A low distortion audio oscillator

15Hz-150kHz output, .003% distortion

This new audio oscillator provides a low distortion sinewave output with a THD of typically .003% and covers the frequency range 15Hz to 150kHz in four overlapping ranges. Square wave output is also provided using a VMOS output stage with rise and fall times less than 20ns.

by RON DE JONG

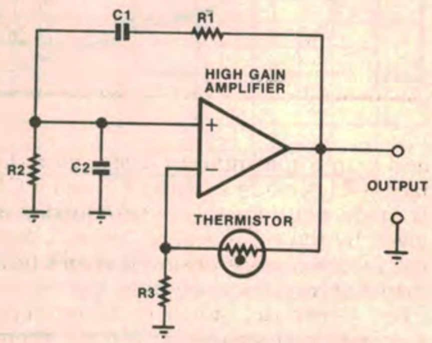


Fig. 1: basic scheme for a Wien bridge oscillator.

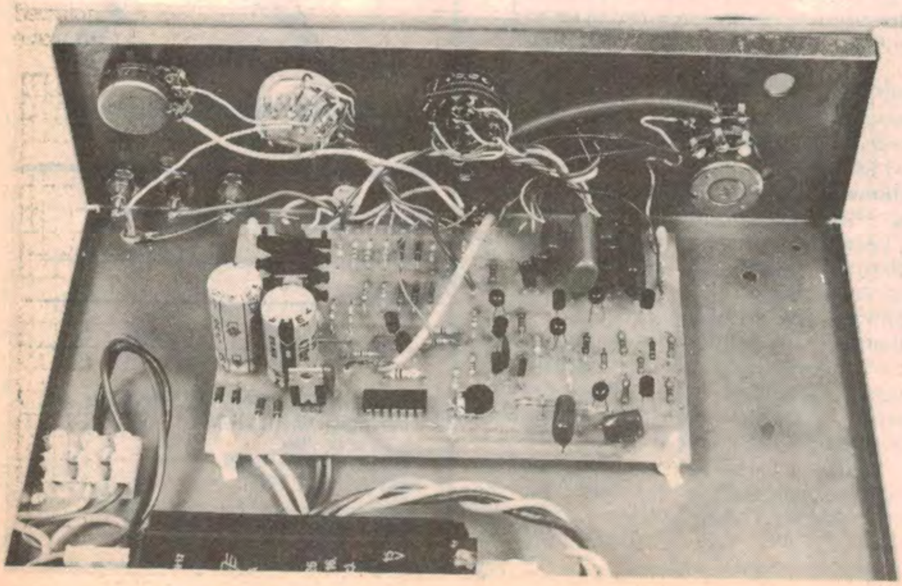
Audio oscillators are among the most useful pieces of equipment a hobbyist can have, and this is reflected in the number of oscillator projects we have published in the past. These have ranged from complex transistor designs to single IC circuits. Not for many years, however, have we designed a really high performance audio oscillator, and since there are few commercially high performance oscillators available at reasonable cost, we felt well justified in presenting this design.

The frequency range is 15Hz to 150kHz, covered in four overlapping ranges with a variable adjustment in frequency via a calibrated dial. Maximum output voltage is 3V RMS and this can be adjusted down to 1mV via a seven position attenuator with 10dB attenuation steps. Output level is continuously variable within each 10dB range via an output level control. Output impedance is nominally 600Ω regardless of attenuator setting or output level adjustment.

Well so far there is nothing to get excited about, but one special feature of the oscillator design is its low distortion output, typically .003% including hum and noise. The distortion versus frequency of the oscillator is shown in an accompanying graph and it can be seen that the distortion remains fairly low right up to the top of the band where it reaches .014% distortion at 100kHz.

As an added feature we have also provided a square wave output which features excellent rise and fall times of 20ns from a 50Ω output socket as well as via the 600Ω output. The fast rise times, extremely small overshoot and minimum ringing of the square wave signal are attributable to the simple VMOS output stage used.

One application of these "fast" square waves is in measuring the slew rates of audio amplifiers. The equivalent slew rate of the square wave output is about 300V/μs which is the sort of figure required to test some modern amplifiers with quoted slew rates up to 200V/μs. Audio signals never approach this sort of figure of course, and the SR is just a measure of an amplifier's ability to cope with transients without generating distortion.



LEFT: View inside the prototype. Note that the position of the LM320T15 regulator has been altered since this photograph was taken.

Circuit description

The circuit uses nine transistors in a Wien bridge oscillator. This configuration offers low distortion plus simplified frequency control and is universally used in high quality audio oscillators.

Turning now to Fig. 1, we can discuss the operation of the Wien bridge oscillator. A frequency selective network is used to apply positive feedback to a high gain amplifier. This network is formed by R1, C1, R2 and C2.

At a particular frequency, this network has a "pseudo-resonance", where a signal applied to R1 at the output of the amplifier is transmitted back to the amplifier input without any phase shift,

and with a minimum of attenuation. In fact, if R1 is made equal to R2, and if C1 is made equal to C2, this frequency is given by the reciprocal of $2\pi R1C1$, and the feedback transmission loss falls to a minimum of 3.0.

For a feedback amplifier to produce sustained and steady oscillations, there must be positive feedback with zero phase shift at a particular frequency, while at the same time the overall loop gain must be unity. We can achieve these conditions in the present case by applying negative feedback to give the amplifier a gain of 3.0.

This is the purpose of the network formed by R3 and the thermistor. The thermistor serves a second purpose in this case though, and that is to stabilise the amplitude of the oscillations. If the thermistor was replaced by a fixed resistor equal in value to 2R3, the circuit would oscillate, but the oscillations would continue to increase in amplitude until clipping occurred.

This is obviously undesirable. The thermistor acts to prevent this however, because as the output signal rises the power dissipated in the thermistor increases, and its temperature increases. This causes the resistance of the thermistor to reduce, (it has a negative temperature coefficient), so that the amount of negative feedback is increased, and hence the gain is reduced.

The thermistor also ensures reliable starting of the oscillator, because when there is no oscillation, there is minimal power dissipation in the thermistor, so that the gain of the amplifier is quite high. As the signal level then increases, the thermistor acts to stabilise the amplitude. With the thermistor specified, and the value of R3 used, the final output amplitude is just a little over 3V RMS.

Looking at the main circuit now, transistors Q1 to Q9 form a high-gain amplifier, the 1kΩ fixed resistors and 10kΩ variable resistors from R1 and R2, while C1 and C2 are selected via switches S1a and S1b. The capacitors are in multiples of 10 from .001μF to 1μF, giving four frequency ranges.

The 10kΩ variable and 1kΩ fixed resistors provide an 11-to-1 variation in resistance and hence frequency within each range. This means that the frequency ranges overlap.

We used a dual 10kΩ linear potentiometer for frequency adjustment because these provide reasonable tracking between the two gangs. This is important because small differences in the resistance of each gang will alter the attenuation of the Wien bridge and hence require a change in the gain of the amplifier. Due to the thermal lag of the thermistor it will therefore take the oscillator a few seconds to settle down whenever the frequency is changed (this is the characteristic "boinging" of thermistor controlled oscillators).

The amplifier itself is a fully complementary design which gives excellent linearity, and because it is complementary, second harmonic distortion is greatly reduced. The circuit consists of an NPN differential pair, Q1 and Q2 and a PNP differential pair, Q3 and Q4. The inputs of the differential pairs are connected in parallel and the output of each pair drives a separate transistor in the output stage.

Each differential pair uses a 15kΩ resistor as a current source "tail". Additionally, 680Ω resistors are included in series with each emitter to improve

SPECIFICATIONS

FREQUENCY RANGE: 15Hz to 150kHz in four ranges

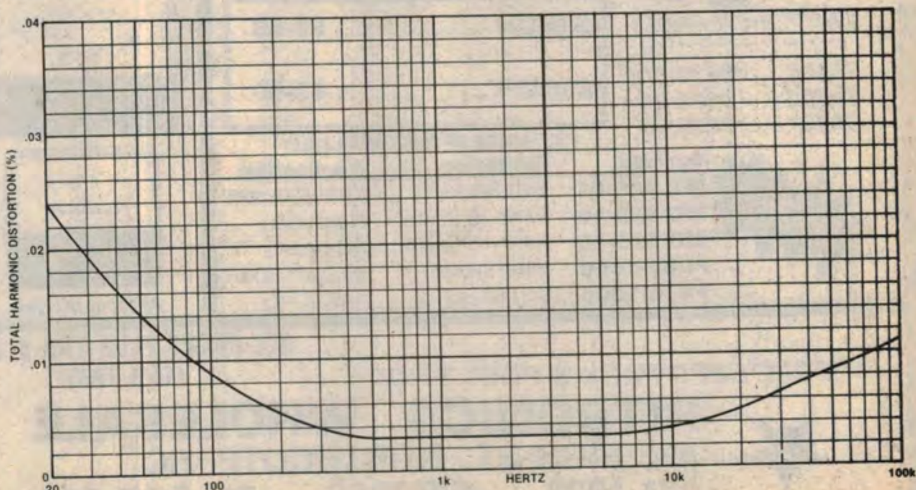
OUTPUT ATTENUATOR: 10dB steps viz 3V, 1V, 300mV, 100mV, 30mV, 10mV, 3mV.

OUTPUT LEVEL ADJUSTMENT: 10dB range

DISTORTION: .003% midband, .011% at 100kHz, .024% at 20Hz (see curves)

AMPLITUDE STABILITY: ±1dB

SQUARE WAVE: less than 20ns rise and fall times



This graph plots the total harmonic distortion against the output frequency.

the linearity of each differential pair but more importantly to provide high frequency stability.

Output from each differential pair is via a 3.3kΩ collector resistor with the NPN pair's output driving Q5 and the PNP pair's output driving Q6. The collectors of Q5 and Q6 are in turn connected together via a "Vbe-multiplier" circuit Q7. Quiescent current through each differential pair is about 0.5mA giving a voltage across each collector load resistor of 1.65V which results in a quiescent current through Q5 and Q6 of about 5.5mA.

The output from the collectors of Q5 and Q6 is an amplified and inverted version of the input signal and can swing symmetrically between the ±15V supply rails. This output is buffered by Q8 and Q9 which form a complementary-symmetry voltage follower.

To reduce crossover distortion in the output stage a quiescent current of about 5mA is passed through Q8 and Q9. The actual value of the current is determined by the fixed voltage across the "Vbe-multiplier", Q7 and the voltage drop across the two 22Ω resistors due to the quiescent current. Note that, due to variations in the Vbe of transistors Q7, Q8 and Q9, the 1.2kΩ resistor in the Vbe multiplier may have to be changed to obtain the required 5mA quiescent current. More about this later in the article.

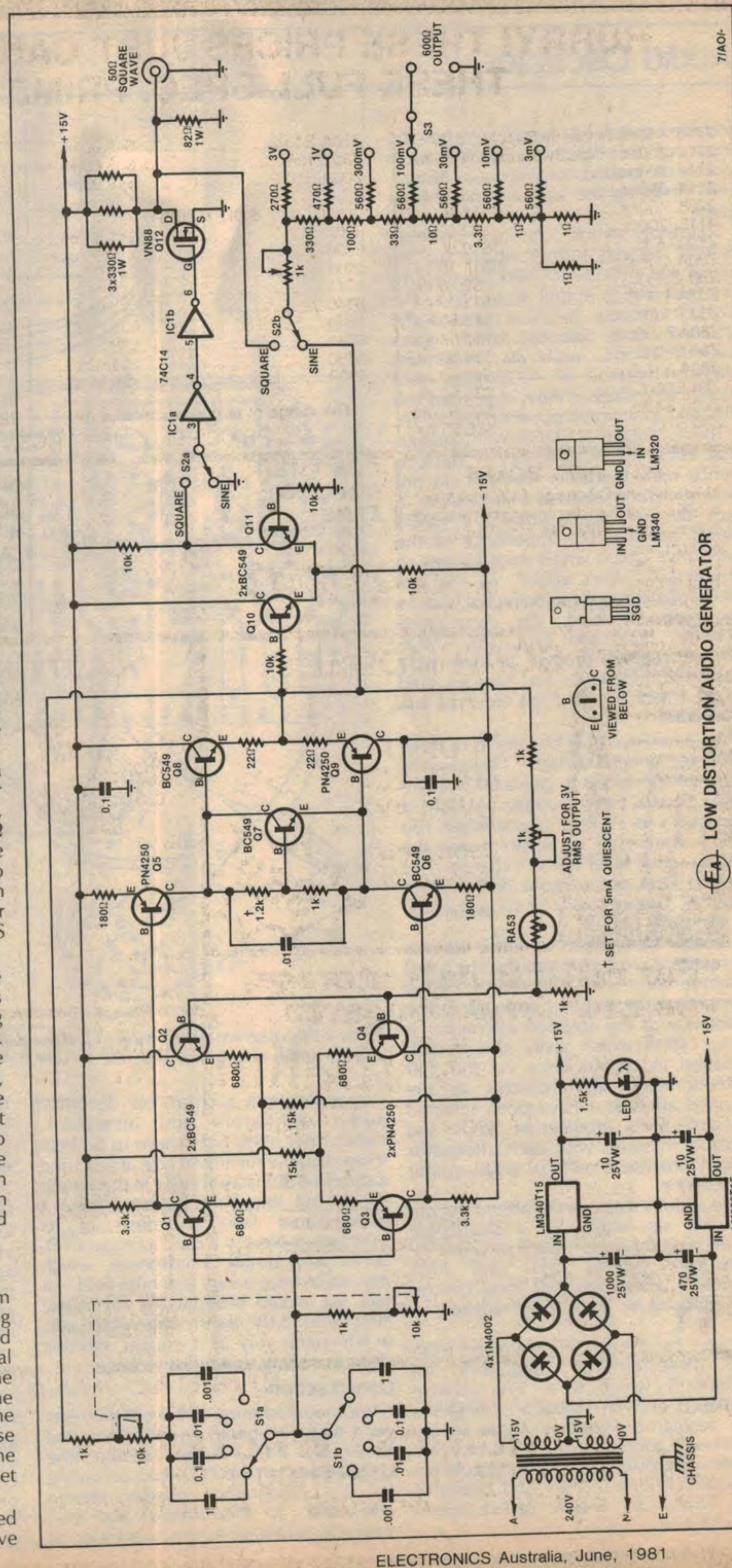
Feedback, as we have mentioned, is via a thermistor in a voltage divider. Rather than use a thermistor by itself however, we have included a 1kΩ fixed and a 1kΩ variable resistor which permits the output voltage level to be adjusted to precisely 3V RMS. (The output level can be quite easily set using a multimeter since most are calibrated to read RMS value for a sinewave signal.)

Turning to the distortion curve of the oscillator for a moment, note that it rises at both low and high frequencies, being .011% at 100kHz and .024% at 20Hz. Two quite different mechanisms are involved here. At the low frequencies, the thermistor itself generates the distortion because its thermal lag is not sufficiently long, causing its resistance to vary during any low frequency sinewave cycle, resulting in distortion. At high frequencies, the reduced open-loop gain is responsible for the increased distortion.

Square wave derivation

The square wave signal is derived from the sinewave output firstly by squaring the signal up via transistors Q10 and Q11. These are arranged as a differential pair with one input going to the sinewave output of the oscillator and the other to ground via a 10kΩ resistor. The 10kΩ resistor in series with the Q10 base is to reduce any loading effects while the 10kΩ in the other base reduces the offset voltage due to biasing currents.

Because of the fast spikes associated with the square wave circuit we have



included switch S2a which switches the input of the following square wave circuit to ground for sinewave output, thus disabling the square wave output stage.

Following switch S2a, two inverting Schmitt triggers further square up the signal and provide a direct interface to the final VMOS output stage, Q12. The VMOS transistor we have used is the VN88AF from Siliconix which offers specifications such as minimum transconductance of 150mmhos, gate threshold voltage of 0.8V, maximum Vds 80V and a minimum current rating of 1A.

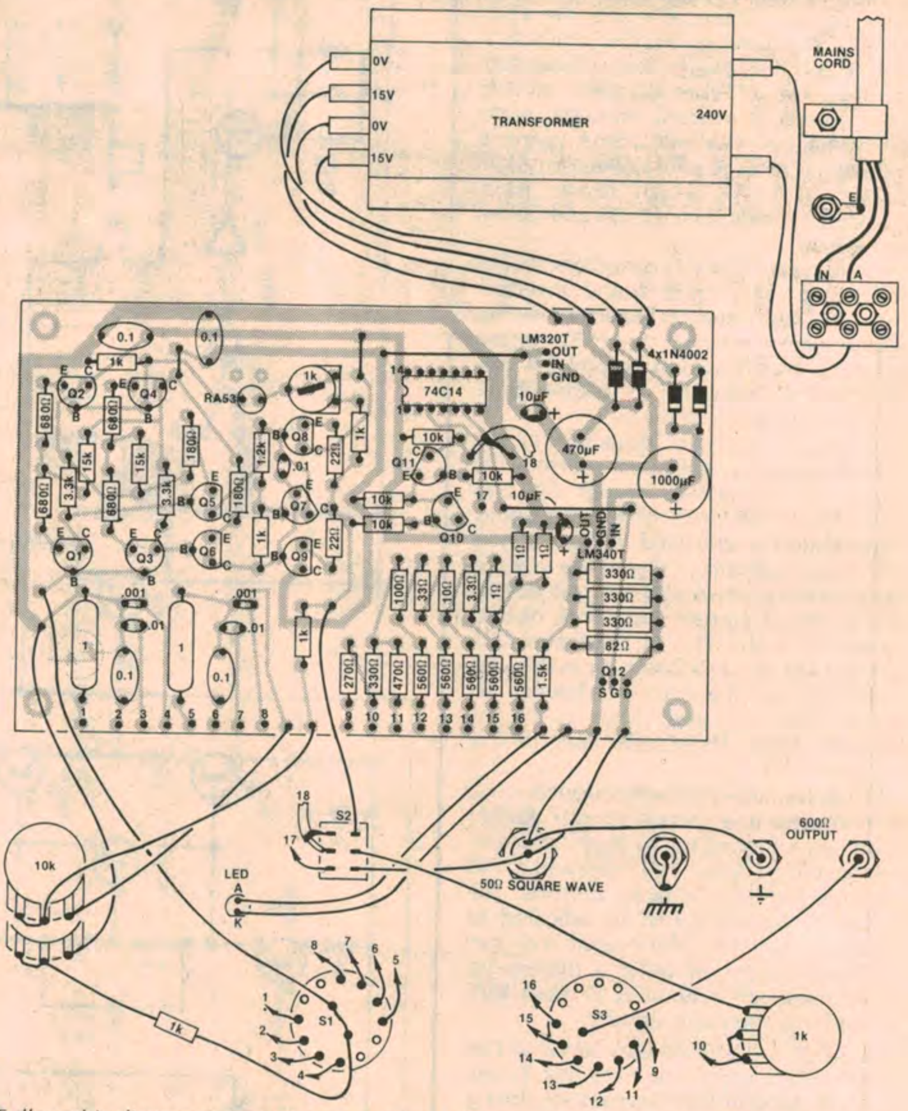
The advantage of VMOS over bipolar is that FETs are majority-carrier devices, hence minority-carrier lifetimes do not slow switching speed as with bipolars. In fact, the only thing slowing a VMOS circuit down is the impedance of the gate drive circuit which must be low to counteract "Miller effect", ie, the fast changing drain voltage being fed back to the gate via Cgd, the capacitance between drain and gate. The CMOS Schmitt triggers provide a reasonably low input impedance and have enabled us to obtain rise and fall times of less than 20ns.

The drain load of Q12 consists of three 330Ω and one 82Ω resistor. These values were selected so as to give a 6V peak-to-peak square wave output, which is equivalent to 3V RMS; hence when you switch from sine to square wave the output signal level doesn't change. This square wave output goes directly to a 50Ω BNC output socket or via switch S2b to the 600Ω output.

Variable attenuation of either the sine or square wave is accomplished by a 1kΩ potentiometer and the resistor ladder comprising the stepped attenuator. The resistance of the stepped attenuator is about 480Ω which gives about 10dB variation in output level via the 1kΩ potentiometer. The resistors in the stepped attenuator have been selected to give 10dB attenuation steps and resistors in series with each attenuator output provide a nominal 600Ω output impedance.

The various steps on the attenuator are selected via switch S3, a single-pole seven-position switch which provides 3V maximum output and 3mV RMS minimum. Using the 1kΩ level control the output can be further reduced to 1mV RMS.

Power for the circuit is derived from two three-terminal regulators, an LM340T-15 which is a 15V positive regulator, and an LM320T-15 negative 15V regulator. Together, these supply ±15V with good regulation and very low ripple and noise. Input to the regulators is from a bridge rectifier and capacitor filter driven by a 30V centre-tapped transformer.



Follow this diagram when wiring up the oscillator and keep all mains wiring neat and tidy. A small aluminium heatsink should be fitted to the LM340T15.

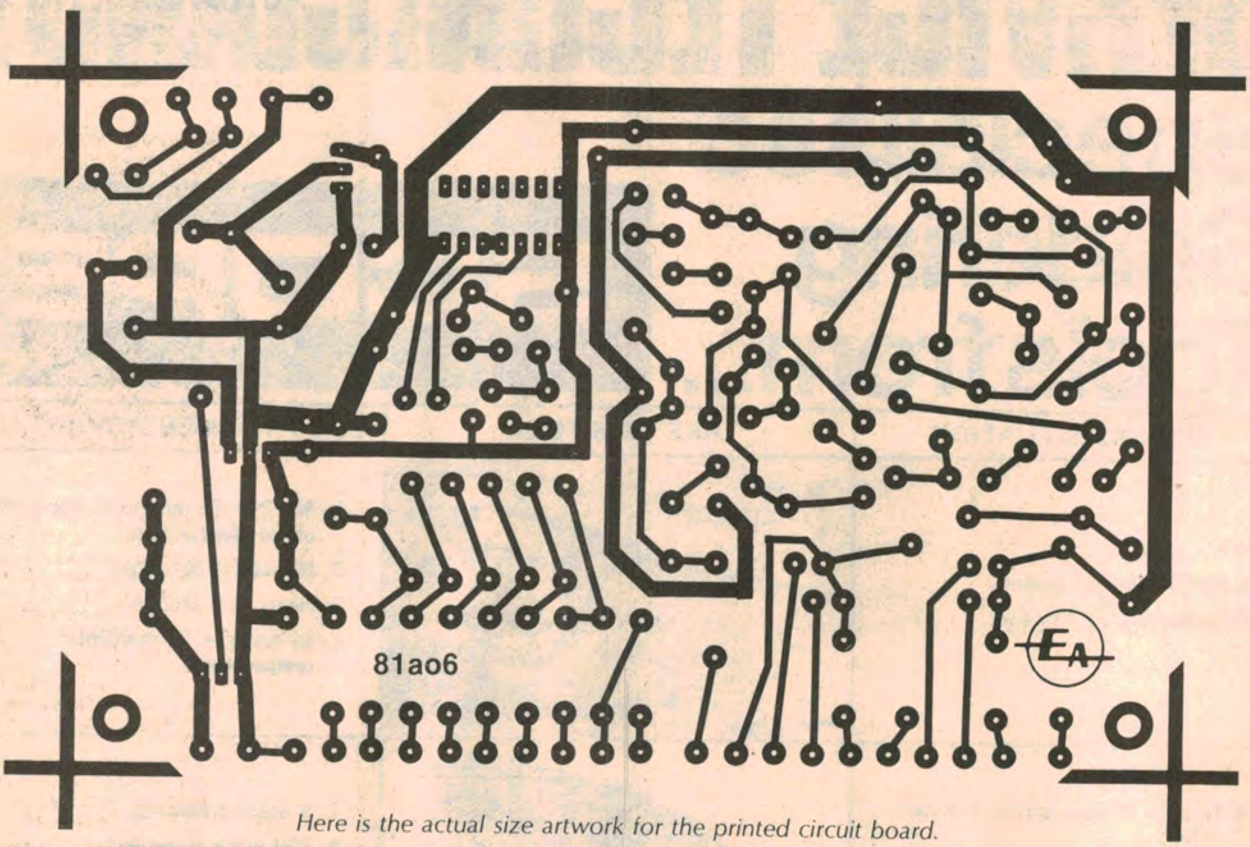
In applications such as low distortion audio oscillators and amplifiers, transformer hums fields have to be kept to an absolute minimum. We at first tried a standard 30V transformer in the power supply but found this radiated quite a strong hum field regardless of its orientation. Next, we used a Ferguson PL 30/40 low profile transformer which proved to have a very low hum field – a fact we would attribute to the metal strap around the transformer which acts in the same way as a copper shorting strap in cutting down flux leakage.

Construction

We mounted most of the components on a single PC board coded 81ao6 and measuring 145 × 91mm. Mount the components on the PC board according to the component overlay shown elsewhere in this article and pay particular attention to the orientation of the transistors, diodes and electrolytics. Note that the VMOS transistor and the CMOS IC can be damaged by static electricity so take the usual precautions.

We housed our audio oscillator in a Musicolour chassis (available from most kit retailers), though any metal case of similar dimensions would be appropriate. Drill mounting holes for the transformer, PC board etc, using internal photographs of our unit as a guide, and make sure that when the unit is fully assembled the PC board does not interfere with the front panel controls.

No power switch is provided on this unit as mains wiring has to be kept well away from the front panel, though one could be fitted to the rear of the unit if desired. The mains cable should enter on the rear of the chassis close to the transformer, passing through a grommetted hole, and be securely



Here is the actual size artwork for the printed circuit board.

clamped and terminated in a 3-way terminal block. The earth lead should be soldered directly to a lug which is securely bolted to the chassis.

To avoid possible earth loops we have left the oscillator outputs floating with respect to chassis and mains earth and provided a separate chassis earth connection on the front panel next to the 600Ω outputs.

To ensure the outputs are floating we have specified an insulated BNC socket for the 50Ω square wave output.

The front panel can be made up from a sheet of 18G aluminium 90 × 275mm. Finished Scotchcal aluminium labels for the front panel and dial are available from Radio Despatch Service, 869 George St, Sydney or Rod Irving Electronics, 425 High St, Northcote, Victoria. We have not included actual-size front panel artwork with the article because it is too big to fit on the page.

The dial was calibrated using a 10kΩ linear pot which explains why the markings are crowded at one end (ideally we would have used an "inverse law" potentiometer but these are not readily available). Affix the Scotchcal label for this dial to, say, a 20-gauge piece of aluminium, which can then be cut to the correct shape and glued to a suitable knob, ie one which matches the other control knobs.

PARTS LIST

1 metal case (see text)
1 PC board coded 81ao6, measuring 145 × 91mm
1 mains transformer Ferguson PL 30/40
1 single-pole, seven-position rotary switch
1 DPDT miniature toggle switch
1 2-pole (or 3-pole) 4-position rotary switch
4 Richco CBS-6N PCB supports
1 dual ganged 10kΩ linear potentiometer
3 terminal posts and banana sockets, green, red, black
1 insulated BNC socket
1 mains cord and plug
1 mains cord clamp and grommet
1 1kΩ (lin) potentiometer
1 1kΩ miniature horizontal trimpot

SEMICONDUCTORS
1 74C14 CMOS hex Schmitt trigger
1 LM340T-15 positive regulator
1 LM320T-15 negative regulator
1 VN88AF VMOS transistor
7 BC549 NPN transistors
4 PN4250 PNP transistors
4 1N4002 diodes
1 small red LED
1 RA53 thermistor (made by ITT)

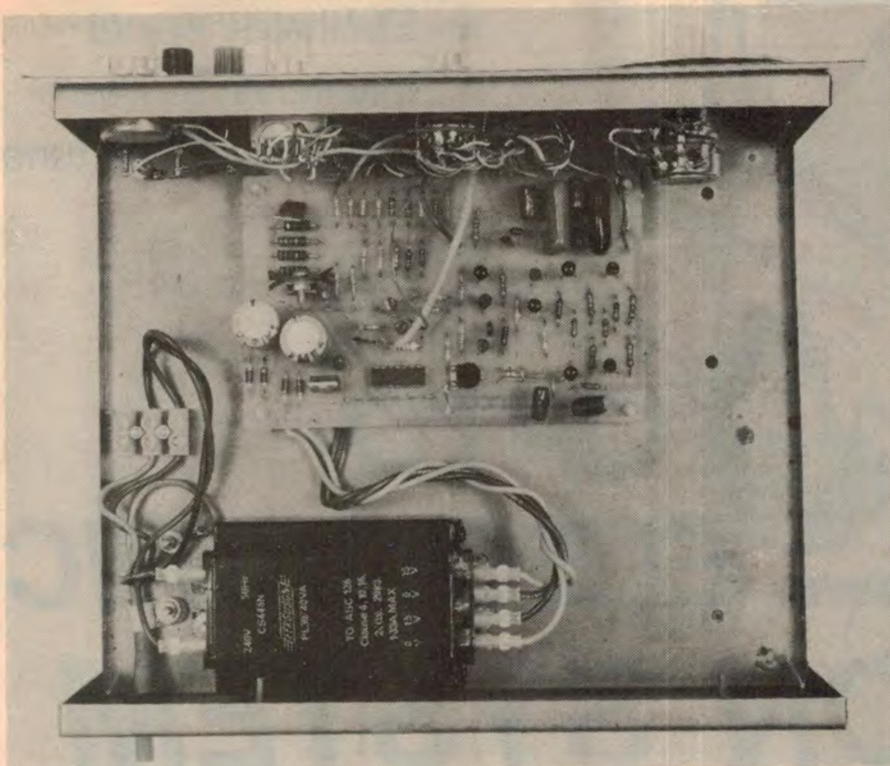
CAPACITORS
1 1000μF/25VW PC electrolytic
1 470μF/25VW PC electrolytic
2 10μF/25VW tantalum
2 1μF greencap (metallised polyester)
4 0.1μF greencap
3 0.01μF greencap
2 .001μF greencap

RESISTORS (all ¼W 5%):
2 × 15kΩ, 4 × 10kΩ, 2 × 3.3kΩ, 1 × 1.5kΩ, 1 × 1.2kΩ, 5 × 1kΩ, 4 × 680Ω, 5 × 560Ω, 1 × 470Ω, 1 × 330Ω, 1 × 270Ω, 2 × 180Ω, 1 × 100Ω, 1 × 33Ω, 2 × 22Ω, 1 × 10Ω, 1 × 3.3Ω, 3 × 1Ω
1W RESISTORS: 3 × 330Ω, 1 × 82Ω

Setting up

Wire up the board and front panel controls according to the wiring diagram keeping leads as short as possible. Make another check of the board and wiring then, switch on and check the +15 and −15V supplies. Next check that the quiesent current is about 5mA or slightly more. This can be accomplished by measuring the voltage across the 2 × 22Ωresistors (should be 220mV) or inserting a milliammeter in place of the link from the collector of Q8 to +15V.

If the quiescent current through Q8 and Q9 is less than 5mA, increase the



The PCB is mounted in the chassis using Richco plastic standoffs.

1.2kΩ resistor connected to the base of Q7 to say 1.5kΩ or greater. Similarly if it's more than 5mA, reduce the 1.2kΩ resistor.

Now set the attenuator to 3V RMS and adjust the output level control to maximum. Connect a multimeter to the output with the unit switched to sinewave output and adjust the 1kΩ trimpot for a 3V output at 1kHz.

We estimate that the current cost of parts for this project is approximately

$70

including sales tax.