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AN002

Using the OM1654 Simple Triac Control IC

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IES

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CONTENTS

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1 SUMMARY

The operating principles and external circuit component requirements of OM1654 are discussed. Specifically designed for triac control of medium and high power resistive heating loads, the OM1654 offers the simplicity of direct mains operation via two high ohmic resistors, affording inherent immunity to mains-born transients and significantly reduced power dissipation.

2 INTRODUCTION

The OM1654 is intended as a simple triac trigger IC requiring a minimum number of support components to realise an electronic thermostat for use in domestic heating or cooking appliances.

In order to achieve simple operation and yet retain important features such as zero-crossing triggering, immunity to mains conducted transients or to RFI, and minimum system dissipation, a number of constraints have to be recognised. These include the choice of a triac with a suitably specified gate sensitivity and the provision of a generated gate pulse width which has been designed to suit the chosen load.

An understanding of how these factors were chosen and implemented with the OM1654 will show how easily it can be used in other application areas.

3 STANDARD OM1654 APPLICATION CIRCUITS

In many applications a standard power supply and sensing circuit can be used. These are described in this part of this report. For more unusual applications, where the standard circuit does not offer the optimum performance, it is helpful to understand the general operating principles of OM1654. This is covered in more detail in the following section (Section 4) firstly by description and then with calculations and some examples.

3.1 Powering the chip

OM1654 uses just one powering resistor from ‘AC’ to Mains as shown in Fig.1. For operation over the range 210 V_{rms}. to 260 V_{rms}, a powering resistor of 220 kΩ can be used.

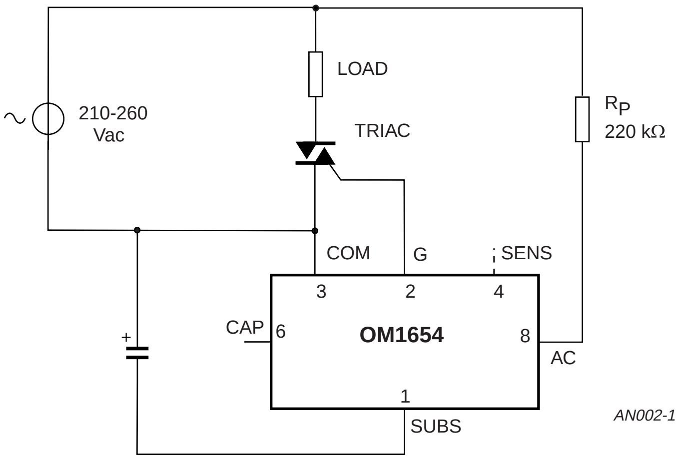

Fig.1 Standard powering circuit for OM1654

The triac should be specified to have a gate current that will trigger all devices of 10 mA maximum at 25 ^{o}C for negative gate current drive. Its current rating and heatsink should be chosen to suit the load.

The triac should also be chosen to have latching currents specified such that:

1. T2 positive, Gate negative, $I_{latching} <$ (7.5 V ÷ max. load resistance).
2. T2 negative, Gate negative, $I_{latching} <$ (5 V ÷ max. load resistance).

For operation over the range 100 V_{rms} to 130 V_{rms}, the preferred powering circuit is shown in Fig.2.

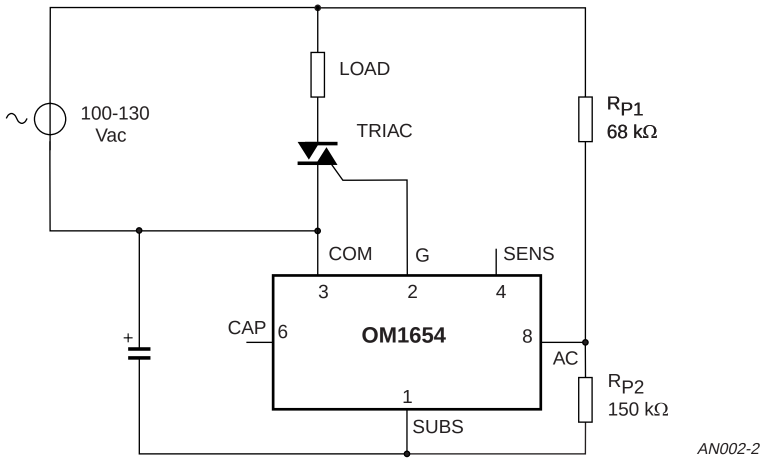

Fig.2 Powering circuit for 110 V operation

Here the powering resistor, R_{P1}, is 68 kΩ and a second resistor, R_{P2}, 150 kΩ is connected between pins `AC' and `SUBS' to modify the gate pulse timing. Again the triac gate sensitivity is 10 mA, but in this case both the specified latching currents must be less than 5 V ÷ max. load resistance.

Note:- The addition of resistor R_{P2} modifies the positive threshold voltage of the OM1654, thereby controlling the gate pulse timing. A further detailed explanation of this may be found in section 4.2, and in Appendices 1 and 4.

3.2 Supply Capacitor

As shown in Figure 3, a 47 µF 10 V capacitor makes an adequate supply bypass. In section 4.4.1 the choice of supply bypass capacitor is discussed in greater detail.

3.3 Timing capacitor

The usual timing capacitor value of around 0.47 µF gives an ‘on’ time of approximately 2 seconds. Capacitance values of the timing capacitor connected to ‘CAP’ that can be used may be typically in the range of 0.22 µF to 10 µF. The ‘on’ time is approximately 4 seconds per µF (microfarad) of timing capacitor.

3.4 Typical sensor circuits

Figure 3 shows a typical sensor configuration and Table 1 below gives suitable component values for some common applications.

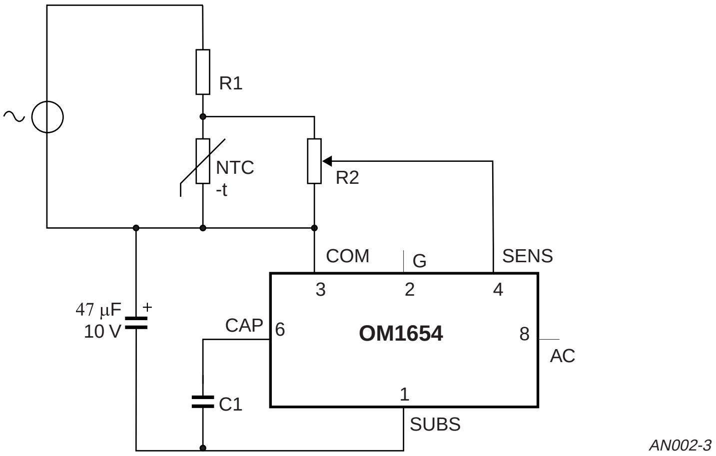

Fig.3 Typical sensor configuration for OM1654

Table 1 Typical component values for sensor circuits

APPLICATION	R1	R2	NTC @ 15 ^{O}C	C1
Frypan 230 V	220 kΩ	47 kΩ log.	100 kΩ see note 1	0.47 µF
Frypan 115 V	110 kΩ	47 kΩ log.	100 kΩ see note 1	0.47 µF
Deep fryer 230 V	270 kΩ	22 kΩ lin.	100 kΩ see note 1	0.47 µF
Deep fryer 115 V	137 kΩ	22 kΩ lin.	100 kΩ see note 1	0.47 µF
Heater 230 V	470 kΩ	47 kΩ log.	2.2 kΩ see note 2	4.7 µF

Note

1. e.g. NTC thermistor Philips 2322 633 83104
2. e.g. NTC thermistor Philips 2322 640 63222

4 OPERATING PRINCIPLES OF THE OM1654

For correct function of any triac trigger circuit, the following factors need to have been considered:

Gate current amplitude
Gate current pulse timing
Total supply current for IC, including the gate drive component
Sensor signal processing.

4.1 Gate current amplitude

The gate current drive is internally determined, and no adjustment is provided. The amplitude is set to provide a minimum gate current of 10 mA at 25 °C and the internal current limiting circuitry is temperature compensated, so that the gate current is increased as the temperature is decreased to match triac drive characteristics. This temperature compensation is normally required only when the circuit is first powered because the triac will usually increase in temperature once it has been conducting for a few cycles.

Note that the OM1654 IC has such low dissipation that self heating is negligible; unless there is thermal coupling between the IC and the triac, the IC drive will not track with the dynamic requirements of the triac. Such tracking is not normally required, the more important factor being that the triac drive requirements are met when first driven ‘cold’.

The reason to note this is that the IC power supply requirements increase at low temperatures to cover the increased gate drive requirements, and sufficient supply current must be provided for the lowest designed operating temperature, and also to maintain operation with ICs where the actual gate current has a design margin over the minimum specified limit of 10 mA.

The minimum 10 mA gate drive is designed to match Philips series ‘-E’ triacs available in the BT134 (4 A) through to the BT139 (16 A) range, although other triacs with a minimum gate drive requirement of 10 mA at 25 °C (for negative gate current) can obviously be used.

The gate drive output of OM1654 is designed to tolerate transient voltage spikes, as may be generated by the triac, of up to 50 V above or 30 V below the IC substrate (‘SUBS’, pin 1) voltage.

Although the gate current cannot be increased above its internally current-limited value, it remains a possibility to decrease the gate current by use of an external resistor, there will be limited practical advantage after tolerances and temperature effects are considered. (The nominal source, if not run in current limit, is 5 V ahead of 110 Ω (± 30 Ω)

4.2 Gate current pulse timing

For minimal generation of RFI, the triac requires a gate pulse that starts as the mains voltage approaches zero – before the triac ceases to conduct. In particular, it must start before the load current in the triac falls below the specified `holding' current of the chosen triac (that is the minimum current at which all triacs are guaranteed to remain `on'). The gate pulse must continue until the current in the triac has crossed zero and increased in amplitude again to the specified `latching' current (the minimum current at which all triacs have latched `on'). OM1654 senses the mains voltage via its supply resistor on pin `AC' and generates the required gate pulses.

The holding and latching current requirements are found in the triac data sheet and are normally specified only at 25 °C. They have temperature coefficients which are similar to the gate current temperature coefficients; that is negative temperature coefficients, so they all increase as the junction temperature falls. In the absence of published data it is wise to allow, for example, some 25 – 30% margin on the 25 °C figures if operation to 0 °C is likely.

For historic reasons it is usual to calculate the gate pulse in terms of pulse width and timing, perhaps because this timing is relatively easy to display on an oscilloscope. However an alternative design method offering simpler calculations is described here. For comparison with conventional (timing) design methods a discussion including timing is included in Appendix 1.

The currents in the triac are simply related to the instantaneous mains voltage and the load resistance by the relationship...

$$I_{TRIAC} = \frac{V_{MAINS} - V_{TRIAC}}{\text{load resistance}}$$

Where V_{TRIAC} = the voltage drop across the triac when I_{TRIAC} flows in the triac and the load, and, for the relatively small latching currents, it is safe to use 1 Volt for this drop when making calculations.

So the minimum allowed instantaneous mains voltage at the time the gate pulse is terminated by the IC is that value which allows the minimum latching current to flow. This voltage must be the threshold level of the mains `zero crossing' detector circuit. In fact there are two thresholds, one for T2 positive and one for T2 negative, and these are further specified for positive and negative gate currents. The OM1654 uses negative gate drive.

The required IC threshold voltages are calculated for each of the values of latching current, and are given by:

$$V_T = (I_{latching} \times R_{load}) + 1\text{ Volt}$$

Where the appropriate latching currents are found in the Triac data and may be identified as $I_{latching}$ 1– for the case where T2 is positive and $I_{latching}$ 3- for the case where T2 is negative.

As an example, the BT139–600E has latching currents of 30 mA for T2 negative and 40 mA for T2 positive. The negative threshold (VT–) of the OM1654 is –6 V, the positive threshold (VT+) is given as:

$$V_{T+} = 45\ \mu A \times R_P$$

Where R_P = powering resistor at ‘AC’.

For the 210 – 260 V application with 220 kΩ, that VT+ threshold is +9.9 V nominal and minimum +8.5 V. So the maximum load resistance for this (16 A) triac is limited by the T2 negative requirement for $I_{latching}$ = 30 mA with the -6 volt VT– threshold.

Max Load Resistance:

$$R_{load_{max}} = \frac{(6-1)\ V}{30\ mA} = 167\text{ Ohms}$$

This corresponds to a minimum load power of 317 W on 230 V.

OM1654 is published for loads exceeding 400 W, to allow for further tolerances on nominal load, Mains voltage, lower temperatures etc.

Of course, for low power loads it makes sense to choose a lower current triac than BT139 (16 A). The 4 A rated BT134 has a latching current of only 15 mA, allowing a minimum load power of 159 W at 25 °C.

4.3 Modifying zero crossing detector thresholds

It now becomes clear that the positive threshold (VT+) can, in fact, be modified by simply adding a further external resistor from pin ‘AC’ to ‘SUBS’ to modify the current in the feed resistor and the negative threshold can be increased by a divider to ‘COM’ that attenuates the voltage at ‘AC’. As will be seen, these two thresholds may be independently modified.

For T2 positive the OM1654 uses the ‘COM’ rail as the voltage threshold for terminating the gate pulse. So connecting an external resistor between ‘AC’ and ‘SUBS’ will cause additional current to be flowing in the Mains feed resistor when the gate pulse terminating threshold voltage is crossed. The additional current is given by the OM1654 supply voltage of typically 6.7 Volts divided by the added external resistor ‘AC – SUBS’.

4.3.1 EXAMPLE: T2 +ve:

Putting 220 kΩ there will cause a nominal current of 6.7 V divided by 220 kΩ or 30.45 µA. The current in the mains feed resistor at the end of the gate pulse then becomes 45 + 30.45 = 75.45 µA. For the same mains feed resistor of 220 kΩ, the mains voltage at the end of the gate pulse is 75.45 µA x 220 kΩ = 16.6 V.

Allowing 1 V drop on the triac, the load voltage is 15.6 V.

To achieve the minimum latching current of 40 mA requires a maximum load resistance of 390 Ω, which corresponds to a minimum load power of 136 W on 230 V at 25 °C for the BT139-600E.

Note: When calculating the total power supply current requirements this 220 kΩ ‘AC-SUBS’ resistor will cause an additional current of 30.45 µA only during positive half mains cycles, so an average current of 15.23 µA will be added to the supply current requirement.

4.3.2 EXAMPLE: T2 –VE:

A similar calculation can be made for negative half cycles.

By using a resistor from ‘COM’ to ‘AC’ it is possible to also alter the effective negative voltage threshold of the zero-crossing detector. This resistor forms, with the mains feed resistor, a simple voltage divider. So fitting 220 kΩ between the ‘AC’ and ‘COM’ means the mains must reach –12 V for the voltage at ‘AC’ to reach the OM1654 detection threshold of –6 V. In this way the minimum load power to meet the $I_{latching}$ 3– requirement is decreased to just 72 W on 230 V for the BT134-600E at 25 °C.

(Note that this resistor also diverts a current of 30.45 µA from the power supply current during negative mains half cycles, so it is necessary to add 15.23 µA DC to the average power supply requirement.)

As will become clear when the additional power supply current requirements are considered, it is not practical to apply OM1654 with large gate pulse extensions to handle very low power loads. OM1682 is more flexible in this regard and will be a better choice for loads less than about 50 W on 230 V.

It is also recommended that the minimum detector threshold for positive half cycles be 6 V. There is no theoretical need to set this limitation, but in practice lesser thresholds seem to unduly risk interference from quite small noise voltages. The magnitude of the negative threshold is determined by the IC supply rail so it can not be user-set to less than 6 V in any case.

4.4 IC supply current

The IC supply current has 4 major components, the supply voltage regulator, the current used in the zero-cross detection circuit, the charge current for the timing capacitor on ‘CAP’, and the average component of the gate current pulses. (In special applications there may be additional currents caused by resistors, added to modify the normal operation, as discussed above.)

Both the average gate current and the zero-cross detection current (which averages around 25 µA) are required only during the triac ‘ON’ period, while the ‘CAP’ charging current for normal cyclic operation is gated on only during triac ‘OFF’ periods.

Normally the charging current of ‘CAP’ cannot reach the sum of gate and zero detect currents, so this current is not usually considered. However it may become important under unusual conditions, e.g. if a large capacitor is fitted at ‘CAP’ and charged by circuits external to OM1654 which do not limit this current or inhibit it during ‘ON’ periods.

The OM1654 requires about 75 µA for its basic internal regulator and logic circuits at 25 °C. For calculation proposes the sum of the regulator and zero-cross circuit currents (i.e. the normal operating supply current of OM1654) may be taken as maximum 150 µA at 100 °C, and 120 µA at 0 °C.

The OM1654 provides nominal gate current of 12.5 mA to allow for IC processing spreads. An additional 0.5 mA is used in the internal drive circuit. For calculations, a maximum of 15.5 mA total should be used. This has a temperature coefficient of approximately –0.5% per °C. So for operation down to 0 °C an additional 25 x 0.5% = 12.5% should be allowed, making the design value 17.4 mA maximum.

The average supply current required for gate drive can be calculated by multiplying this current by the duty cycle for which it is provided.

As derived in Appendix 3 the average current is given by:

$$I_{G_{ave}} = 17.4 \times \left(\frac{V_{T+} + V_{T-}}{\sqrt{2} \times V_{rms} \times \pi} \right)$$

Where V_{T+} and V_{T-} are the positive and negative thresholds respectively of the mains used to set the gate pulse.

In 210 V – 260 V operation, with a 220 kΩ feed resistor, the gate pulses last from +9.9 V to –6 V resulting in a maximum average gate current components of 296 µA at 210 V at 0 °C.

Add to this the worst case static supply current of 120 µA at 0 °C, and the total requirement becomes 416 µA.

For half wave rectification, as used in OM1654, the DC component of the Mains feed current is derived in Appendix 2 and is:

$$I_{S_{ave}} = \frac{210 \times \sqrt{2}}{\pi \times 220\ \text{k}\Omega} - \frac{6.7}{2 \times 220\ \text{k}\Omega} = 415\ \mu\text{A}$$

While OM1654 was designed for nominal 220 – 240 V operation, a calculation for 110 V operation is presented in Appendix 4 to illustrate the points so far discussed.

4.4.1 CHOICE OF SUPPLY BYPASS CAPACITOR

The IC is quite tolerant to supply voltage ripple, 500 mV or more is allowable, and the supply bypass should not be too generous because the supply current is quite small and will take a considerable time to charge large supply filter capacitors at first application of the power.

A reasonable range would be 47 µF to 100 µF. The 47 µF restricts the ripple to around 150 mV. For high temperature operation, e.g. at 100 °C, the use of solid aluminium capacitors will ensure high reliability and very long life.

4.5 Sensor signal processing

OM1654 is intended for applications with NTC sensors which have typical temperature coefficients of around 3% per °C, and for relatively non-critical temperature control of domestic cooking and heating appliances. Such applications will typically feature continuously variable manual control knobs with scales that are not intended to be accurate to better than 5 degrees. Indeed, it will challenge bi-metal controllers which will typically have 10 degrees hysteresis in addition to their nominal manufacturing tolerances.

So the IC only needs sufficient accuracy to process a signal from the NTC which is changing by 5 ^oC x 3% per ^oC = 15% in magnitude in order to achieve this target control accuracy.

In the OM1654 a simple diode threshold detector is used. In combination with a potentiometer for threshold level adjustment, this produces a voltage threshold detector of N times one junction diode level, where N is the divider ratio set by the potentiometer. It is simply a matter of adjusting the attenuation range to cover the range of voltage that appears on the NTC over the design operating temperature range.

The internal threshold is approximately 0.65 V, and has a normal junction temperature coefficient of 2 mV per ^oC. To determine the effect on performance due to this simple detector, consider a change in IC ambient of 50 ^oC. This will change the detection threshold by 100 mV or 15%. That will require an equivalent change in the sensor NTC resistance of 15%, which corresponds to an error in controlled temperature of about 15 ÷ 3 = 5 ^oC.

In most applications the IC will have a predictable temperature by the time the load has reached its controlled temperature, and the spread in possible IC ambient temperatures would be much less than 50 ^oC. This expected IC ambient temperature can be used in calculations so that the overall target accuracy of 5 degrees will be easily achieved.

4.5.1 SENSOR CIRCUIT OPERATION

An equivalent circuit for the internal threshold detector of the OM1654 `SENS' input is shown in Figure 4.

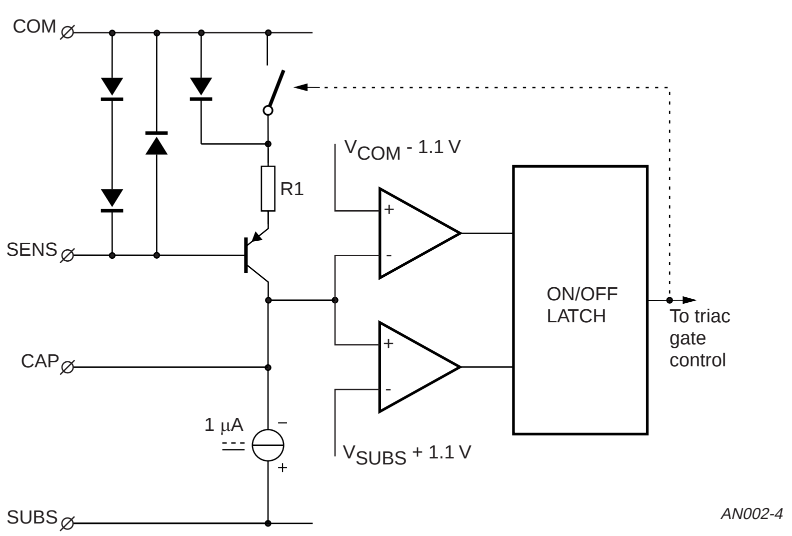

Fig.4 Equivalent sense input circuit of OM1654

In the following description it is assumed that the timing capacitor is connected between `CAP' and `SUBS' for reasons discussed below. There is a constant discharge current of 1 µA at all times pulling the `CAP' pin towards `SUBS'. When the voltage on `CAP' reaches a detection threshold, 1.1 V above `SUBS', switch SW1 is closed and the input detector circuit at `SENS' is enabled so that `CAP' may be charged. The triac drive circuitry is disabled, creating the `OFF' period for the load. There will be no charge current into the capacitor at `CAP' unless the threshold level at `SENS' of one junction voltage (about 0.65 V) is exceeded, by a margin sufficient to cause a collector current in T1 which exceeds the continuous 1 µA discharge current. For higher voltages at ‘SENS’ the capacitor charge current via T1 increases and is approximately given (for DC voltages) by:

$$I_C = \frac{(V_{SENS} - 0.65)}{4\ \text{k}\Omega}$$

The charge current is limited to approximately 150 µA by limiting the voltage at ‘SENS’ to two junction voltages (about 1.3 V).

The charge current from T1 charges ‘CAP’ towards V_{COM}. When ‘CAP’ is charged up to a second threshold detector level, which is approximately 1.1 volts below ‘COM’ (i.e. V_{COM} – 1.1 volts), the triac drive is enabled and the load enters an ‘ON’ period. The zero-crossing circuitry is enabled and the triac will be fired at each zero crossing. The input circuit at ‘SENS’ is desensitised by opening switch SW1 during the triac ‘ON’ time so that there is no charge current into ‘CAP’ and the capacitor is discharged by the internal 1 µA current towards the ‘OFF’ detection threshold again.

4.5.2 100% DUTY CYCLE

For input signals at ‘SENS’ which are about double the normal threshold level, and which cause the input clipping diodes to conduct, a small charging current is fed to ‘CAP’ preventing discharge and therefore achieving 100% ‘ON’ time of the load.

Because this charging path is operational during the ‘ON’ period of the load (normal charging of the capacitor occurs during the ‘OFF’ period) it is necessary to power the sensor circuit directly from the mains, and not from the voltage across the triac, if this feature is required.

Alternatively, if it is not required, the sensor circuit may be powered by connecting the feed resistor to the triac. In this way the sensor circuit is powered via the load resistor and power is available only during the ‘OFF’ periods.

This method of powering the sensor circuit is advantageous whenever it is essential to achieve the lowest possible power dissipation in the control circuit, but the highest practical duty cycle of the load will be limited to about 95% ‘ON’ time for the AC powered sensor circuits usually employed.

4.5.3 SETTING ‘ON’ TIMING PERIOD

The ‘ON’ period of the load, set by the discharge time of the capacitor at ‘CAP’, depends only on the difference in the voltage between the two detection thresholds, the 1 µA discharge current, and the capacitor fitted at ‘CAP’.

The ‘ON’ period timing is calculated from the expression:

$$C \times V = I \times t$$

Where C = timing capacitor,
V = voltage between thresholds which is approximately: V_{SUBS} – 2.2 V = 4.5 V and I = 1 µA

So the timing is approximately 4.5 seconds per microfarad at ‘CAP’. It is clear that a low leakage capacitor is required if the ‘ON’ timing is to be independent of capacitor leakage. It is also necessary to connect ‘leaky’ capacitors, such as electrolytic capacitors, between ‘CAP’ and ‘SUBS’ so that any leakage simply adds to the internal discharge current and simply shortens the ‘ON’ time. A leaky capacitor between ‘COM’ and ‘CAP’, with more than 1 µA leakage could never be charged to the ‘OFF’ threshold and would simply leave the load ‘ON’.

With the timing capacitor connected to ‘SUBS’, a small leakage, even a microamp or so, will not cause any significant error in the controlled temperature because only a very tiny change to the voltage at ‘SENS’ is required to increase the charge current by this amount.

For example the charge current changes from 2 µA to 4 µA for less that 2 or 3 ^{o}C change in controlled load temperature.

It is difficult to calculate the effective charge current obtained by half-wave, peak rectification of the ac signal at ‘SENS’, so a table of approximate rms voltages and the corresponding load ON/OFF duty cycles is shown in Table 2. Some worked examples of sensor circuit designs are given in Appendix 4.

Table 2

SENS VOLTS (RMS)	DUTY CYCLE
0.5	5%
0.52	25%
0.54	50%
0.58	75%
0.80	95%
0.92	100%

5 SYSTEM DESIGN CONSIDERATIONS

5.1 Thermal design

In every control system the thermal design is as important as the circuit design.

When controlling high power loads it is important to recognise the rise in load temperature during the ‘ON’ period. For example a domestic Iron having a typical heater of 1200 W and a 400 gram sole-plate will heat by 3 ºC in just one second of ‘ON’ time. If an ‘ON’ time of 2 seconds was selected to guarantee a minimum load switching time, as required in some countries, then a minimum cycling of 6 ºC would result. A temperature ‘error’ of 6 ºC will inhibit ‘CAP’ charging with the result that ‘proportional’ control is effectively lost and the system will behave in a manner similar to an ON/OFF controller.

While this performance will already offer an improvement over bi-metal controls, further improvement would require modification of the thermal components of the system. Some suggestions regarding the attachment of the NTC sensor are given in Section “5.4 Mounting the sensor NTC”

5.2 Wiring layout

In any triac controller design it is important that the gate drive connections be made as close to the triac as possible. If the control circuit must be mounted some distance away from the triac then it is important to make the connection to the `COM' of the circuit by a separate wire which connects to the triac T1 lead or close to it. This is to prevent `degeneration' of the gate drive voltage caused by current flowing in the T1 wiring. Fig.5 and Fig.6 illustrate both the correct, and the incorrect way of wiring the triac and connecting the sense input.

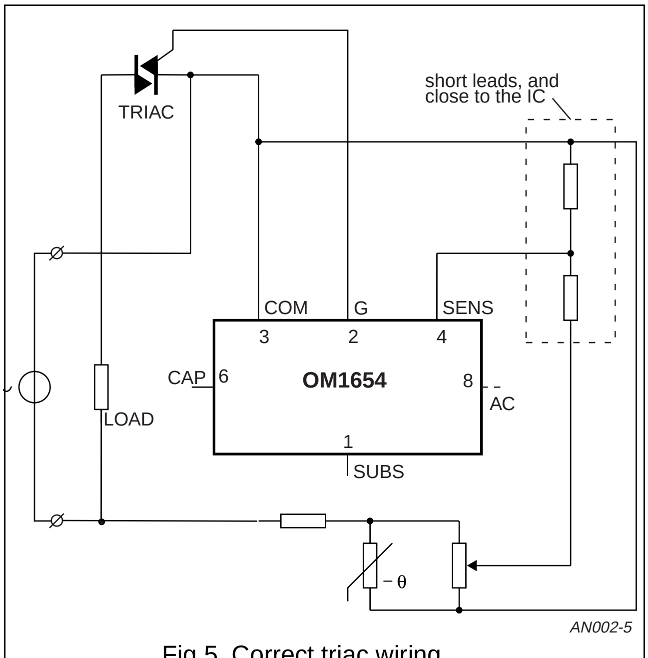

Fig.5 Correct triac wiring

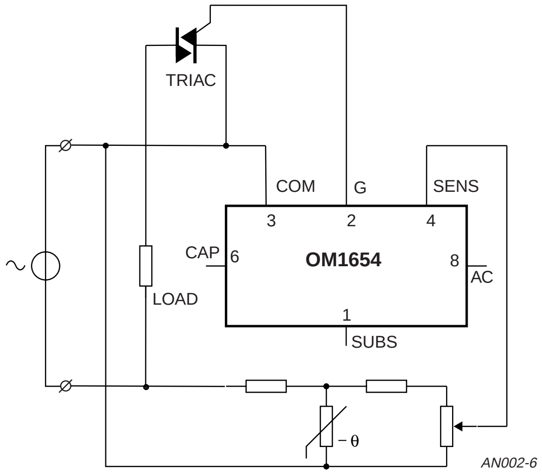

Fig.6 Incorrect triac wiring

Likewise, the signal at `SENS' is small, so good practice will be to fit any attenuator of the NTC voltage close to the IC. Where long sensor wiring is necessary it will be helpful if the NTC can be operated at a relatively high voltage, e.g. 10 V_{rms}, and the signal be attenuated from this level by components close to the IC. If extreme environments require RF suppression then capacitors can be fitted between `COM' and `SENS' and/or `COM' and the sensor lead.

5.3 Triac heatsinking

Reference should be made to the triac data to determine the exact dissipation, but it is often worthwhile to make an early estimate of the heatsinking requirements. A good guide is to allow 1 Watt of triac dissipation per 1 Amp of load current.

Where the typical load duty cycle can be accurately estimated this duty cycle can also be applied to the total triac dissipation to determine its average dissipation. A first estimate of the required heatsink area can be based on an approximation to its thermal resistance. A dissipation of 1.5 milliwatts over a surface area of 1 square centimetre will cause about one degree C rise in that surface above its surrounding ambient air temperature.

So the temperature rise of a heatsink, above its surrounding ambient, is approximately:

$$T_{rise} = \frac{\text{Total watts dissipated}}{\text{heatsink area} \times 0.0015} \quad (\text{in } {}^{\circ}\text{C})$$

Where the heatsink area is in square cm. and is the total of all exposed surfaces.

5.4 Mounting the sensor NTC

In high temperature applications it is not possible to use solder connections to the NTC unless special high melting point solders are used. It may be found more convenient to use mechanical crimping of the NTC leads to the connection wires leading away to the controller. The most convenient way to heat a ‘diode style’ NTC is by thermal conduction along the connection wires. The heat loss to ambient via the tiny body area will usually be negligible if the leads can be held at the temperature of the surface to be sensed. That is, if the leads can be thermally coupled to the object to be controlled then accurate sensing will be achieved. It is necessary to achieve full mains electrical insulation.

A suitable approach is illustrated in Fig.7 which shows a relatively large area of a conducting metal thermally coupled by reason of its significant area, yet electrically insulated. The NTC lead is then crimped, together with the lead that goes to the controller, to this electrically live piece of metal. The NTC is then heated by conduction along the lead. Attempts to couple heat via the glass of the ‘diode’ are likely to be less effective because the metal connection leads would then provide some thermal coupling or ‘heat sinking’ to ambient.

An effective thermal coupling can often be made by simply clamping a significant length of a high temperature, mains insulated, connecting lead (e.g. Teflon covered or sleeved) against the surface to be sensed, without any attempt to couple directly to the body of the NTC. An example of this mounting approach is illustrated in Fig.8.

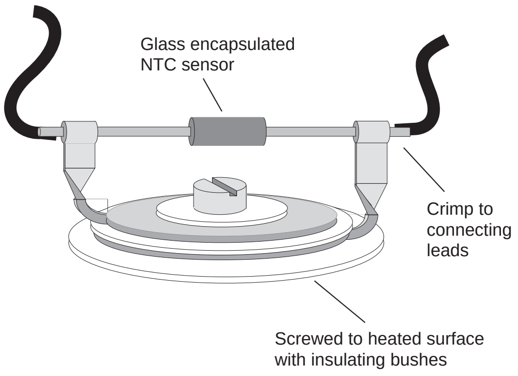

Fig.7 Thermal coupling via insulated crimp terminations

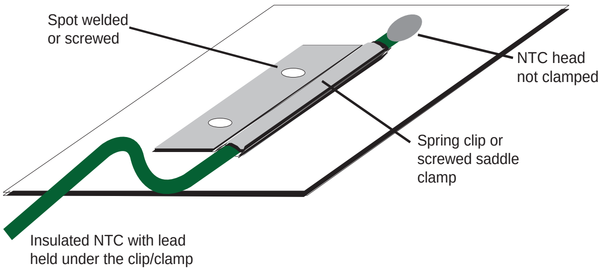

Fig.8 NTC with the lead electrically insulated, secured under clip/clamp

6 APPENDIX 1: CALCULATION OF GATE PULSE WIDTH

The times at which the holding or latching current limits will be reached, or the times at which the gate pulse timing detector threshold voltages will be crossed, are calculated as follows.

The slope or rate of increase of the sinusoidal mains voltage around its zero-crossing is found by differentiation.

If the peak amplitude of the Mains voltage is V volts, then the instantaneous voltage v is a sine-wave described by:

$$v = V \cdot \sin\omega t$$

Where ω is the angular frequency.

So by differentiation, the slope of the voltage is given as:

$$\frac{dv}{dt} = V \cdot \omega \cdot \cos\omega t$$

and the slope near the zero crossing, when t is near zero and $\cos 0 = 1$ is:

$$\frac{dv}{dt} = V \cdot \omega$$

In the more usual terms of frequency, $\omega = 2 \cdot \pi \cdot f$,(where f = frequency in Hz), so:

$$\frac{dv}{dt} = 2 \cdot \pi \cdot f \cdot V$$

$$= 2 \cdot \pi \cdot f \cdot v \cdot \sqrt{2} \qquad \text{volts/sec}$$

$$= 8.886 \cdot f \cdot v \qquad \text{volts/sec}$$

$$= 102.2 \text{ kV/sec for 230 V, 50 Hz mains.}$$

$$(= 61.3 \text{ kV/sec for115 V, 60 Hz mains.})$$

For OM1654, which has two voltage thresholds that determine the start and finish of each gate pulse, the total duration of the gate pulse may be calculated as:

$$t_g = \frac{V_{t+} + |V_{t-}|}{dv/dt}$$

Where V_{T+} and V_{T-} are respectively the positive and negative thresholds of the OM1654.

For example, with a 220 kΩ powering resistor on 230 V 50 Hz mains the thresholds are –6 V and +9.9 V, so the gate pulse width $t_g = 156\,\mu s$.

With a load resistance of R ohms, the rate of increase of load current near the mains voltage zero will be:

$$\frac{di}{dt} = \frac{102.2}{R} \qquad \text{kA/sec. (or more conveniently mA/}\mu\text{s)}$$

Remember the triac has a diode-like characteristic, so the load current (I_L) is given by:

$$I_L = \frac{\text{mains voltage} - \text{triac ON voltage}}{\text{load resistance}}$$

So for the first 0.7 to 1 volt increase of mains voltage there is no load current. That is, for 1 V ÷ 102.2 V/millisecond, (= 9.8 µs).

The additional time, after this delay, to reach the required latching current can be calculated by dividing the latching current by the di/dt.

7 APPENDIX 2: CALCULATION OF AVERAGE GATE DRIVE CURRENT

For a sine-wave of peak amplitude V volts, the time to reach a level of v volts is calculates as follows:

$$v = V \cdot \sin\omega t$$

$$\omega t = \operatorname{asin}\frac{v}{V}$$

$$t = \frac{1}{\omega} \cdot \operatorname{asin}\frac{v}{V}$$

For small angles (i.e. gate pulses which are a small fraction of the total half cycle) then:

$$\operatorname{asin}\frac{v}{V} \approx \frac{v}{V}$$

$$\text{so} \qquad t \approx \frac{1}{\omega} \cdot \frac{v}{V}$$

The period of a complete half cycle = π ÷ ω.

So the duty cycle for a pulse lasting until level v is reached is t ÷ time for one half cycle;

$$\text{Duty cycle} = \frac{t \cdot \omega}{\pi}$$

$$\text{and here} \qquad t = \frac{1}{\omega} \cdot \frac{v}{V}$$

So the duty cycle becomes:

$$\text{Duty cycle} = \frac{v}{\pi \cdot V} \qquad \text{and} \qquad V = \sqrt{2} \cdot \text{rms voltage}$$

$$= \frac{v}{\sqrt{2} \cdot \pi \cdot \text{rms mains voltage}}$$

For the OM1654 there are two portions of the Mains voltage, each side of the zero crossing, for which the gate pule is enabled. Each one has a duty cycle as given, and the total duty cycle is the sum of the two.

$$\text{total gate pulse duty cycle} = \frac{\text{sum of threshold voltage magnitudes}}{\sqrt{2} \cdot \pi \cdot \text{rms mains voltage}}$$

Then the contribution to the average supply current is simply the actual gate current (plus internal drive currents) multiplied by the gate current duty cycle. The internal drive current used by the IC during the gate pulse is approximately 0.5 mA. The nominal gate drive at 25 ^{o}C is 12.5 mA. The maximum gate pulse current at 25 ^{o}C to use for calculation purposes is 15.5 mA, including the internal drive current component. This has a temperature coefficient of approximately –0.5% per ^{o}C.

8 APPENDIX 3: CALCULATING AVAILABLE SUPPLY CURRENT

The supply current available from the supply resistor is found by integrating the current in the supply resistor over one complete cycle. The effective voltage across the resistor is closely equal to the mains voltage minus the chip supply. Half wave rectification is used, so the effective voltage during one half of the cycle is zero.

The voltage across the resistor = ($V \cdot \sin\omega t - 6.7$) volts where V = peak mains voltage and 6.7 V is the typical supply voltage.

The average voltage, by integrating over a half cycle (T/2), is:

$$V_{av} = \frac{1}{T}\int_0^{\frac{T}{2}}(V \cdot \sin\omega t - 6.7) \qquad \text{and} \qquad \omega = \frac{2 \cdot \omega}{T}$$

$$= \frac{1}{T}\left[\frac{V \cdot T}{2 \cdot \omega}\cos\frac{2 \cdot \pi}{T}t - 6.7 \cdot t\right]_0^{\frac{T}{2}}$$

$$= \frac{V}{\pi} - \frac{6.7}{2} \quad \text{volts}$$

Therefore the average supply current becomes:

$$I_{av} = \frac{V_{av}}{R_s}$$

$$= \frac{\left[\frac{V}{\pi} - \frac{6.7}{2}\right]}{R_s} \qquad \text{mA}$$

where R_S = the supply resistor in kilohms.

9 APPENDIX 4: DESIGN EXAMPLES

9.1 Example 1: Design for a 100V, 60 Hz, 50 W load, using BT134-500E

Assume maximum load resistance is nominal (242 Ω) + 10% = 266 Ω.

Assume minimum mains voltage = 100 V_{rms}.

Triac latching current ($I_{latching}$))
= 15 mA negative T2, 20 mA positive T2.

Required positive mains threshold
= 1 V for triac + 20 mA × 266 Ω = 6.3 V.

Required negative mains threshold
= 1 V + 15 mA × 266 Ω = 5 V,

so no modification of the negative threshold will be required (it is already – 6 V).

Before the positive threshold can be calculated it is necessary to know the supply dropping resistor value. This is calculated from the supply current requirement. Assume operation to 0 ^{o}C, so use 17.4 mA for the gate current.

The average gate current:

$$I_{G_{ave}} = 17.4\left[\frac{12.3}{\sqrt{2}\cdot\pi\cdot 100}\right] = 482\ \ \mu A$$

Add 120 μA for IC supply, making a total of 602 μA.

The supply resistor (R_S) to supply this average current is given by:

$$R_s = \frac{\left[\frac{\sqrt{2}\cdot 100}{\pi} - \frac{6.7}{2}\right]}{602\mu A} = 69.2\ \ k\Omega$$

Let's choose 62 kΩ, which will provide 672 μA because we know some additional current will be required by the resistor that we will need to increase the positive threshold.

Now the positive threshold will be:

$$V_{T+} = 62\ k\Omega\cdot 45\ \mu A = 2.8\ \ V$$

So a resistor connected between AC and SUBS is required to increase this to 6.3 V. An additional 56 μA is required, the supply is –6.7 V so the required resistor is 119 kΩ, so choose the nearest convenient value of 110 kΩ. The average supply current contribution of the added resistor is 30.5 μA which is already available because the 62 kΩ supply resistor gave a 70 microamp margin.

9.2 Example 2: Electric frypan

This type of electric cooker has an operating temperature range, manually adjusted, from room temperature to 200 ºC. The lowest important temperature setting is around 60 ºC. The control knob should provide an ‘OFF’ condition when fully anti-clockwise. The biggest problem for NTC sensing of high temperature cooking equipment is the very large resistance variation of the NTC and its much lower resistance at the highest temperatures.

One NTC intended for high temperature applications is the glass encapsulated ‘diode style’ type thermistor, Philips 2322 633 82104. The nominal resistance/temperature characteristic is given in Table 3 and shows that its resistance falls from 100 kΩ at 25 ºC to just 640 Ω at 200 ºC. The range over which control is required is from about 12.56 kΩ to 640 Ω which is about a 20:1 range.

Table 3 Philips Thermistor 2322 633 82104

TEMPERATURE DEGREES CELSIUS	THERMISTOR RESISTANCE (OHMS)
–40	3,306 k
0	325.5 k
25	100 k
50	36 k
100	6,770
125	3,400
150	1,840
160	1,460
170	1,170
180	950
190	780
200	640
210	520
220	440
230	370
240	310
250	270
260	230
270	200
280	170
290	150
300	130

The OM1654 takes advantage of the relative accuracy and stability of the mains voltage to achieve an effective ‘constant current’ feed of the sensor circuit by choosing the series feed resistor to be more than 10 times the NTC resistance at the lowest important temperature. This results in an AC voltage across the NTC which also has about a 20:1 ratio over the control temperature range.

We must now choose a potentiometer circuit with a similar attenuation range so that the input to the OM1654 threshold detector will be more or less constant at the threshold level of about 0.5 V_{rms}.

The most accurate solution will be to apply a linear potentiometer with shunt and series resistors, but for such a non-critical application the use of a simple log potentiometer looks attractive. At 50% rotation a ‘log’ potentiometer has a nominal attenuation ratio of 9:1. The total potentiometer resistance should be larger than the NTC resistance at the lowest important control temperature. If, for example, it was chosen equal to the NTC resistance then the effective ‘sensitivity’ of control is halved because a 1 degree C temperature change will change the NTC by 3%, but the combination with a fixed, equal, parallel resistor results in a change of only 1.5% for 1 deg C.

Notice here that the use of a very large feed resistor from the mains is an important factor in maintaining this ‘sensitivity’ of control because that feed resistor has the same effect as the parallel loading of the potentiometer. The potentiometer should not be chosen with too high a resistance because the input impedance at ‘SENS’ is around 150 kΩ and needs to be included in the calculations if the source impedance of the ‘SENS’ signal is large. So select a log potentiometer of 47 kΩ and connect it directly across the NTC. The wiper arm is connected to ‘SENS’.

With the potentiometer around 3/4 rotation and the NTC therefore well below 3 kΩ, the maximum source resistance of the ‘SENS’ signal will be always less than 12.5 kΩ. (47 kΩ + 3 kΩ total resistance, max source impedance with wiper at 50% resistance point, giving two equal resistors of 25 kΩ.)

With the potentiometer wound fully clockwise, the hottest setting, we require a duty cycle of, say, 75% corresponding to a ‘SENS’ voltage of about 0.54 V_{rms}, allowing for, say, 20 ºC heating of the IC by the appliance when hottest.

We must be able to achieve this operation even for lowest mains and lowest NTC; so allow 10% tolerance for each. The effect of pot tolerance is negligible because the NTC resistance is below 1 kΩ. So we require 0.54 V across 640 Ω – 10% in parallel with the 47 kΩ Pot and the effective input impedance at ‘SENS’, which will be greater than 150 kΩ. That resistance combination gives 567 Ω.

For a mains voltage of 230 V nominal, 210 V minimum, a series resistor of 220 kΩ is required. At 1/2 of full rotation, and assuming here the duty cycle falls because a lower temperature is selected, there must be say, 0.51 V at ‘SENS’ and the NTC voltage is then:

$$V_{NTC} = 9 \times 0.51 \text{ V} = 4.59 \text{ V}$$

That corresponds, for the nominal 230 V mains, to a resistance of 4.48 kΩ. The potentiometer is 47 kΩ, so the NTC must be 4.95 kΩ, which corresponds to an operating temperature of just over 110 ºC.

At 3/4 rotation the potentiometer ratio is about 0.37 : 1, allowing for the 150 kΩ loading by ‘SENS’ of the nominal 0.4 : 1 pot tap. The appliance will need a slightly increased duty cycle, let’s say ‘SENS’ = 0.52 V, so the NTC resistance must become 1.39 kΩ corresponding to 160 ºC.

At 1/4 rotation the nominal temperature is about 60 ºC, and at half this rotation there will not be enough ‘SENS’ voltage for any operation at all.

So for rotation less than about 30 degrees we are guaranteed an ‘OFF’ condition. Clearly at zero rotation there must be a total ‘OFF’, so for many applications it will not be necessary to design any mechanical switch.

Because we allowed for NTC and mains tolerances, the 200 ºC will typically be obtained before full clockwise rotation and the user scale would be marked for the typical case as illustrated in Fig.9.

The scale can be seen to be very evenly divided from 60 ºC to 200 ºC even for the most simple attenuation method. It is something less than 1 ºC per one degree of rotation and such appliance knobs will rarely be marked with, or expected to be accurate to, better than 20 ºC intervals.

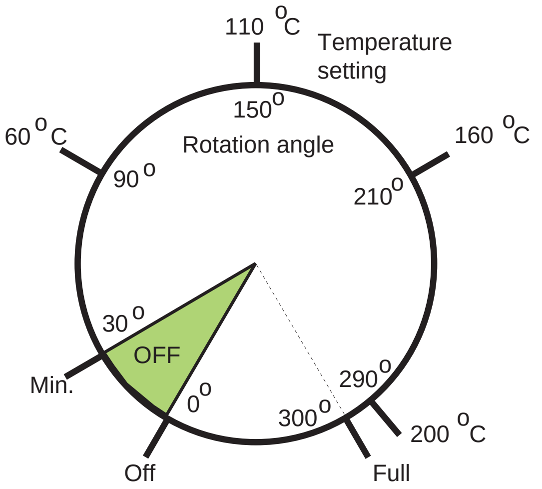

Fig.9 Control scale for Fry-pan, using a 47 kΩ Log. potentiometer.

9.3 Example 3: Deep fat fryer

This appliance has a similar maximum temperature of operation to the Frypan. The major difference is that the most useful section of its operating range is that part between about 130 °C and 190 °C. Again it is possible to calculate accurate networks, but simply exchanging a linear potentiometer for the log pot will serve to expand the sensitivity at the higher temperatures at the expense of those below 130 °C.

Because only higher temperatures are important, loading of the NTC by the potentiometer is less critical than ‘SENS’ loading it, so choose a value of 22 kΩ for the potentiometer. Note that the temperature coefficient of the NTC at 190 °C has fallen to around 2% per degree C, so it becomes important to specify the NTC in a way which minimises errors.

A standard NTC is specified at 25 °C to have a certain tolerance, say 5% resistance tolerance. It also has a tolerance on the characteristic shape of resistance versus temperature, called ‘B value’ tolerance.

Data shows that the B-value tolerance at 190 °C causes an additional 6.5% tolerance in resistance at 190 °C. Therefore the total tolerance at 190 °C is about 11.5%, corresponding to 5.7 degrees C.

For small quantity production the best solution is to use a trim-pot to individually adjust controls for the necessary accuracy at 190 °C. Large manufacturers should approach the NTC supplier to have a custom specification in which the maker guarantees a tolerance at 190 °C to replace the 25 °C specification. A typical specification would be R190 = 776 ± 6% which corresponds to just ± 2 degrees C at 190 °C.

If we again design for a mains of 230 V, with spread assumed 210 – 240 V, then we choose the NTC feed resistor to achieve 0.54 V_{rms} when the NTC is –6% and the mains is 210 V. In this way we guarantee 190 °C can still be reached even with worst case tolerances.

We can then determine the typical appliance scale for nominal conditions. The NTC at 190 °C can be 776 Ω – 6% = 729 Ω. The potentiometer can be 22 kΩ – 10% across the NTC giving a resultant 703.5 Ω. The feed resistor for 0.54 V across 703 Ω at 210 V mains is 273 kΩ so choose 270 kΩ.

Now for nominal 230 V and nominal components the voltage at the NTC will be 0.637 V_{rms} at 190 °C, so the potentiometer must be adjusted to 85% to give 0.54 V at `SENS'. That is, it will be at about 250 degrees mechanical rotation. (300 degrees is the maximum rotation angle).

At the centre, 50% attenuation, the NTC voltage must be 1.08 V_{rms}. and the corresponding NTC resistance is 1.35 kΩ, equivalent to 165 °C.

At 130 °C the rotation can be calculated as 75 degrees.

`OFF' will occur for rotation less than approximately 30 degrees from the anticlockwise stop.

The wanted control range of 60 °C spans almost 180 mechanical degrees allowing a convenient user scale as shown in Fig.10.

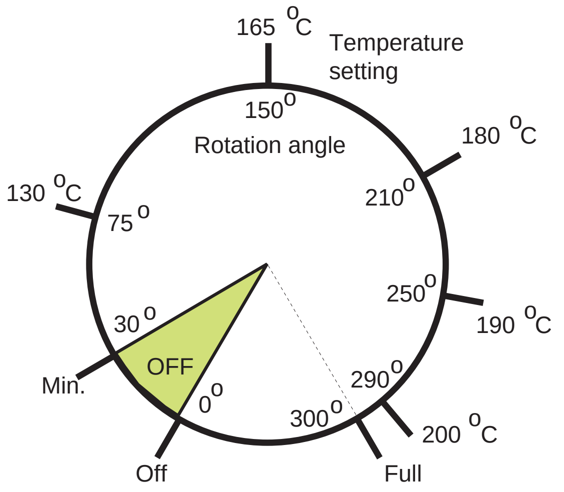

Fig.10 Control scale for “Deep Fat Fryer”, using a 22 kΩ linear potentiometer.

10 REFERENCES

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11 DEFINITIONS

Data sheet status	
Engineering sample information	This contains draft information describing an engineering sample provided to demonstrate possible function and feasibility. Engineering samples have no guarantee that they will perform as described in all details.
Objective specification	This data sheet contains target or goal specifications for product development. Engineering samples have no guarantee that they will function as described in all details.
Preliminary specification	This data sheet contains preliminary data; supplementary data may be published later. Products to this data may not yet have been fully tested, and their performance fully documented.
Product specification	This data sheet contains final product specifications.
Limiting values	
Limiting values given are in accordance with the Absolute Maximum Rating System (IEC 134). Stress above one or more of the limiting values may cause permanent damage to the device. These are stress ratings only and operation of the device at these or at any other conditions above those given in the Characteristics sections of the specification is not implied. Exposure to limiting values for extended periods may affect device reliability.	
Application information	
Where application information is given, it is advisory and does not form part of the specification.	

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