentiometer P1.

Having built the circuit, and before connecting the motor, an initial check is advisable. The slider of P2 is turned fully clockwise. Power is then applied, and P1 is set to maximum resist-

Figure 1. In the basic regulator circuit, the IC adjusts the output voltage to maintain a constant −2.23 V between its control input and the 'common' connection. This means that the output voltage is determined by R1 and R2.

Figure 2. The complete circuit. P1 sets the motor speed; preset P2 is adjusted so that the speed remains constant under load. On some drills, a lower value for C2 and/or C3 may give better results. In fact, one of the drills we tried ran best when these capacitors were omitted!

Figure 3. Printed circuit board and component layout. Note that only two connections are provided to P1: the connection between the wiper and one end is made at the potentiometer.

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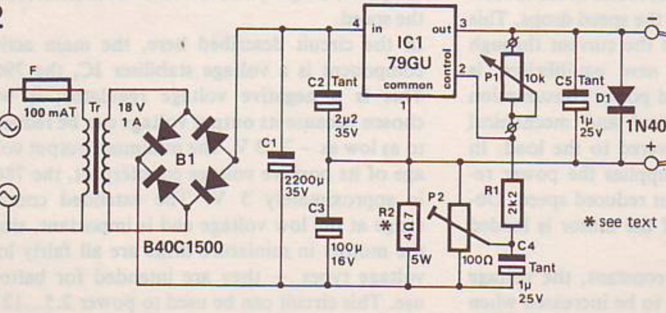

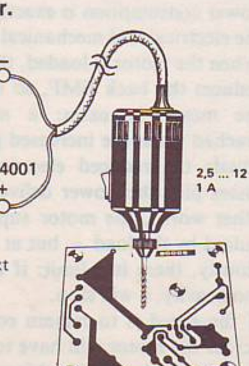

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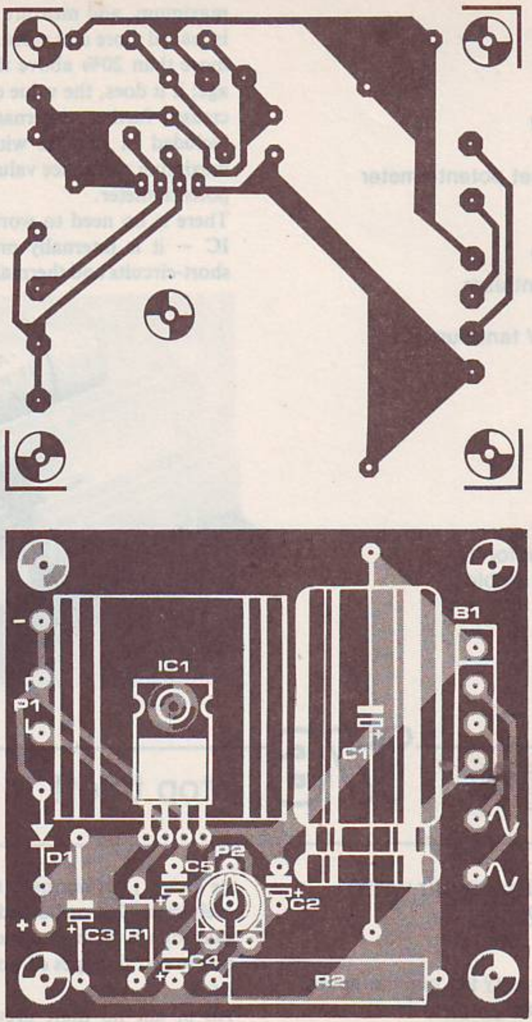


ance – corresponding to the maximum output voltage. This voltage (between the '+' and '–' output terminals) is measured. It should be safely below the maximum permissible motor voltage – say about 20% down. If it is too far off this value, the value of R1 will have to be modified: increasing R1 reduces the voltage, and reducing R1 brings the voltage up.
P1 is now turned back about halfway, and the drill is connected. Preset P2 is carefully adjusted so that the motor speed is on the verge of increasing. The idea is that too much feedback will cause the drill speed to run right up, out of control; too little feedback, on the other hand, will make the circuit less effective. It is possible that, with a given motor, even the lowest setting of P2 is not low enough: the speed still drops when the motor is loaded. In that case, the value of R2 will have to be increased and the calibration procedure repeated.
Obviously, this circuit is no miracle worker. If the motor is loaded further when it is already running flat out, at maximum voltage, the speed will drop. Which is just as well – a higher voltage than the maximum permissible will burn out the motor. This is why it is so important to select the correct value for R1 – it determines the maximum voltage that can be applied to the motor. For that matter, it is a good idea to check this again once P2 has been adjusted: set P1 to

Parts list

Resistors:

R1* = 2k2
R2* = 4.7 Ω /5 W
P1 = 10 k lin.
P2 = 100 Ω preset potentiometer

Capacitors:

C1 = 2200 µ /35 V
C2 = 2µ2/35 V tantalum
C3 = 100 µ /16 V
C4, C5 = 1 µ /25 V tantalum

Semiconductors:

IC1 = 79GU
D1 = 1N4001
B1 = B40C1500

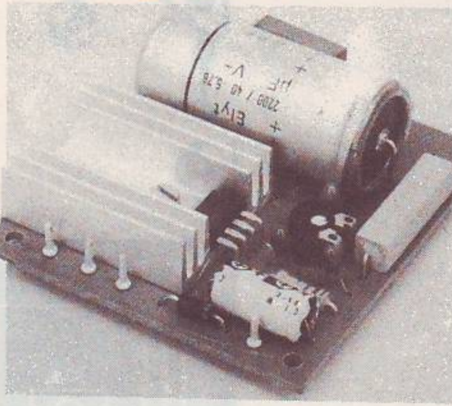
Sundries:

Tr = 18 V/1 A transformer
F = 100 mA fuse, sloblo
Heatsink for IC1

* see text

maximum, and measure the motor voltage as it is loaded more and more. It should not run up to more than 20% above the nominal motor voltage; if it does, the value of R1 will have to be increased further. Alternatively, a resistor can be included in parallel with P1 – reducing the maximum resistance value that can be set by this potentiometer.

There is no need to worry about damaging the IC – it is internally protected against output short-circuits and thermal overload.



176 Stop thief!

who wants to steal a car with engine trouble?

Protecting one's property is a popular hobby – particularly when that property is attractive to others. There are all sorts of systems for protecting cars, but the one described here is unusual: it is deception, rather than protection. It doesn't make it impossible to steal the car (for that matter, no system does), but it makes it very unattractive: who wants to steal a car with an engine that stalls every few yards?

Even the most effective of theft prevention systems normally suffer from the drawback that it is immediately apparent to a thief that some kind of protective device is built in. If he is sufficiently courageous, persistent and experienced, he can put the device out of action and make off with the car. If he's a professional thief, it's 'Goodbye, car!'; if he's joy-riding, you *may* just be lucky... but normally the vehicle ends up severely crumpled or burned-out. An alarm system that sounds the car horn even seems to be attractive to certain types of 'joy-rider'. Amazing, but true.

All in all, no theft prevention system can be guaranteed to keep thieves out of your car. Once

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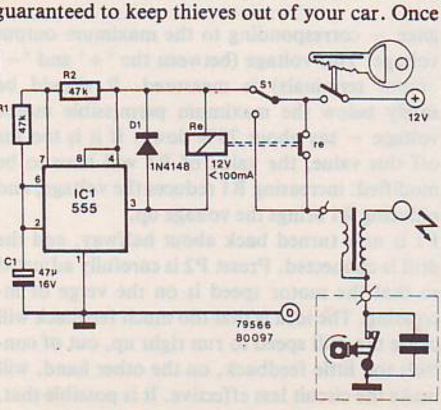

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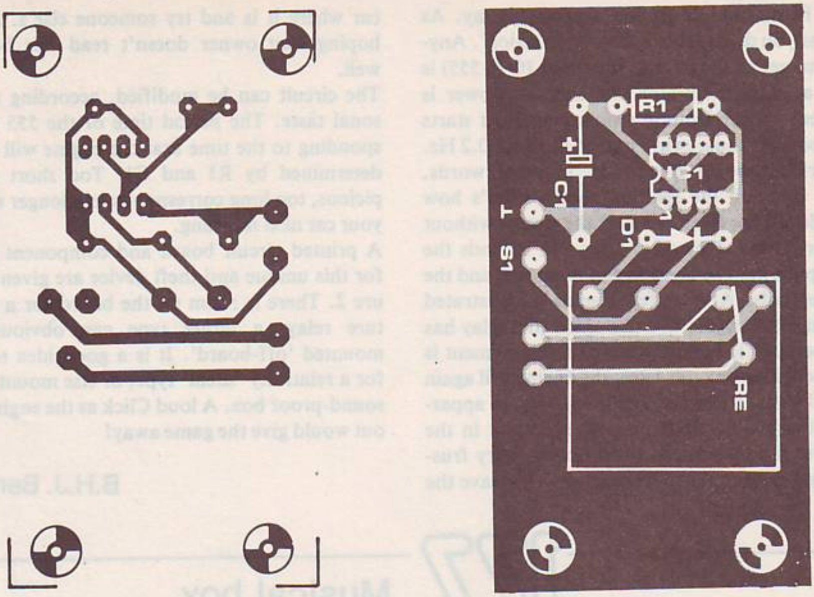

in, the chances that they can get it moving are quite high.
The system described here should give adequate protection against joy-riders. It should even discourage most 'professional' thieves – all except the type who is prepared to steal a furniture removal van first and then run your car into it! At the same time, this system has the advantage that it does the job on its own, without giving silent or raucous alarm signals to the owner or other passers-by. There's no need to chase after your property in the middle of the night, wearing only pyjamas and a dressing gown: you can rest assured that it will never end up far away. It is highly unlikely that the would-be thief will drive more than a few hundred yards. What is the basic principle behind such an effective system? Simple! The engine is about as reliable as that of a twenty-year old car with water in the petrol tank!

In practice

The wire from the positive side of the ignition coil to the 12 V supply from the ignition switch is cut, and passed through the break contact of a relay. As long as the relay is not energised, power is supplied to the ignition coil and the engine runs smoothly. However, when the relay pulls in it breaks the connection. No power to the ignition coil – no spark – no go. One dead engine! It can be started again – no problem – and it will run as smoothly as before. Until the relay pulls in again, that is.
The circuit is given in figure 1. It is put into operation by operating switch S1. A 'secret' switch, of course. Bear in mind that the best place to hide a switch is very often the most obvious:

Figure 1. Only six components – cheap ones, at that! – are needed for a quite effective anti-theft device.

Figure 2. Printed circuit board and component layout. There is room for a miniature relay, but a larger type can be mounted off-board. 'Sound-proofing' is important!

Parts list:

Resistors:
R1 = 82 k

Capacitors:
C1 = 47 µ /16 V

Semiconductors:
IC1 = NE555 or equ.
D1 = 1N4148

Miscellaneous:
S1 = single-pole changeover switch
Re = 12 V/100 mA relay with heavy-duty break contact.

right in the middle of the dashboard, say. As long as you don't lable it 'theft protection'. Anyway, to get to the circuit: the timer IC (a 555) is used as a multivibrator. As soon as power is applied, via the ignition switch and S1, it starts to produce a squarewave output at about 0.2 Hz. A period time of 5 seconds, in other words. After bridging the ignition switch (that's how they do it), the thief can start the engine without any problems. However, after five seconds the relay pulls in. The ignition coil is cut off, and the engine stalls. After a few seonds of frustrated fiddling, the engine will fire again (the relay has dropped out!), but the feeling of achievement is doomed. Five seconds later, the engine will again stall. To sum it up: the engine *will* run, so apparently there is no theft prevention circuit in the car, but it conks out at short notice. Very frustrating for any thief. His best bet is to leave the car where it is and try someone else's. Always hoping *that* owner doesn't read this book as well.

The circuit can be modified, according to personal taste. The period time of the 555 (corresponding to the time that the engine will run) is determined by R1 and C1. Too short is suspicious, too long corresponds to a longer walk to your car next morning.

A printed circuit board and component layout for this unique anti-theft device are given in figure 2. There is room on the board for a miniature relay; a larger type can obviously be mounted 'off-board'. It is a good idea to look for a relatively 'silent' type, or else mount it in a sound-proof box. A loud Click as the engine cuts out would give the game away!

B.H.J. Bennink

177 Musical box

Readers who collect musical boxes will probably think that an 'electronic musical box' sounds as crazy as a gas telephone or a steam radio. After all, what made the musical box so enjoyable was winding it up and listening to its familiar tune. The circuit presented here shows that electronics can be used to replace the wear-prone internal workings of a musical box. In fact, an advantage over its old-fashioned counterpart is that this circuit is able to play no less than 27 tunes. Applications can also include toys, video games and doorbells.

As can be seen from figure 1, the actual melody generator is a single IC (IC4). It is the AY-3-1350 from General Instrument Microelectronics, a company with an excellent name for solid state musical devices. The circuitry around IC4 generates the clock signal, selects the melody required and amplifies the output level.

To select a particular tune, one of the connections marked A...E will have to be grounded and pin 15 of the melodic chip must be connected to one of the points marked 1...4. There are several ways in which the desired code can be presented to the IC. One method is to use wire links, another is to incorporate switches and a combination of the two is also possible. The printed circuit board has been designed to accomodate either of the two methods shown in figure 2.

If the circuit is constructed exactly as shown in figure 1 and wire links are placed between points K...N and R...V (see figure 2a), the following procedure will take place.

When one of the pushbuttons, S_A...S_E is pressed, one of the points marked A...E will be connected to ground via one of the diodes D1...D5. Each pushbutton has a total of five melodies at its disposal. The choice can be cut down to one by means of a wire link. Thus, one of five predetermined melodies can be selected per switch and, in addition, two well-known chimes may be 'played' by depressing S_F or S_G. Table one shows the melodies which are available and the combination of connections required to select each one. The code numbers and letters correspond to those given in the circuit and in the component layout shown in figure 3.

The second method is to use a pair of multi-way

Figure 1. The complete circuit of the 'electronic musical box'.

Figure 2. Using wire links, as shown in figure 2a, five melodies can be pre-selected. If this is considered too much of a restriction, two five-way switches can be used, as shown in figure 2b.

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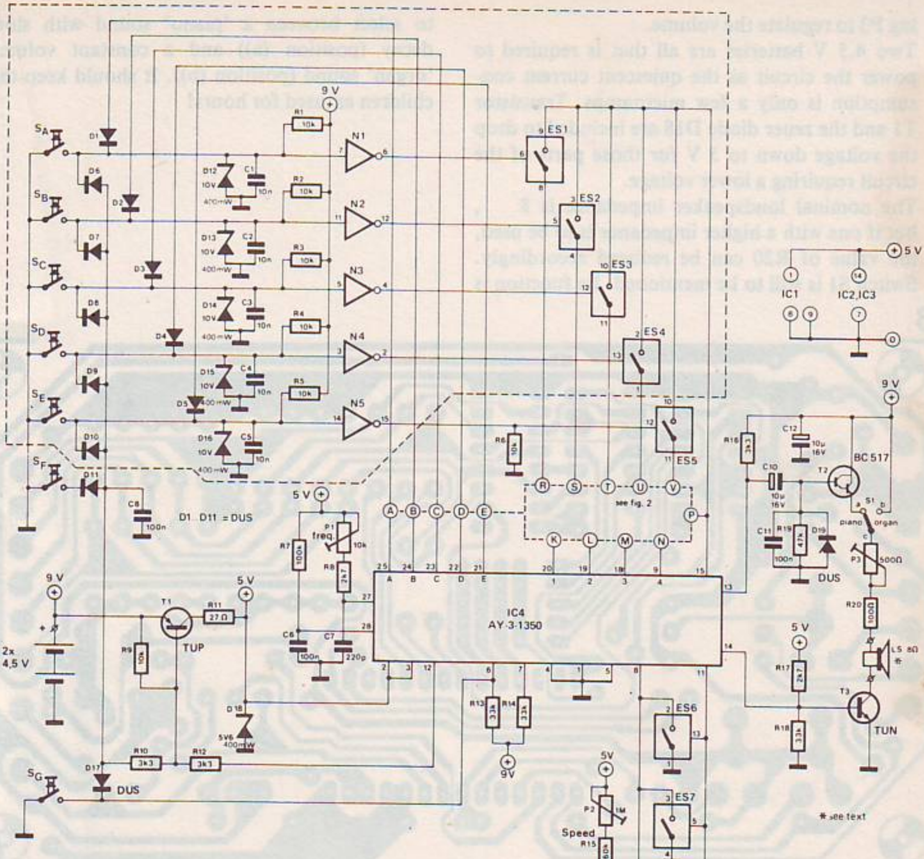

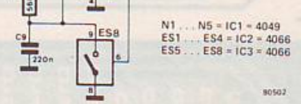

switches, in which case the area inside the dotted line in figure 1 may be omitted. This will enable any one of 25 melodies to be selected. As can be seen from figure 2b, points A...E can be grounded by means of a six position wafer switch, S3. Switch S2 connects one of the points K...N to point P. The melody will be initiated upon depressing S_F. Resistor R6 and the electronic switch ES5 are not necessary for this latter option. They are shown outside the dotted line as ES5 is contained in a separate IC to ES1... ES4.

The oscillator is formed by C7, R8 and P1 together with part of IC4. The pitch of the melody being played can be adjusted by P1, the length of each note can be adjusted by P2, leav-

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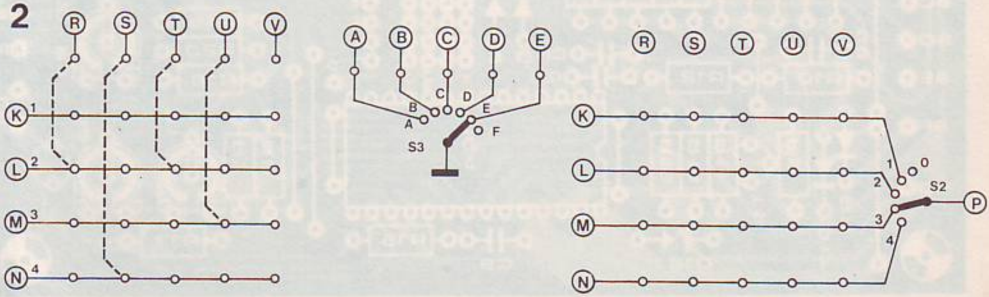
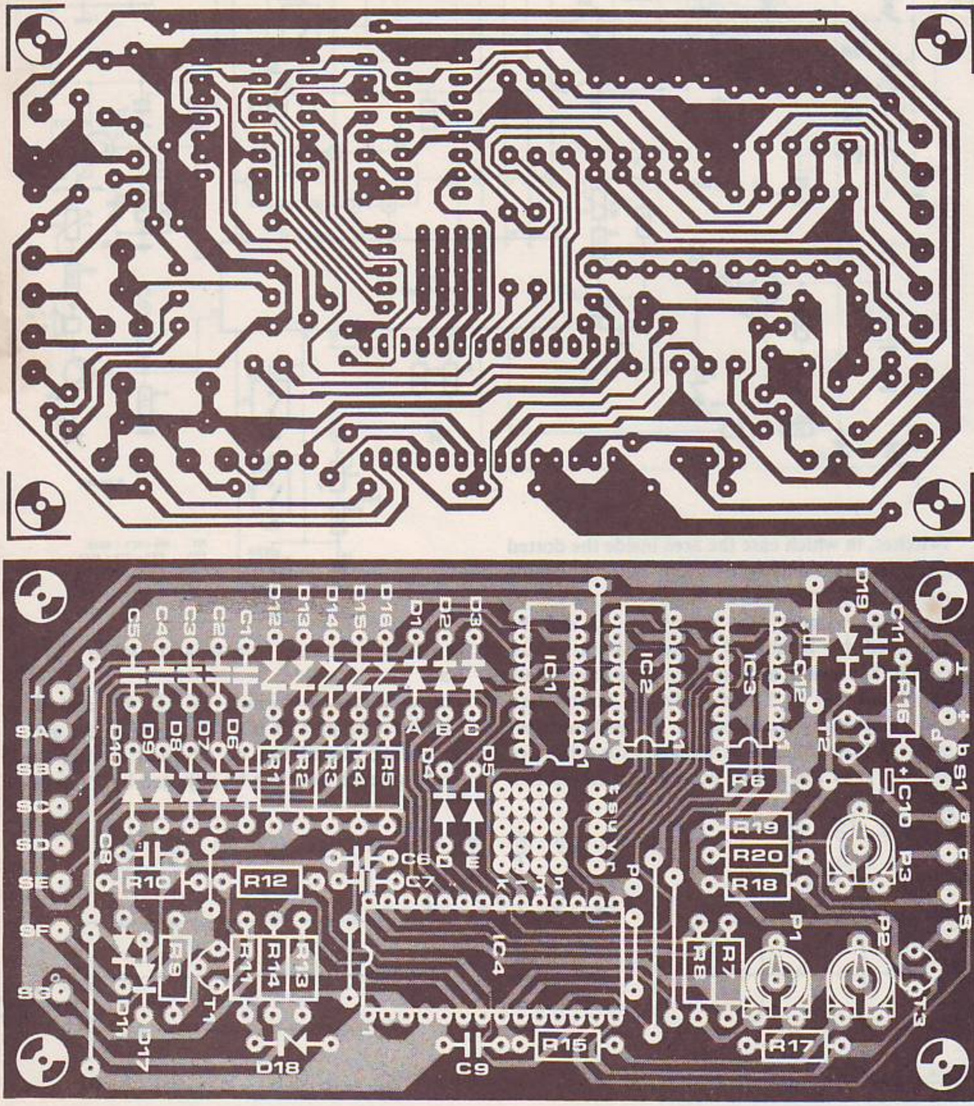


ing P3 to regulate the volume.
Two 4.5 V batteries are all that is required to power the circuit as the quiescent current consumption is only a few microamps. Transistor T1 and the zener diode D18 are included to drop the voltage down to 5 V for those parts of the circuit requiring a lower voltage.
The nominal loudspeaker impedance is 8 , but if one with a higher impedance is to be used, the value of R20 can be reduced accordingly. Switch S1 is still to be mentioned. Its function is to select between a 'piano' sound with slow decay (position (a)) and a constant volume 'organ' sound (position (b)). It should keep the children amused for hours!

3



Parts list

Resistors:
R1...R6, R9 = 10 k
R7 = 100 k
R8, R17 = 2k7
R10, R12, R16 = 3k3
R11 = 27 Ω
R13, R14, R18 = 33 k
R15 = 560 k
R19 = 47 k
R20 = 100 Ω
P1 = 10 k preset
P2 = 1 M preset
P3 = 500 Ω preset

Capacitors:
C1...C5 = 10 n
C6, C8, C11 = 100 n
C7 = 220 p
C9 = 220 n
C10, C12 = 10 Ω/16 V

Semiconductors:
D1...D11, D17, D19 = DUS
D12...D16 = 10 V/400 mW zener
D18 = 5V6/400 mW zener
T1 = TUP
T2 = BC 517
T3 = TUN
IC1 = 4049
IC2, IC3 = 4066
IC4 = AY-3-1350

Miscellaneous:
S_A...S_G = pushbutton switch
S1 = S.P.D.T.
S2 = 5 position wafer switch
S3 = 6 position wafer switch
LS = 8 Ω /0.5 W loudspeaker (see text)

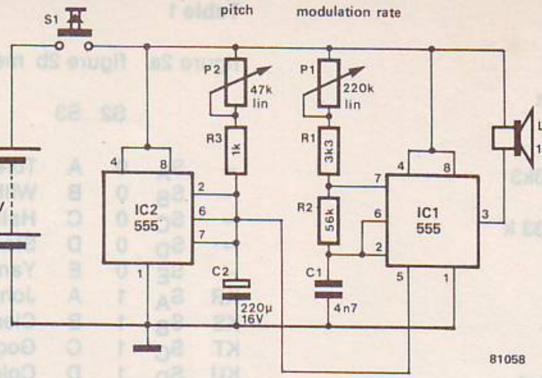
Table 1

figure 2a	figure 2b	S2	S3	melody
–	S_A	0	A	Toreador
–	S_B	0	B	William Tell
–	S_C	0	C	Hallelujah Chorus
–	S_D	0	D	Star Spangled Banner
–	S_E	0	E	Yankee Doodle
KR	S_A	1	A	John Brown's Body
KS	S_B	1	B	Clementine
KT	S_C	1	C	God Save The Queen
KU	S_D	1	D	Colonel Bogey
KV	S_E	1	E	Marseillaise
LR	S_A	2	A	America, America
LS	S_B	2	B	Deutschland Lied
LT	S_C	2	C	Wedding March
LU	S_D	2	D	Beethoven's 5th
LV	S_E	2	E	Augustine
MR	S_A	3	A	A Sole Mio
MS	S_B	3	B	Santa Lucia
MT	S_C	3	C	The End
MU	S_D	3	D	Blue Danube
MV	S_E	3	E	Brahm's Lullaby
NR	S_A	4	A	Hell's Bells (specially composed)
NS	S_B	4	B	Jingle Bells
NT	S_C	4	C	La Vie en Rose
NU	S_D	4	D	Star Wars
NV	S_E	4	E	Beethoven's 9th
	S_F			Descending Octave Chime
	S_G			Westminster Chime

178 Din can

The number of possible applications for noise emitting circuits is absolutely astounding – far too many to be listed in the available space. This is particularly true of the unit described here. Although not exactly original in concept, the can is capable of producing a din which will almost certainly leave your ears tingling for quite a while.

The circuit consists of very few components which together form a 'Kojak' type siren. The majority of the work is done by two (555) timer ICs. The first (IC1) generates an audible tone whose frequency can be adjusted by means of potentiometer P1. The output of this timer is fed directly to a loudspeaker. The loudspeaker however does not emit a constant tone, as IC1 is



modulated by a low frequency sawtooth waveform generated by IC2. The frequency of the sawtooth can be regulated by means of potentiometer P2. As a result, the complete circuit generates a frequency modulated signal which sounds just like a siren. The pitch can be adjusted with P1 and the modulation rate with P2.
The circuit is compact enough to be mounted in a can together with the batteries and loudspeaker and can be used quite successfully on children's cycles, skate boards etc. or as an 'anti-attack' alarm.

Figure 1. IC1 generates a tone which is modulated in frequency by IC2.

L. van Ginderen

179 Transistor curve tracer

I_C/U_{CE} characteristics directly onto the screen

There are never enough simple circuits which provide useful and low-cost additions to the 'home lab'. This particular design possesses all the advantages to make it a glorious example of its kind. It offers oscilloscope owners a neat, additional measurement facility. It is easy to build, contains common parts and is inexpensive. Reason enough to design a printed circuit board for it.

The design

This is an economic curve tracer for transistors and diodes. No really professional test instru-

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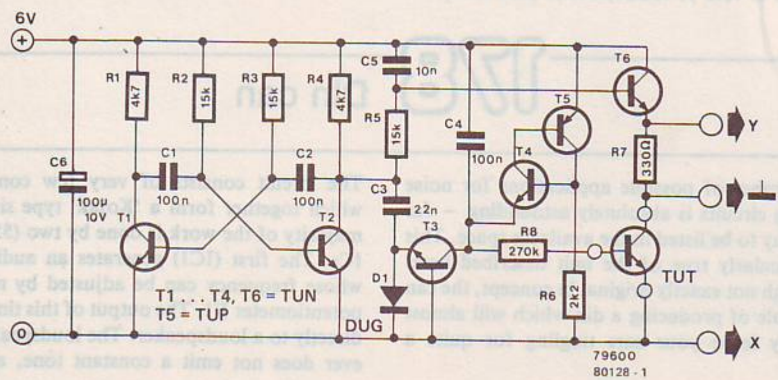

ment, of course, but an extremely useful aid to quickly carry out a general test either to compare transistors or select them. Naturally, hobbyists will have to have an oscilloscope (with separate x and y inputs), because the curves will be displayed on the oscilloscope screen.

Since it is impossible to tell which transistor characteristic is more important than another, there is no such thing as the 'most important curve'. Transistor handbooks speak of the most read curve. This involves the I_C/U_{CE} characteristics where the collector current is plotted as a function of the collector/emitter voltage at different drive currents. Figure 2 gives an example of such a characteristic. At the same time it (roughly) indicates the drive currents the curve tracer uses. The current amplification may be directly derived from the I_C/U_{CE} characteristics and, after a few calculations, so may the transistor's output impedance. The latter is affected by the curve's slope. Generally speaking, the more horizontal and straight it is, the higher the collector/emitter impedance.

Back to the schematic. The transistor under test is indicated as 'TUT' as usual. Between the points which are connected to the Y input and the ground connection of the oscilloscope 'hangs' resistor R7. This is the TUT's collector resistor and the voltage across it is naturally proportional to the collector current of the transistor tested. In this way, an 'I_C' will appear on the vertical axis of the oscilloscope. The TUT's emitter is connected to the X input so that the collector/emitter voltage (U_{CE}) can be read horizontally on the screen.

What causes the curves to appear on the screen? Two signals are fed to the TUT. A 5 step position staircase waveform is fed to the base and during each step a sawtooth is fed to the collector. This means the collector voltage changes continually at a certain base drive current. This occurs at quite a speed so that the oscilloscope screen simultaneously shows 5 characteristics for 5 different base drive currents. The staircase signal and the sawtooth waveform are controlled by means of an astable multivibrator. The AMV consists of T1 and T2 and generates a square wave with a frequency of approximately 1 kHz.

Figure 1. The circuit diagram of the curve tracer.

Figure 2. I_C/U_{CE} curves of a transistor. In our circuit five different base drives are measured.

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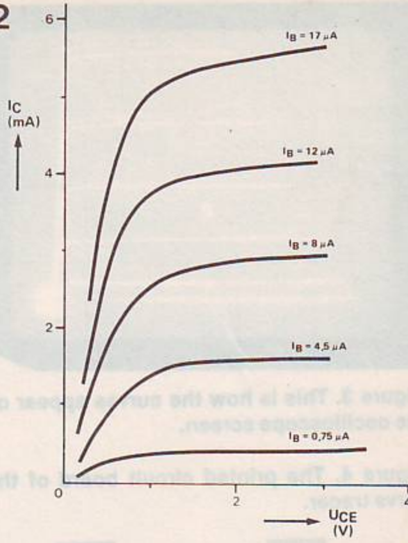

The sawtooth is obtained very easily by integrating the square wave via R5 and C5. Creating the staircase voltage is a little more complicated. During the positive half-cycle of the square wave produced by the AMV, C3 is charged to a maximum which is equal to the supply voltage. During the negative half-cycle, C3 will turn on transistor T3 and thus the voltage at T4's emitter (connected to the TUT's base via R8) will become a little lower. By loading C4 intermittently, each successive negative half-cycle will reduce the emitter voltage of T4 in steps until T4 starts to conduct turning on T5. C4 is soon discharged and a new cycle starts.

The number of stages which make up a single cycle is determined by the ratio of C3 to C4 and is 5 here. By adjusting the value of C4 the number of stages (and thus the number of curves indicated on the screen) can be changed as required.

In practice

The photo in figure 3 shows how the curves appear on the oscilloscope screen. The circuit's only flaw now comes to light – the characteristics are traced from right to left, instead of the other way around. Unfortunately, nothing can be done about this. In practice it does not present a problem. What is serious however, is that the tracer is only suitable for NPN transistors. NPN types cannot be tested with it. If this is considered to be a drawback, however, there is a cheap solution: two printed circuit boards may

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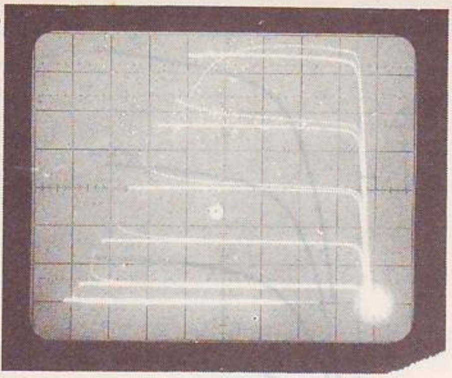

Figure 3. This is how the curves appear on the oscilloscope screen.

Figure 4. The printed circuit board of the curve tracer.

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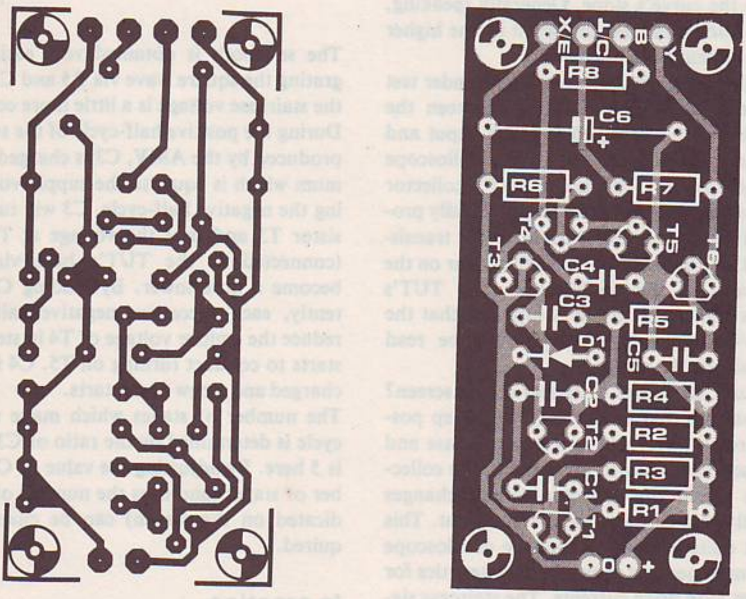


Parts list

Resistors:
R1, R4 = 4k7
R2, R3, R5 = 15 k
R6 = 2k2
R7 = 330 Ω
R8 = 270 k

Capacitors:
C1, C2, C4 = 100 n
C3 = 22 n
C5 = 10 n
C6 = 100 µ /10 V

Semiconductors:
T1...T4, T6 = TUN
T5 = TUP
D1 = DUG

be built instead of one. The circuit requires few components, so why not? The second circuit will then be a PNP version. For T1...T4 and T6 use TUPs, T5 will be a TUN, C6, D1 and the supply leads will be switched around. Furthermore, such a PNP version will trace the curves from left to right, only now the Y axis will be negative so that they will appear upside down on the screen. A little strange perhaps, but you'll soon get used to it...

As mentioned above, diodes may also be tested. These are connected with the anode to R7 (⊥) and the cathode to the supply zero (X). The I/U characteristics of the diode in question will now appear on the screen. Figure 4 shows the printed circuit board. It is highly compact and can be built in less than no time.

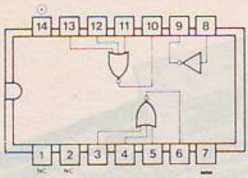
Last word. Since the circuit only requires a few mA, the supply will not have to be very 'heavily' tested. However, the supply voltage must be well regulated for it to work properly.

B. Darnton

BLANK

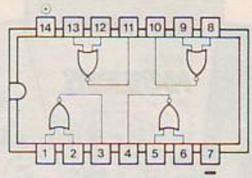
DUAL 3 INPUT NOR GATE PLUS INVERTER

4000



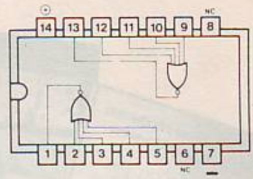
QUADRUPLE 2 INPUT NOR GATE

4001



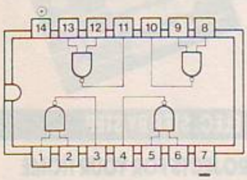
DUAL 4 INPUT NOR GATE

4002



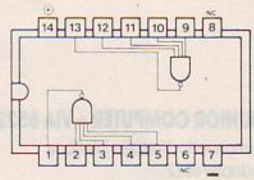
QUADRUPLE 2 INPUT NAND GATE

4011



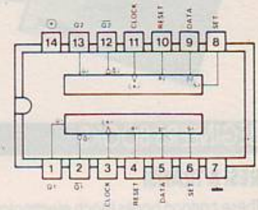
DUAL 4 INPUT NAND GATE

4012



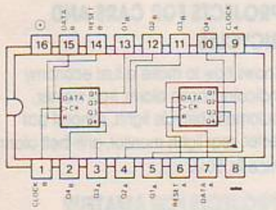
DUAL D FLIP FLOP

4013



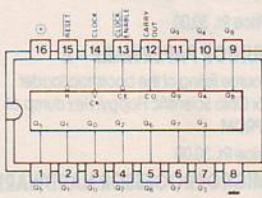
DUAL 4 BIT STATIC SHIFT REGISTER

4015



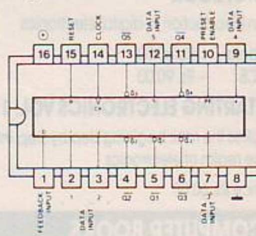
DIVIDE BY 10 SYNCHRONOUS COUNTER

4017



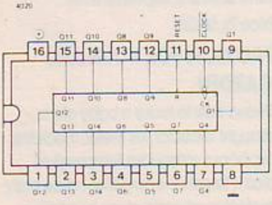
SYNCHRONOUS PRESETTABLE DIVIDE BY 'N' COUNTER

4018



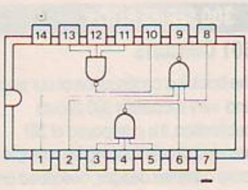
14 BIT BINARY RIPPLE COUNTER

4020



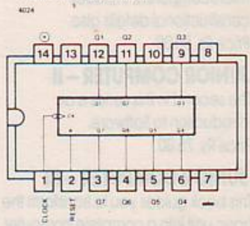
TRIPLE 3 INPUT NAND GATE

4023



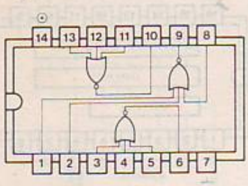
7 STAGE BINARY RIPPLE COUNTER

4024



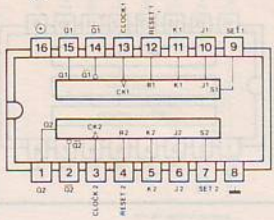
TRIPLE 3 INPUT NOR GATE

4025



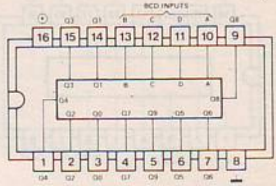
DUAL JK FLIP FLOP

4027



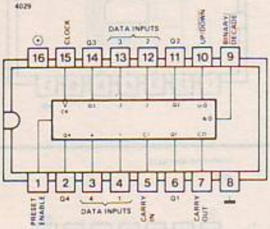
BCD TO DECIMAL DECODER

4028



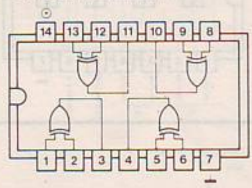
SYNCHRONOUS PRESET TABLE BINARY/DECADE UP/DOWN COUNTER

4029

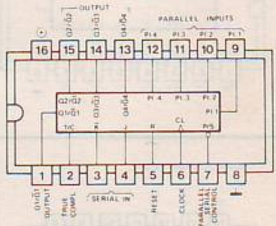


QUADRUPLE 2 INPUT EXCLUSIVE OR GATES

4030
4070 low power TTL compatible (fan out - 2)

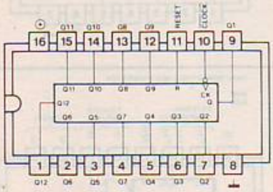


4035
4 BIT
PARALLEL IN/PARALLEL OUT
SHIFT REGISTER



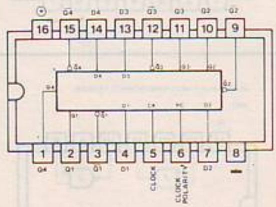
12 BIT BINARY RIPPLE COUNTER

4040



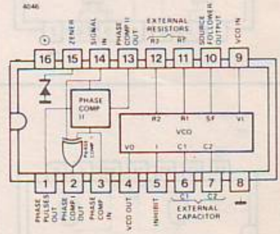
QUAD CLOCKED "D" LATCH

4042



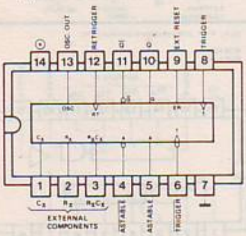
MICROPOWER PLL

4046



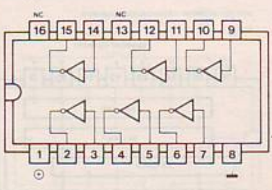
MONOSTABLE/ASTABLE MULTIVIBRATOR

4047



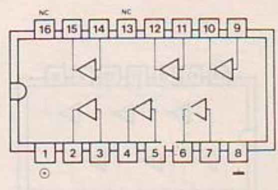
HEX INVERTING BUFFER

4049



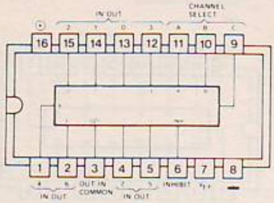
HEX BUFFER

4050



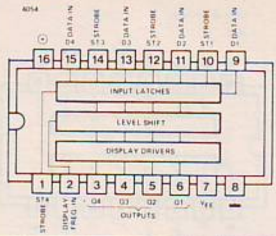
8 CHANNEL ANALOGUE MULTIPLEXER DEMULTIPLEXER

4051



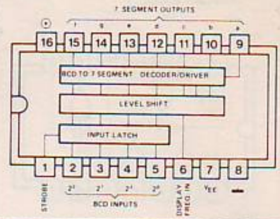
LCD DRIVER

4054



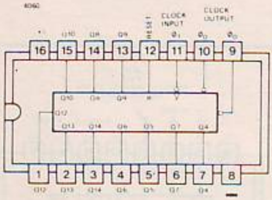
BCD TO 7 SEGMENT DECODER/DRIVER

4056



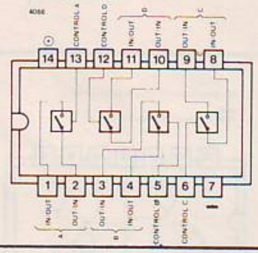
7 · BIT BINARY RIPPLE COUNTER AND OSCILLATOR

4060



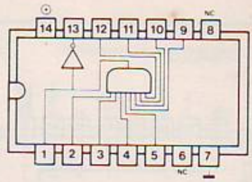
QUAD BILATERAL SWITCH

4066



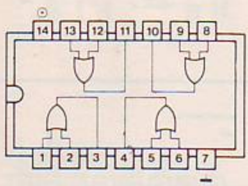
8 INPUT AND/NAND GATE

4068



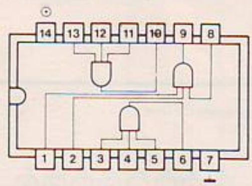
QUADRUPLE 2 INPUT OR GATE

4071



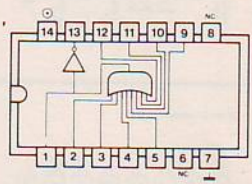
TRIPLE 3 INPUT AND GATE

4073



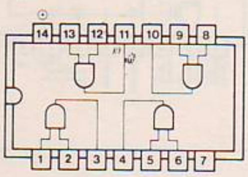
8 INPUT OR/NOR GATE

4078



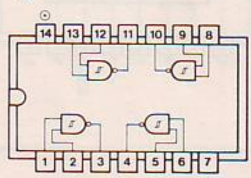
QUADRUPLE 2 INPUT AND GATE

4081



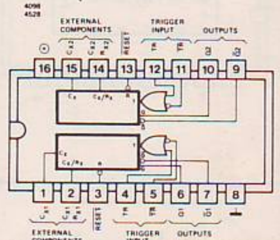
QUADRUPLE 2 INPUT NAND SCHMITT TRIGGER

4093



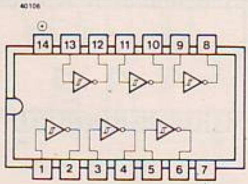
DUAL MONOSTABLE MULTIVIBRATOR

4098
4528



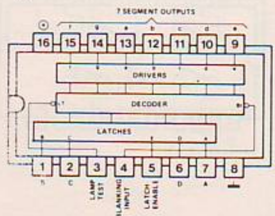
HEX SCHMITT TRIGGER

40106



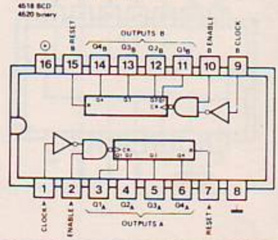
BCD TO 7 SEGMENT LATCH/DECODER/DRIVER

4511



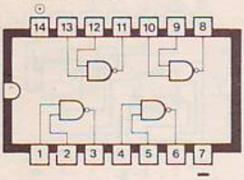
DUAL 4 BIT SYNCHRONOUS UP COUNTERS

4518 BCD
4520 binary



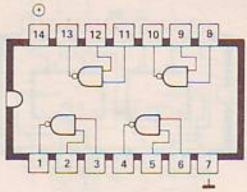
QUADRUPLE 2 INPUT NAND GATES

7400
7403 open collector outputs
7437 power driver (fan out = 30)



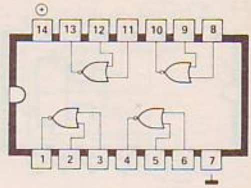
QUADRUPLE 2 INPUT NAND GATE WITH OPEN COLLECTOR OUTPUT

7401



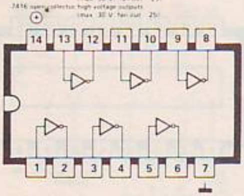
QUADRUPLE 2 INPUT NOR GATES

7402
7428 power driver (fan out = 30)



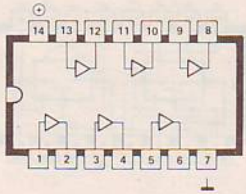
HEX INVERTERS

7404
7405 open collector outputs
7406 open collector high voltage outputs (max. 30 V; fan out = 25)
7416 open collector high voltage outputs (max. 30 V; fan out = 25)



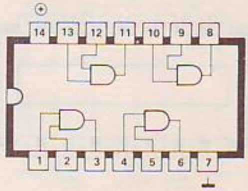
HEX BUFFER/DRIVER WITH OPEN COLLECTOR HIGH VOLTAGE OUTPUTS (max. 30 V; fan out = 25)

7407



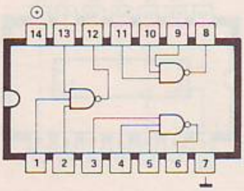
QUADRUPLE 2 INPUT AND GATES

7408
7409 open collector outputs



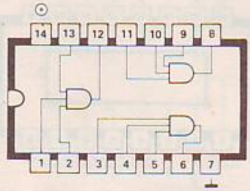
TRIPLE 3 INPUT NAND GATES

7410
7412 open collector outputs



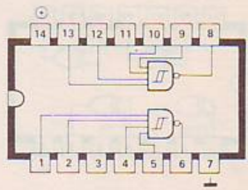
TRIPLE 3 INPUT AND GATE

7411



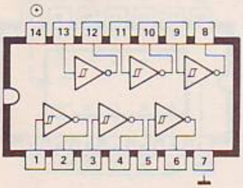
DUAL 4 INPUT NAND SCHMITT TRIGGER

7413



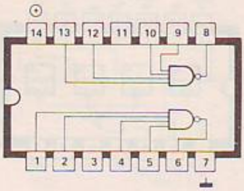
HEX SCHMITT TRIGGER INVERTER

7414



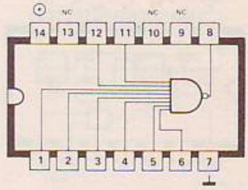
DUAL 4 INPUT NAND GATES

7420
7440 power driver (fan out = 30)



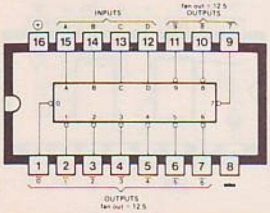
8 INPUT NAND GATE

7430



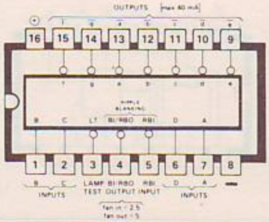
BCD TO DECIMAL DECODER/DRIVER WITH OPEN COLLECTOR OUTPUTS (max. 30 V)

7445



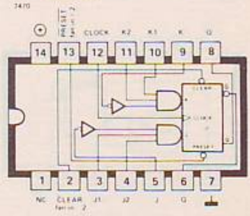
BCD TO 7 SEGMENT DECODER/DRIVER

7447



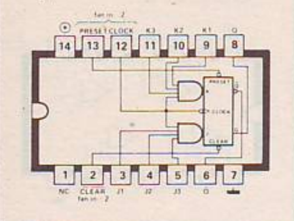
AND GATED J K POSITIVE EDGE TRIGGERED FLIP FLOP WITH PRESET AND CLEAR

7470



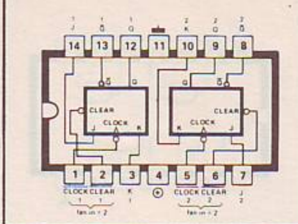
AND GATED J-K FLIP FLOP WITH PRESET AND CLEAR

7472



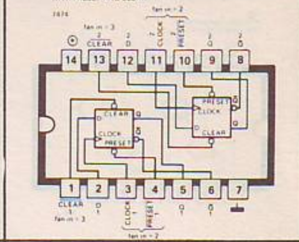
DUAL J-K FLIP FLOP WITH CLEAR

7473



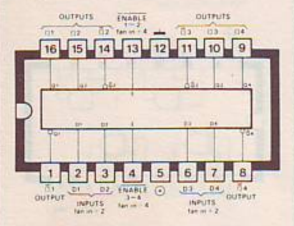
DUAL D TYPE POSITIVE EDGE TRIGGERED FLIP FLOP WITH PRESET AND CLEAR

7474



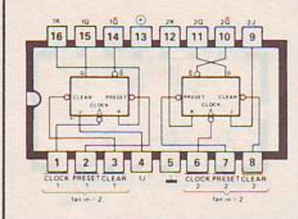
4 BIT BISTABLE LATCH

7475



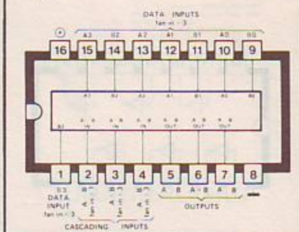
DUAL J-K MASTER SLAVE FLIP FLOP WITH PRESET AND CLEAR

7476



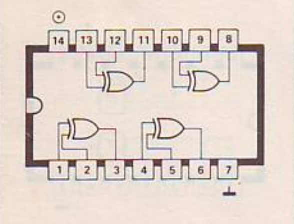
4 BIT COMPARATOR

7485



QUADRUPLE 2 INPUT EXCLUSIVE OR GATE

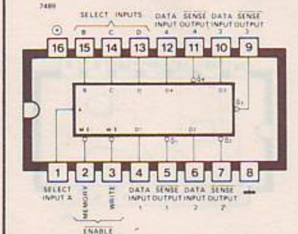
7486



64 BIT READ/WRITE MEMORY

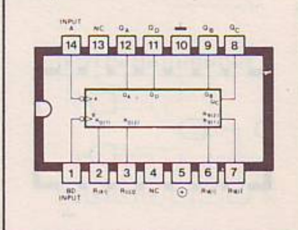
(fan out = 7.5)

7489



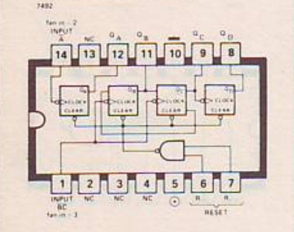
DECADE COUNTER

7490



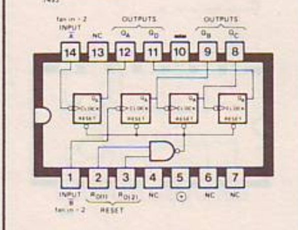
DIVIDE BY TWELVE COUNTER (÷ 2 and ÷ 6)

7492



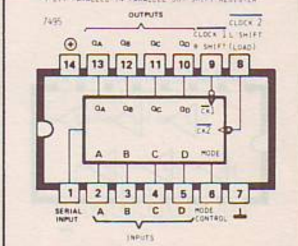
4 BIT BINARY COUNTER

7493



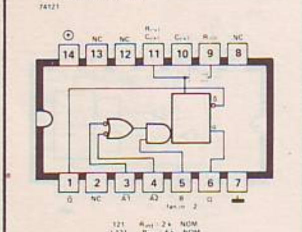
4-BIT PARALLEL-IN PARALLEL-OUT SHIFT REGISTER

7495



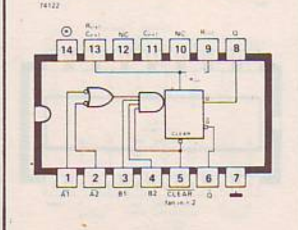
MONOSTABLE MULTIVIBRATOR

74121



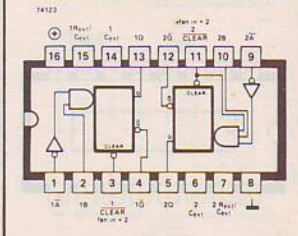
RETRIGGERABLE MONOSTABLE MULTIVIBRATOR WITH CLEAR

74122



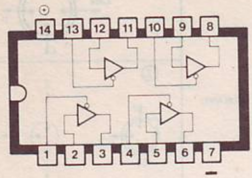
DUAL RETRIGGERABLE MONOSTABLE MULTIVIBRATOR WITH CLEAR

74123



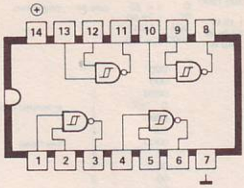
QUAD BUFFER (3 STATE)

74125



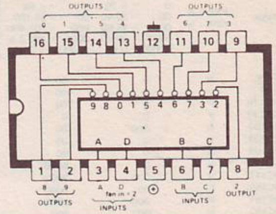
QUADRUPLE 2 INPUT NAND SCHMITT TRIGGER

74132



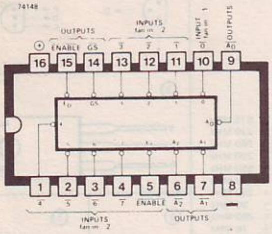
BCD TO DECIMAL DECODER DRIVER (outputs max 60 V max 7 mA)

74141



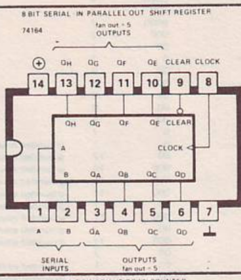
PRIORITY ENCODER

74148



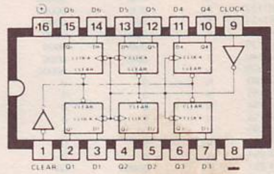
8 BIT SERIAL IN PARALLEL OUT SHIFT REGISTER

74164



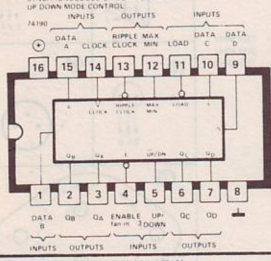
HEX D FLIP FLOP WITH CLEAR

74174



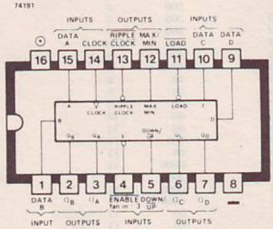
SYNCHRONOUS BCD UP DOWN COUNTER WITH UP DOWN MODE CONTROL

74190



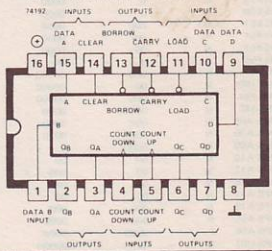
SYNCHRONOUS 4 BIT BINARY UP/DOWN COUNTER

74191



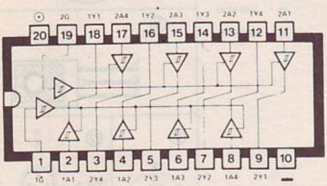
SYNCHRONOUS UP DOWN DECADE COUNTER

74192



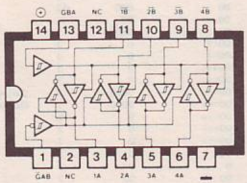
OCTAL BUFFER AND LINE DRIVER (3 STATE)

74LS241



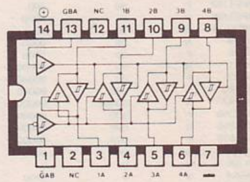
QUADRUPLE BUS TRANSCEIVER (3 STATE)

74LS242



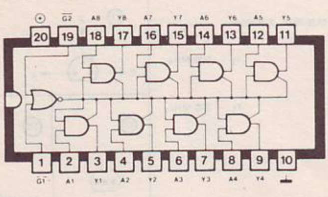
QUADRUPLE BUS TRANSCEIVER (3 STATE)

74LS243



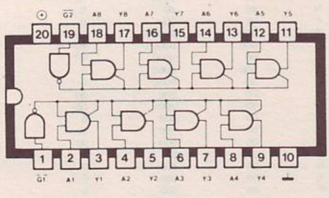
OCTAL BUFFER (3 STATE)

81LS95

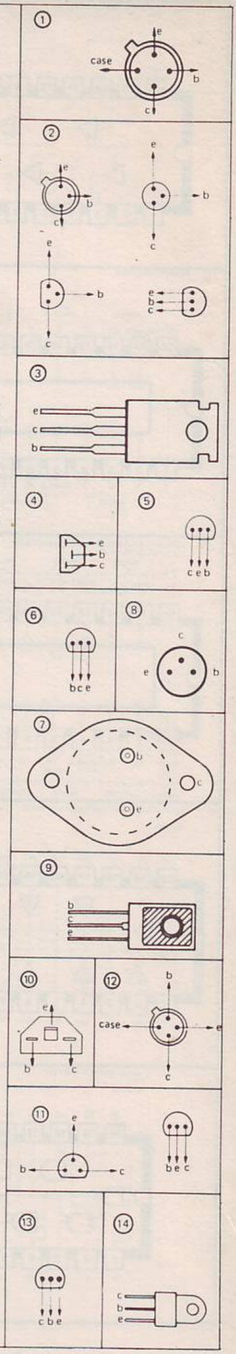


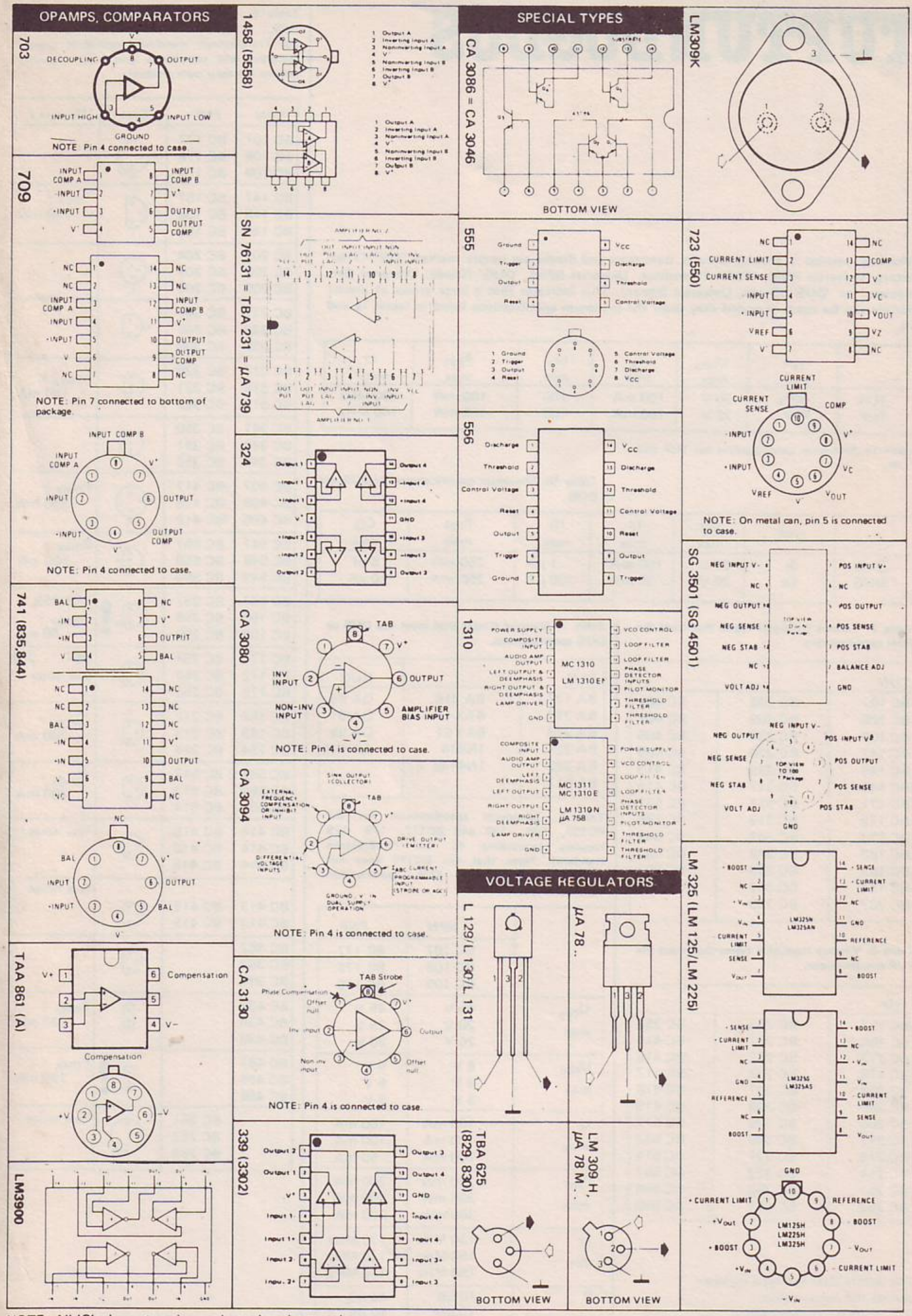
OCTAL BUFFER (3 STATE)

81LS97



Type	PNP = P NPN = N	U_{CEO} (Volt) 0 = ≤ 20 00 = 25-40 000 = 45-60 0000 = 65-80 00000 = ≥ 85	$I_{c(max)}$ (mA) 0 = ≤ 50 00 = 55-100 000 = 105-400 0000 = 405-2 A 00000 = ≥ 2 A	P_{max} (mW) not cooled: 0 = ≤ 300 00 = 305-1000 cooled: 00☆ = 1-10 W 00☆☆ = 10-35 W 00☆☆☆ = ≥ 40 W	$h_{FE(min)}$ 0 = ≤ 20 00 = 25-50 000 = 55-120 0000 = ≥ 125	case nr.	comments
TUN	N	0	00	0	000		
TUP	P	0	00	0	000		
AC126	P	0	00	00	0000	2	
AF239	P	0	0	0	0	1	grounded base: f f_T = 700 MHz
BC107	N	000	00	0	000	2	
BC108	N	0	00	0	000	2	
BC109	N	0	00	0	0000	2	low noise
BC140	N	00	0000	00☆	00	2	
BC141	N	000	0000	00☆	00	2	
BC160	P	00	0000	00☆	00	2	
BC161	P	000	0000	00☆	00	2	
BC182	N	000	000	0	0000	2	
BC212	P	000	000	0	000	2	
BC546	N	0000	00	00	0000	2	
BC556	P	0000	00	00	000	2	
BD106	N	00	00000	00☆☆	00	7	
BD130	N	000	00000	00☆☆☆	0	7	
BD132	P	000	00000	00☆☆	00	9	
BD137	N	000	0000	00☆	00	9	
BD138	P	000	0000	00☆	00	9	
BD139	N	0000	0000	00☆	00	9	
BD140	P	0000	0000	00☆	00	9	
BDY20	N	000	00000	00☆☆☆	0	7	
BF180	N	0	0	0	0	1	grounded base: f_T = 675 MHz
BF185	N	0	0	0	00	12	grounded base: f_T = 220 MHz
BF194	N	0	0	0	000	10	grounded emitter: f_T = 260 MHz
BF195	N	0	0	0	000	10	grounded emitter: f_T = 200 MHz
BF199	N	00	0	00	000	11	grounded emitter: f_T = 550 MHz
BF200	N	0	0	0	00	1	grounded base: f_T = 240 MHz
BF254	N	00	0	0	000	11	grounded emitter: f_T = 260 MHz
BF257	P	00000	00	00	00	2	grounded emitter: f_T = 90 MHz
BF494	N	0	0	0	000	11	grounded emitter: f_T = 260 MHz
BFX34	N	000	00000	00	00	2	f_T = 70 MHz
BFX89	N	0	0	0	00	1	grounded emitter: f_T = 1000 MHz
BFY90	N	0	0	0	00	1	grounded emitter: f_T = 1000 MHz
BSX19	N	0	0000	0	000	2	
BSX20	N	0	0000	0	000	2	
BSX61	N	000	0000	00	000	2	
HEP51	P	00	0000	00	000	1	f_T = 150 MHz
HEP53	N	00	0000	00	000	1	f_T = 200 MHz
HEP56	N	0	00	00	000	5	f_T = 750 MHz
MJE171	P	000	00000	00☆☆	00	9	
MJE180	N	00	00000	00☆☆	00	9	
MJE181	N	000	00000	00☆☆	00	9	
MJE340	N	00000	0000	00☆☆	00	9	
MPS A05	N	000	0000	00	00	13	
MPS A06	N	0000	0000	00	00	13	
MPS A09	N	0000	0	00	000	13	
MPS A10	N	00	00	00	00	13	
MPS A13	N	00	000	00	0000	13	
MPS A16	N	00	00	00	0000	13	
MPS A17	N	00	00	00	0000	13	
MPS A18	N	000	000	00	0000	13	
MPS A55	P	000	0000	0	00	13	
MPS A56	P	0000	0000	0	00	13	
MPS U01	N	00	00000	00☆	00	14	
MPS U05	N	000	00000	00☆	00	14	
MPS U56	P	0000	00000	00☆	00	14	
MPS2926	N	0	00	00	00	13	f_T = 300 MHz
MPS3394	N	00	00	00	000	13	
MPS3702	P	00	000	00	000	13	f_T = 100 MHz
MPS3706	N	0	0000	00	00	13	
MPS6514	N	00	00	0	0000	13	f_T = 480 MHz
TIP29	N	00	0000	00☆☆	0	3	
TIP30	P	00	0000	00☆☆	0	3	
TIP31	N	00	00000	00☆☆☆	0	3	
TIP32	P	00	00000	00☆☆☆	0	3	
TIP140	N	000	00000	00☆☆☆	0000	7	Darlington
TIP142	N	00000	00000	00☆☆☆	0000	7	Darlington
TIP2955	P	000	00000	00☆☆☆	0	3	
TIP3055	N	000	00000	00☆☆☆	0	3	
TIP5530	P	000	00000	00☆☆☆	0	3	
2N696	N	000	0000	00	0	2	
2N706	N	0	0	0	0	2	
2N914	N	0	0000	00	00	2	
2N1613	N	000	0000	00	00	2	
2N1711	N	000	0000	00	000	2	
2N1983	N	00	0000	00	000	2	
2N1984	N	00	0000	00		2	
2N2219	N	00	0000	00	00	2	
2N2222	N	00	0000	00	00	2	
2N2925	N	00	00	0	0000	13	
2N2955	P	00	00	0	0	2	≠ MJE2955, TIP2955!
2N3054	N	000	00000	00☆☆☆	00	7	
2N3055	N	000	00000	00☆☆☆	0	7	
2N3553	N	00	0000	00☆	0	2	f_T = 500 MHz
2N3568	N	000	0000	0	000	13	
2N3638	P	00	0000	0	000	13	
2N3702	P	00	000	00	000	13	
2N3866	N	00	000	00☆	0	2	f_T = 700 MHz
2N3904	N	00	000	0	00	13	
2N3905	P	00	000	00	000	13	
2N3906	P	00	000	00	000	13	
2N3907	N	000	0	0	000	13	
2N4123	N	00	000	0	00	13	
2N4124	N	00	000	0	000	13	
2N4126	P	00	000	0	000	13	
2N4401	N	00	0000	00	0	13	
2N4410	N	0000	000	00	000	13	
2N4427	N	0	000	00☆	0	2	f_T = 700 MHz
2N5183	N	0	0000	00	000	2	





NOTE: All IC's shown top view, unless otherwise stated.

TUPTUNDUGDUS

Wherever possible in Elektor circuits, transistors and diodes are simply marked 'TUP' (Transistors, Universal PNP), 'TUN' (Transistor, Universel NPN), 'DUG' (Diode, Universal Germanium) or 'DUS' (Diode, Universal Silicon). This indicates that a large group of similar devices can be used, provided they meet the minimum specifications listed in tables 1a and 1b.

	type	U_{ceo} max	I_c max	h_{fe} min.	P_{tot} max	f_T min.
TUN	NPN	20 V	100 mA	100	100 mW	100 MHz
TUP	PNP	20 V	100 mA	100	100 mW	100 MHz

Table 1a. Minimum specifications for TUP and TUN.

Table 1b. Minimum specifications for DUS and DUG.

	type	U_R max	I_F max	I_R max	P_{tot} max	C_D max
DUS	Si	25 V	100 mA	1 μA	250 mW	5 pF
DUG	Ge	20 V	35 mA	100 μA	250 mW	10 pF

Table 2. Various transistor types that meet the TUN specifications.

TUN		
BC 107	BC 208	BC 384
BC 108	BC 209	BC 407
BC 109	BC 237	BC 408
BC 147	BC 238	BC 409
BC 148	BC 239	BC 413
BC 149	BC 317	BC 414
BC 171	BC 318	BC 547
BC 172	BC 319	BC 548
BC 173	BC 347	BC 549
BC 182	BC 348	BC 582
BC 183	BC 349	BC 583
BC 184	BC 382	BC 584
BC 207	BC 383	

Table 3. Various transistor types that meet the TUP specifications.

TUP		
BC 157	BC 253	BC 352
BC 158	BC 261	BC 415
BC 177	BC 262	BC 416
BC 178	BC 263	BC 417
BC 204	BC 307	BC 418
BC 205	BC 308	BC 419
BC 206	BC 309	BC 512
BC 212	BC 320	BC 513
BC 213	BC 321	BC 514
BC 214	BC 322	BC 557
BC 251	BC 350	BC 558
BC 252	BC 351	BC 559

The letters after the type number denote the current gain:

A: α' (β, h_{fe}) = 125-260
B: α' = 240-500
C: α' = 450-900.

Table 4. Various diodes that meet the DUS or DUG specifications.

DUS		DUG
BA 127	BA 318	OA 85
BA 217	BAX 13	OA 91
BA 218	BAY 61	OA 95
BA 221	1N914	AA 116
BA 222	1N4148	
BA 317		

Table 5. Minimum specifications for the BC107, -108, -109 and BC177, -178, -179 families (according to the Pro-Electron standard). Note that the BC179 does not necessarily meet the TUP specification ($I_{c,max}$ = 50 mA).

	NPN	PNP
	BC 107 BC 108 BC 109	BC 177 BC 178 BC 179
U_{ceo} max	45 V 20 V 20 V	45 V 25 V 20 V
U_{ebo} max	6 V 5 V 5 V	5 V 5 V 5 V
I_c max	100 mA 100 mA 100 mA	100 mA 100 mA 50 mA
$P_{tot.}$ max	300 mW 300 mW 300 mW	300 mW 300 mW 300 mW
f_T min.	150 MHz 150 MHz 150 MHz	130 MHz 130 MHz 130 MHz
F max	10 dB 10 dB 4 dB	10 dB 10 dB 4 dB

Table 6. Various equivalents for the BC107, -108, . . . families. The data are those given by the Pro-Electron standard; individual manufacturers will sometimes give better specifications for their own products.

NPN	PNP	Case	Remarks
BC 107 BC 108 BC 109	BC 177 BC 178 BC 179		
BC 147 BC 148 BC 149	BC 157 BC 158 BC 159		P_{max} = 250 mW
BC 207 BC 208 BC 209	BC 204 BC 205 BC 206		
BC 237 BC 238 BC 239	BC 307 BC 308 BC 309		
BC 317 BC 318 BC 319	BC 320 BC 321 BC 322		I_{cmax} = 150 mA
BC 347 BC 348 BC 349	BC 350 BC 351 BC 352		
BC 407 BC 408 BC 409	BC 417 BC 418 BC 419		P_{max} = 250 mW
BC 547 BC 548 BC 549	BC 557 BC 558 BC 559		P_{max} = 500 mW
BC 167 BC 168 BC 169	BC 257 BC 258 BC 259	!	169/259 I_{cmax} = 50 mA
BC 171 BC 172 BC 173	BC 251 BC 252 BC 253		251 . . . 253 low noise
BC 182 BC 183 BC 184	BC 212 BC 213 BC 214		I_{cmax} = 200 mA
BC 582 BC 583 BC 584	BC 512 BC 513 BC 514		I_{cmax} = 200 mA
BC 414 BC 414 BC 414	BC 416 BC 416 BC 416		low noise
BC 413 BC 413	BC 415 BC 415		low noise
BC 382 BC 383 BC 384			
BC 437 BC 438 BC 439			P_{max} = 220 mW
BC 467 BC 468 BC 469		!	P_{max} = 220 mW
	BC 261 BC 262 BC 263		low noise