PHASE LOCKED LOOP IC FLUTTER METER

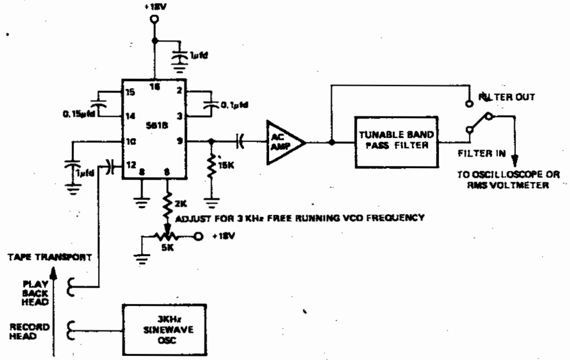
Every day, new applications are being found for the integrated circuit. One interesting use in the audio test field has been published by Signetics: a tape recorder flutter meter using a 'phase locked loop' IC. Signetics, a leader in the development and manufacture of integrated circuits, includes the flutter meter circuit in their recent PLL applications book.

According to the circuit's designer, "the Signetics PLL 561B is used to detect the frequency variations of the playback 3kHz tone. The VCO frequency is set to a nominal 3kHz by C_o and fine tuning trimmer. The demodulated output is AC coupled to a high input impedance amplifier. An oscilloscope can be used to measure peak deviations, and a true RMS voltmeter is used to make RMS flutter readings.

"The output may be calibrated by feeding in a 3 kHz tone from an oscillator, and offsetting the frequency by 1% and measuring the output level shift. Good recorders have RMS flutter of less than 0.1%. The output can be filtered to study selected frequency bands.

"Speed variations in the movement of tape across the heads in a tape recorder cause the playback frequency to vary from the original signal being recorded. These speed variations are caused by mechanical problems associated with the tape drive and tape guidance mechanisms. The variation in frequency of the playback signal is called flutter and is generally measured over a frequency range of 5 to 200 Hz.

"Test tapes with low recorded flutter variations are available to test playback mechanisms. These tapes are standardized at 3 kHz. With systems equipped with record heads, a 3 kHz tone can be recorded for analysis." ■



An I.C. Peak Programme Meter

by L. Nelson-Jones, F.I.E.R.E.

A design using standard i.c. operational amplifiers to achieve a transformerless design to the specification of the British Broadcasting Corporation, who pioneered this type of level indicator. Mono or stereo applications are catered for in the design, with separate or common meter indication. The circuit is stable against temperature and supply voltage variations, and is designed for use with a nominal 24V supply (16–30V). The main design aims were to obtain accuracy, stability, ease of law adjustment, and repeatability from one unit to another.

The peak programme meter dates back some 36 years when it was developed to provide a better means of measuring line levels in sound broadcasting than that provided by normal rectifier instruments such as the VU meter. In particular the instrument was given a slow decay and fast attack time to ease reading and lessen eye strain. Early designs were characterized by a very rapid response to transient peaks, but this was later modified since it was found that in practice the ear cannot easily detect the distortion produced by the clipping of very short duration peaks. The final attack time figure decided upon, and which is still standard, was 2.4 milliseconds. Such a response corresponds to a meter reading reaching 80% of peak using a square wave transient lasting 4 milliseconds. The decay time constant used is 1 second, which is a compromise between ease of reading and a response quick enough to record following peaks.

The graduations on the indicating meter were kept small in number and a black scale with white markings used to make for ease of reading. The basic scale division was chosen as 4dB, this being two steps of the standard B.B.C. fader controls. On a standard meter there are basically 7 divisions, with division 4 corresponding to 0dBm on a 600Ω line (0.775V r.m.s. sine wave, 1.095V peak).

The response of the peak programme meter (PPM) is approximately logarithmic and the divisions on the meter are approximately evenly spaced. The extreme divisions (1 to 2, and 6 to 7) represent a greater change than 4dB, namely 6dB. (Earlier meters differed in having all divisions except 1 to 2 equal to 4dB.) The present standard calibration together with the corresponding current in the meter are shown in Table 1.

The meter figures given are for B.B.C. Meter Specification ED1477, the one chosen for the design to be described.

In order to make good use of a fast charge time, the dynamic qualities of the moving-coil meter movement itself must be tightly controlled and considerably faster than that of normal movements. The meter must also be correctly damped to avoid large overshoots—two rather conflicting requirements. Whilst PPM circuits will work with standard meter movements the accuracy will be somewhat impaired unless the correct movement is used. In particular a circuit using a normal meter movement will, when set up on a standard tone level, tend to seriously underestimate short peaks on actual programme material.

Peak detection

In most previous PPMs a normal full-wave rectifier has been used, (Fig. 1), with a centre tapped signal transformer; the charge and discharge time constants being controlled by the two resistors *r* and *R*.

With the advent of integrated circuit operational amplifiers, however, one can now make an accurate peak rectifier without the need to use large voltage swings in order to overcome the forward drop of the rectifier, and the consequent non-linearity at low levels.

The basic circuit of such a peak detector is shown in Fig. 2. On a rising positive input, the output of the op-amp rises positively until the signal fed back to the inverting input of the op-amp via the diode *D* equals the level at the non-inverting input of the op-amp. When the input level falls, the diode *D* ceases to conduct as it becomes reverse biased, and the previous peak is stored on the capacitor *C* until such time as the input rises above the voltage to which the capacitor is charged, when the voltage on the capacitor again follows the input.

In practice the author has modified the basic circuit of Fig. 2 to that of Fig. 3. Apart from the two resistors *r* and *R*, to control the charge and discharge time constants, a transistor has been added to ensure adequate charging current availability. The practical values of the components are $C = 33\mu F$, $r = 75\Omega$, $R = 30k\Omega$. With such a large capacitance the peak charging current through the diode reaches approximately 100mA, which is well above the

Table 1.

PPM reading	Level dBm	Input voltage (peak)	Meter current (mA)
0	—	0	0
1	−14	0.220	0.10
2	−8	0.436	0.22
3	−4	0.690	0.35
4	0	1.095	0.51
5	+4	1.74	0.67
6	+8	2.75	0.80
7	+14	5.50	0.93
f.s.d.	—undefined		1.00

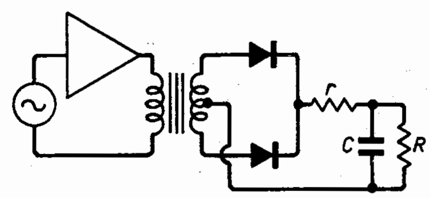


Fig. 1. Conventional PPM using centre-tapped transformer and full-wave rectifier.

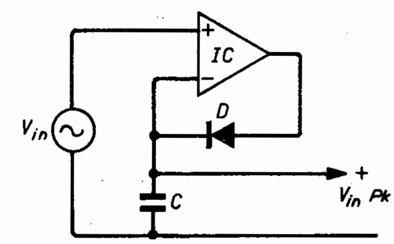


Fig. 2. Peak detecting circuit.

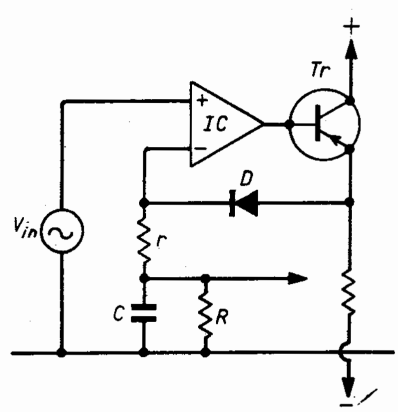


Fig. 3. Peak detecting circuit with time constants added.

capability of a normal i.c. op-amp, hence the additional transistor. (In practice r will be less than 75Ω due to the necessity of allowing for the forward impedance of the diode, and other components in the 'charging' path.)

The peak detector described operates only on positive peaks, whereas in a practical PPM it is necessary to measure positive and negative peaks equally, and to this end it is necessary to either (a) have a similar peak detector of reversed polarity to detect the negative peaks or (b) to have a second similar peak detector and precede it with a unity gain phase inverter.

It was decided to take the second course since it allowed the two positive peak detectors to be combined, sharing a common capacitor, charge and discharge resistors. In this way the highest peak from either detector will automatically be selected.

The unity gain inverter can of course be a centre tapped transformer as in previous PPMs, or else another op-amp connected as a unity gain inverter, as shown in Fig. 4.

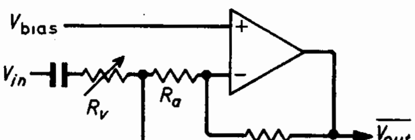


Fig. 4. Unity gain inverting circuit.

Provided that the loading on the two outputs is small, both the a.c. and d.c. levels will be equal except for the phasing. Since the input to the two peak detectors is the non-inverting input of two op-amps the loading is in fact quite low. The difference of d.c. level due to the unequal base supply resistances of the two peak detectors is approximately equal to the typical offsets of the i.cs, and is therefore fairly negligible when compared to the signal levels, i.e. they are less than 10% of the lowest division (1 on the PPM = 0.22V pk). In addition there is a zero set control on the output amplifier which can largely remove the effect from the meter deflection.

Gain adjustment is achieved by the single control R_V for both peak detectors. Whatever the value of R_V, V_{out} and $\overline{V_{out}}$ (Fig. 4) will remain equal and opposite to one another so far as signal excursions are concerned, although at the same d.c. level.

Law corrected output amplifier

The voltage across the peak storage capacitor is applied to a law corrected summing amplifier, whose input resistance (and hence the discharge time constant) will be set by an input resistor R_x to the summing point. The basic principle of this amplifier is illustrated in simplified form in Fig. 5.

The initial gain for voltages close to the bias voltage line (V_{bias}) that is from 0–3 on the PPM scale, is linear, and is set by the ratio R_y/R_x since for small output levels the transistor Tr_1 is reverse biased. When the emitter potential of Tr_1 falls below its bias voltage V_1 the additional feedback

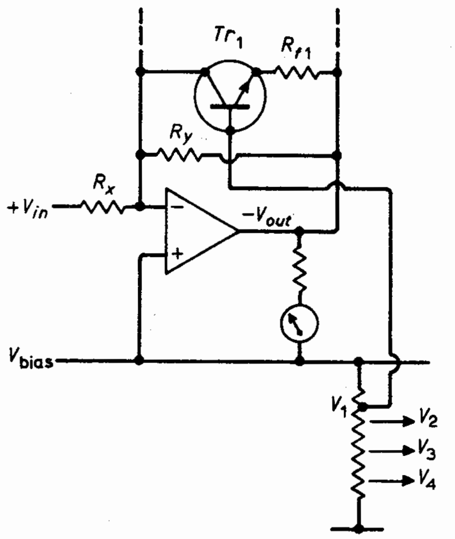


Fig. 5. Use of a transistor in the feedback path to provide law correction of transfer characteristic.

resistor R_{f1} is brought into operation in parallel with R_y so that the gain is reduced to

$$\frac{R_y \cdot R_{f1}}{R_x(R_y+R_{f1})}.$$

Further feedback resistors R_{f2}-R_{f4} together with transistors Tr_2-Tr_4 each controlled by bias voltages V_2-V_4 respectively, are similarly connected to the amplifier to successively reduce the gain with increasing negative output level. The law corrected amplifier therefore approximates the desired curve of input versus output with a five section linear gain curve as shown in Fig. 6. The choice of feedback resistors and bias voltages is made to get the best match to the actual smooth curve. In practice this was done by graphical methods together with calculation. The values were finally adjusted by trial and error to get the best result, together with the use of standard E24 values. The choice of values possible is almost infinite depending on the choice of break points.

Complete practical circuit

Fig. 7 shows the complete circuit of the peak programme meter based on the circuits described above. It is designed to work with a 1mA meter movement to B.B.C. Specification ED1477.

There are a few items in this circuit not covered in the above circuit descriptions. First, in the feedback network of the law corrected amplifier a diode has been added to prevent any appreciable positive excursion of the amplifier's output on switching on or off. Secondly, a zero set potentiometer is added to this amplifier to take out the combined zero errors of the four op-amps which although small enough to hardly affect the working accuracy is nevertheless rather annoying visually in the absence of an input level.

The zero set potentiometer is the usual value for the type 741 op-amp but is connected in a somewhat different manner. Instead of being connected between the two offset points of the 741 and the negative supply line, a resistor is connected to the slider of the potentiometer and returned instead to the 9.1V bias line. This arrangement allows a much wider range of adjustment than the usual connection, which although adequate to cope with the offset of one 741 op-amp is not sufficient to cope with the combined offset of four op-amps if these should unfortunately be additive.

The d.c. operating level of all stages is determined by the bias supply of +9.1V stabilized by the zener diode D_4 which also supplies the bias chain for the output amplifier's feedback network. This bias chain has an overall adjustment in order that the exact law correction of the completed instrument may be set up, and the tolerances of the various elements allowed for—in particular that of the zener diode stabilized voltage.

The 1mA meter to B.B.C. Specification ED1477, has a resistance of 600Ω ±5% so that with its series resistor of 4.7kΩ (R_{15}), full-scale deflection corresponds to −5.3V with respect to the +9.1-volt bias line. Maximum overdrive of the meter is limited therefore to a little less than the bias line

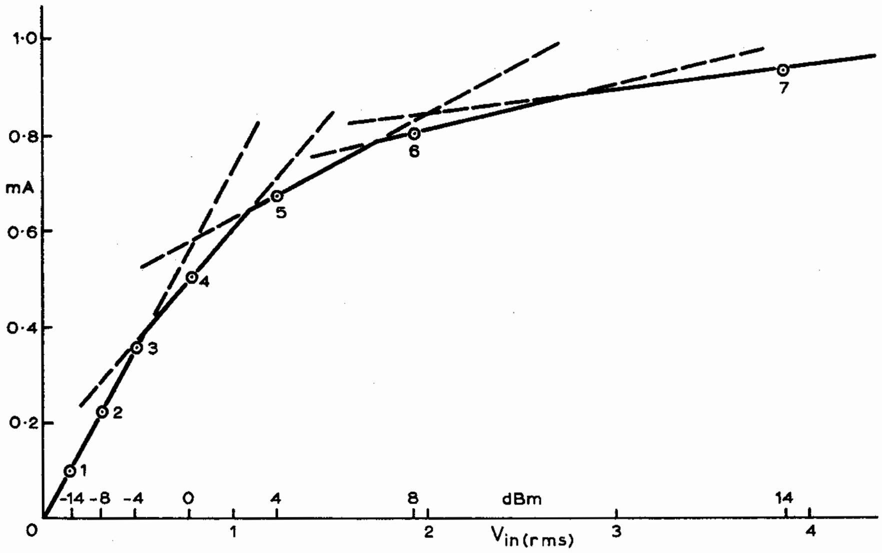


Fig. 6. Low corrected transfer curve approximated by a five section linear characteristic.

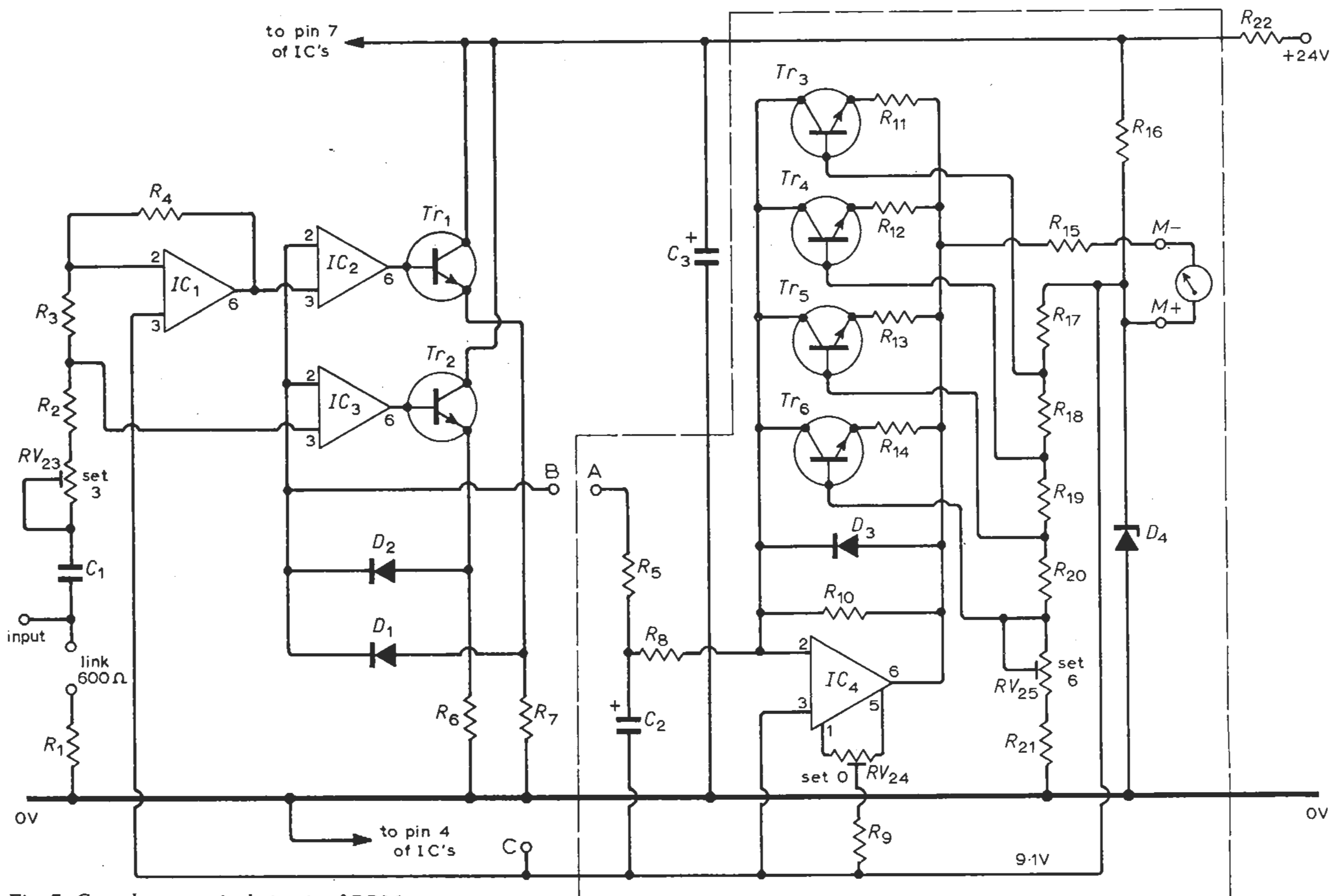


Fig. 7. Complete practical circuit of PPM.

voltage, or approximately some 8V, corresponding to roughly 150% overdrive—a reasonable value for meter protection. The general action of the circuit normally prevents reverse deflection, but in any case the diode in the feedback circuit prevents more than 0.6V being applied to the meter, corresponding to −11% deflection.

Due to the very high peak currents occurring in the peak rectifier circuit, particularly in the collector currents of Tr_1 and Tr_2, some measure of isolation from other equipment sharing the same supply line is necessary. To this end decoupling by R_{22} and C_3 is provided.

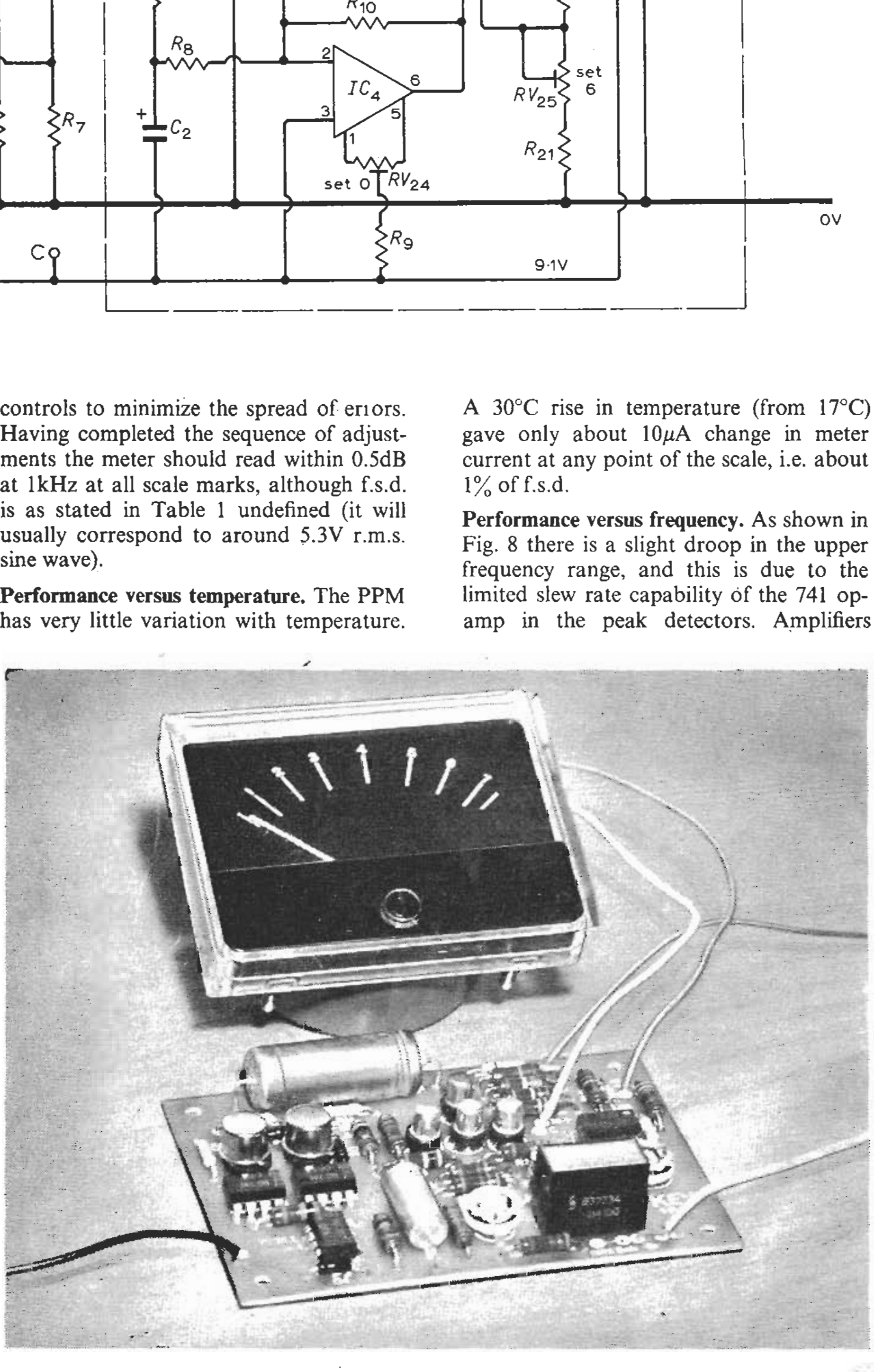
A resistor (R_1) of 620Ω is included so that it can be linked into circuit to give a line terminating impedance of 600Ω instead of the normal line bridging input impedance of around 16kΩ.

Setting-up and performance

The procedure for setting-up the PPM is a simple one. First, with zero input voltage, the zero is set ('Set 0' control) RV_{24}. Next a level corresponding to −4dBm (reading 3 on the PPM scale) or 490mV r.m.s. sine wave, 690mV peak, is applied and the 'Set 3' control (RV_{23}) is adjusted to bring the meter pointer to 3 on the scale. Finally a level of +8dBm, (reading 6 on the PPM scale) or 1.94V r.m.s. sine wave (2.75V peak), is applied to the input and the 'Set 6' control RV_{25} is adjusted to bring the meter pointer to 6 on the scale. The meter is then checked at 0, 1, 2, 3, 4, 5, 6, 7, and f.s.d. points as listed in Table 1, and any small adjustment made to the 'Set' 0, 3, and 6 controls to minimize the spread of errors. Having completed the sequence of adjustments the meter should read within 0.5dB at 1kHz at all scale marks, although f.s.d. is as stated in Table 1 undefined (it will usually correspond to around 5.3V r.m.s. sine wave).

Performance versus temperature. The PPM has very little variation with temperature. A 30°C rise in temperature (from 17°C) gave only about 10μA change in meter current at any point of the scale, i.e. about 1% of f.s.d.

Performance versus frequency. As shown in Fig. 8 there is a slight droop in the upper frequency range, and this is due to the limited slew rate capability of the 741 op-amp in the peak detectors. Amplifiers



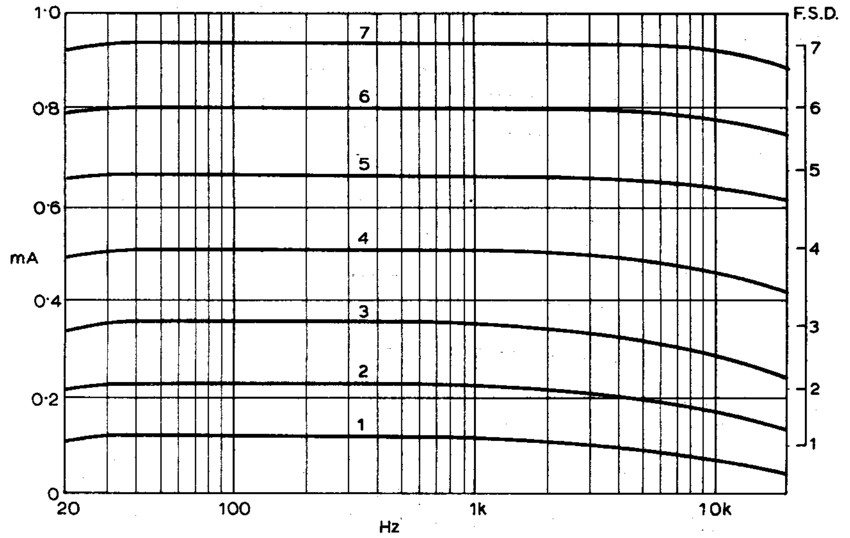
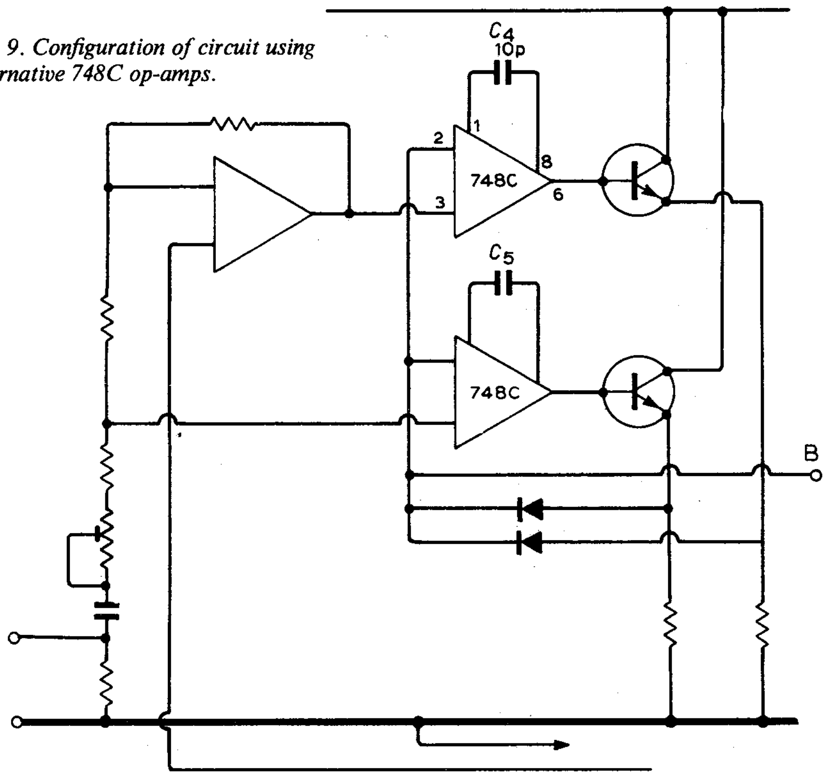


Fig. 8. Frequency response of prototype PPM.

Fig. 9. Configuration of circuit using alternative 748C op-amps.



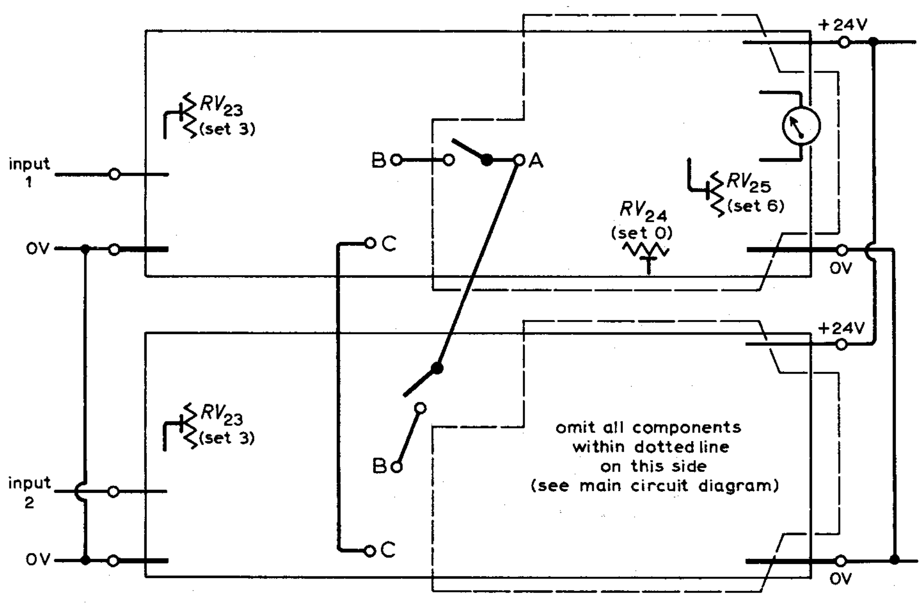


Fig. 10. Outline of printed circuit panel for mono or stereo.

having a higher slew rate have been tested and do remove this limitation in the audio range. The 748 op-amp has a higher speed performance than the 741 but uses external compensation; this allows the response to be tailored to suit any particular need. Fig. 9 shows how two 748 op-amps may be used for IC_2 and IC_3, together with appropriate extra components to obtain a flat frequency response over the whole audio band. There is still a slight fall off at 20kHz but this is greatly reduced as compared to the 741 op-amp.

In practice does this h.f. droop matter? The author would argue that for the monitoring of practical speech and music levels it does not matter to any noticeable extent. This is because of two factors. First, there is the attack response time of 2.5ms used in the circuit, meaning that a level must last for several milliseconds to register near to its true peak level, and secondly, in general, frequencies above about 5kHz do not exist at as high a level as the lower frequencies, and these lower frequencies therefore largely determine the peak amplitude at any time.

Performance versus supply voltage. Over the range of 16 to 30V there is little visible change of reading at any level of input. The circuit is designed for operation from a nominal 24-volt supply. Supply current is somewhat dependent on input level, and is typically 14mA at zero input, rises fairly rapidly as input is applied, and reaches 20mA at full scale. There will be some variation from unit to unit but at 24V the current should remain in the limits 13–22mA. The current demand is also dependent on supply voltage being lowest at 16V and highest at 30V. An absolute maximum supply voltage of 36V should never be exceeded.

Connections for stereo use with a single common meter

For economic or space reasons, it may be desired to use two PPM circuits with a common meter, and the printed circuits were designed with this in mind as an option. The method of interconnection is shown in Fig. 10 where two input circuits up to point *B* are used, with only one output circuit from point *A* onwards. The bias supply is made common to both boards by linking points *C* together.

To set up the meters in this method of connection the zero is first set at nil input level (to both inputs), 'Set 0' control (RV_2).

Next inputs of -4dBm are connected to each input in turn and the appropriate 'Set 3' control (RV_{23}) for that channel is set to give a reading of 3 on the PPM. Finally the 'Set 6' control is set to give a reading of 6 from either input at a level of $+8$dBm.

For the setting of the 'Set 3' and 'Set 6' controls both inputs may be connected in parallel and the switches shown in Fig. 10 operated to select the channel to be set up.

The dotted lines in Fig. 7 show the section of circuit omitted on one board and correspond to the dotted lines in Fig. 10.

Acknowledgements

The author wishes to express his grateful thanks to the B.B.C. Engineering Depart-

ment and in particular to Mr. A. E. Tolladay, for considerable help and encouragement, also to Ernest Turner Electrical Instruments Ltd. for their help in the project.

Constructional appendix

The circuit is built on a printed circuit board $3\frac{1}{2} \times 3\frac{1}{2}$ inches in size with mounting centres of 3.1×2.1 inches (6BA). The board, which is suitable for either the circuit with 741s in all stages or the higher speed circuit with 748s in the peak rectification stages, is shown in prototype form in the photograph. Layout of production boards will differ slightly but all component positions are silk screened onto the component side of the board.

It is essential that the charge storage capacitor C_2 be a low leakage type, hence the specification of a solid dielectric tantalum type. An alternative is the solid dielectric aluminium capacitor such as Mullard type 121 15339 (33μF 16V) or 121 16339 (33μF 25V). However, it should be remembered that these are of 20% selection tolerance, and it may be necessary to select one to the necessary tolerance of 10%. In general normal aluminium electrolytics are not suitable, due to their high leakage (especially at elevated temperature) and very wide tolerance, even of the higher quality type (e.g. Mullard C428 is -10 $+50$%).

No special techniques are used in the construction and the only precautions needed are to ensure correct insertion of the 8-pin dual-in-line op-amp packages, and to avoid shorts on the board due to careless soldering (a miniature soldering iron is, these days, essential for printed circuit work). Mounting pads are used under the 6 transistors but are not absolutely essential. Connections are by 14–0076 p.v.c. covered leads as shown in the photograph.

All component parts in kit form together with Ernest Turner PPM meters type 642 are available from Key Electronics, P.O. Box No. 7, Bournemouth, BH7 7BS, Hants.

Components list

Resistors

R_1	620	R_{13}	56k
R_2	2.2k	R_{14}	15k
R_3, R_4	10k	R_{15}	4.7k
R_5	68	R_{17}	1.2k
R_8	30k	R_{18}	620
R_{10}	160k	R_{19}	560
R_{11}	220k	R_{20}	270
R_{12}	120k	R_{21}*	

All the above are 2% metal oxide or metal film (e.g. Welwyn MR5 or Electrosil TR5).

*Resistor R_{21} will normally be a wire link. (For use only where a higher reference line voltage than 9.1V is to be used.)

R_6, R_7	22k	R_{16}	2.2k
R_9	220k	R_{22}	47

All the above are $\frac{1}{8}$W 5% carbon film (e.g. Iskra UPM033 or Mullard CR25).

Capacitors

C_1	1μF, 100V poly (15mm mounting centres)
C_2	33μF, $\pm$10%, 20V solid dielectric tantalum
C_3	220μF, 35V, aluminium electrolytic.
C_4, C_5	10pF, ceramic tube, disc or poly.

Transistors

Tr_1, Tr_2	BFY52 or 2N2219.
Tr_3 to Tr_6	BC109.
D_1, D_2	OA200, 1S920.
D_3	1S44, 1N914, 1N916.
D_4	BZY88–C9V1 (9.1V, $\pm$5%, 400mW).
IC_1	741C (8 pin d.i.l.).
IC_2, IC_3	741C (8 pin d.i.l.) or 748C for high-speed version.
IC_4	741C (8 pin d.i.l.).

N.B. The TO–99 versions (multi-lead TO–5), of the 741 and 748 may also be used since they have the same lead layout and are easily arranged in d.i.l. lead configuration. Jermyn Industries Ltd., type MON–8L mounting pad may be used to achieve this end.

Preset potentiometers

RV_{23}, RV_{24} 10k $\pm$20% RV_{25} 5k $\pm$20%

Open cermet potentiometers (R.S. Components or A.B. Electronic Components).

Semi-sealed type Morganite 81E may also be fitted (also from R.S. Components).

Announcements

Information Retrieval is the title of a course of six lectures to be held at Twickenham College of Technology on Monday evenings at 19.00 commencing 6th November. Details from The Principal, Twickenham College of Technology, Egerton Road, Twickenham, Middlesex TW2 7SJ. Fee £3.75.

Pickup arm and turntable parts. Longdendale Technological Products, suppliers of parts for the pickup arm and turntable designs published in the October and November 1971 issues, have moved to Wood End Industrial Estate, Manchester Road, Mossley, Manchester.

Arrow Electronics Ltd, 7 Coptfold Road, Brentwood, Essex, has been formed to provide a retail mail order distribution service for a range of components from opto-electronics to $\frac{1}{4}$W carbon resistors; temperature controlled soldering irons to aluminium boxes; test instruments to the latest double-wound toroid chokes.

Television Systems and Research Ltd, 63 Woodside Road, Amersham, Bucks, formed to acquire the business and assets of the Top Rank Television Division of Rank Audio Visual Ltd. have been appointed by TeleMation Ltd to be exclusive distributors in the U.K. and Ireland of the TeleMation range of closed circuit TV equipment.

Marconi Space and Defence Systems Ltd is to supply a **cockpit procedures trainer** to Flight Safety Inc., the professional pilot training organization in America, acting on behalf of Aeroformation. The trainer, to be completed by late 1973, will be installed in a new multi-million dollar Aeroformation Training Centre in Toulouse, Blagnac, France.

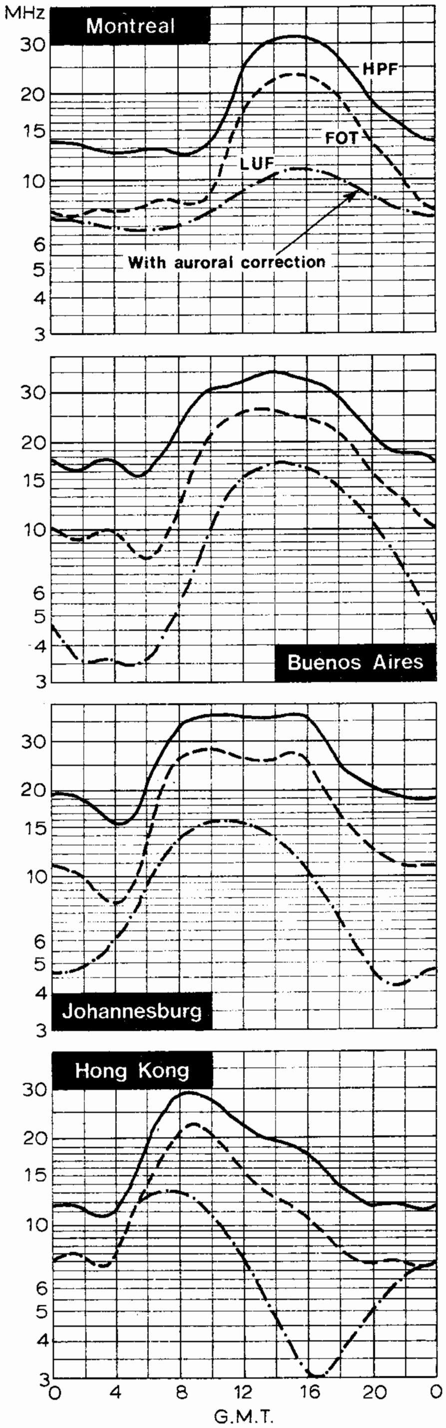
Marconi Communications Systems Ltd and Amalgamated Wireless (Australasia) Ltd are jointly to supply equipment for the Australian Broadcasting Commission's new **colour television service.** Scheduled to be operational by March 1975, Marconi are supplying four of their Mark VIII automatic colour cameras and ancillary equipment with an outside broadcast vehicle.

Consolidated Fisheries Ltd., Grimsby, are to fit their entire fleet of 15 distant-water trawlers with the new, world wide, **Redifon Omega Navigator.**

H.F. Predictions — November

Solar activity is still not declining as rapidly as expected. The past three years have shown a check in decline during each September which has been maintained. This could mean that the current sunspot cycle will be two years longer than the eleven year average giving a minimum in 1975/6.

With this relatively high activity the 26-MHz broadcast and 28-MHz amateur bands should be open between 08.00 and 16.00 GMT throughout the month to South Africa. These two bands will also be usable for South America but not so consistently and to North America only for very short periods, if at all.



MULTIPLIER/OP AMP CIRCUIT DETECTS TRUE RMS

To get an RMS value when you can't afford the time it takes to heat an element, try this technique. It may not be feasible for a multimeter but how about a sampling voltmeter good up to 600 kHz?

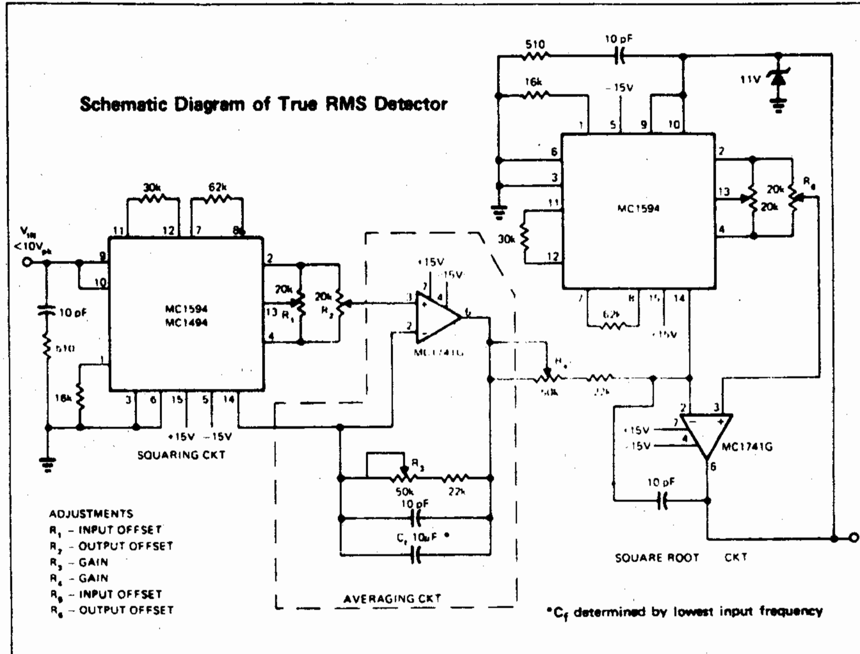
Mathematically, the RMS value of a function is obtained by squaring the function, averaging it over a time period I and then taking the square root:

$$V_{RMS} = \sqrt{\frac{1}{T}\int_{0}^{t} V^2 dt}$$

In a practical sense this same technique can also be used to find the RMS value of a waveform. Using two multipliers and a pair of op amps, an RMS detector can be constructed. The first multiplier is used to square the input waveform. Since the output of the multiplier is a current, an op amp is customarily used to convert this output to a voltage. The same op amp may also be used to perform the averaging function by placing a capacitor in the feedback path. The second op amp is used with a multiplier as the feedback element to produce the square root configuration.

This method eliminates the thermal-response time that is prevalent in most RMS measuring circuits.

The input-voltage range for this circuit is from 2 to 10 Vpk. For other ranges, input scaling can be used. Since the input is dc coupled, the output voltage includes the dc components of the input waveform.



Nuts & Bolts

Peter Hiscocks

In preparing to perform the extensive series of tests on cassette wow and flutter that appear elsewhere in this issue, we were faced with a problem: we did not have a readily available wow and flutter meter. We were quoted something like $300 a month to rent such an instrument, so we decided to build one ourselves. Fortunately, a wow and flutter meter can be simple and inexpensive. The schematic of our unit is shown below.

The heart of the meter is a Signetics phase locked loop (PLL) Chip, the NE565. A phase locked loop is basically a negative feedback system that compares the reference frequency (in this case a 3 kHz tone played back from the tape recorder) to the frequency generated by its own internal voltage controlled oscillator (VCO). If the two are reasonably close together in frequency, the action of the PLL is to synchronize the two frequencies.

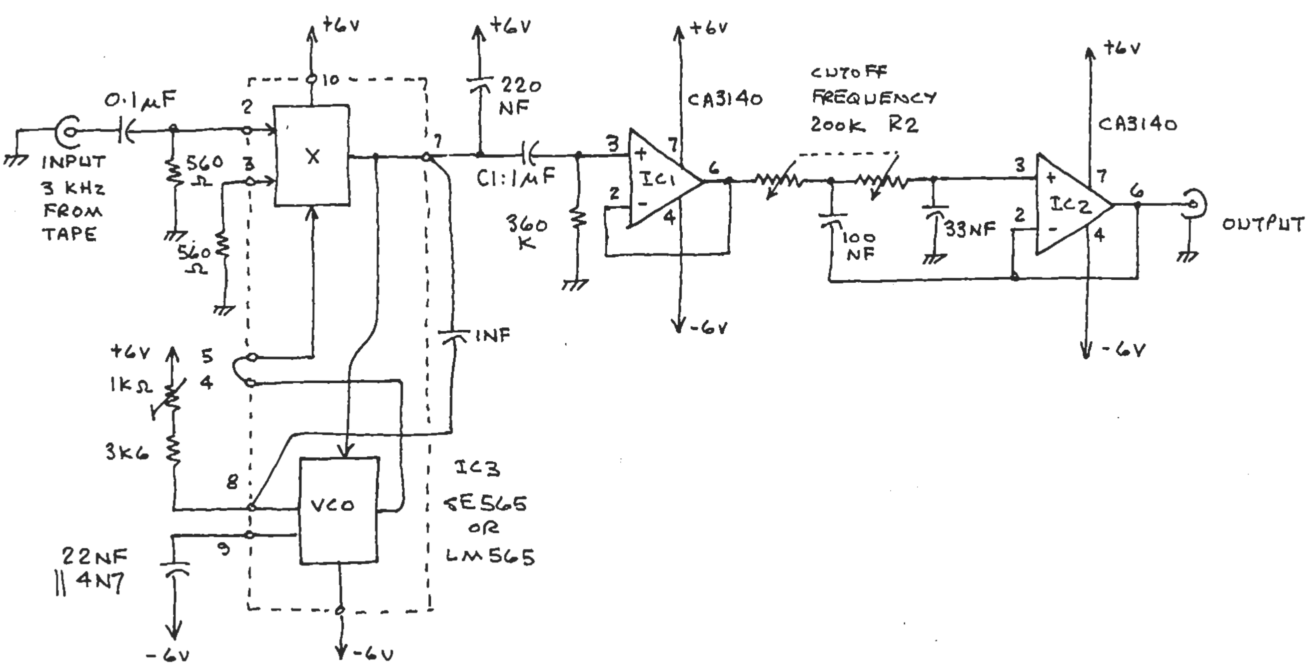
Synchronism is maintained by an "error signal" inside the phase locked loop. As the 3 kHz tone from the tape varies in frequency, due to variations in the speed of the tape recorder, the error signal will vary to cause the internally generated 3 kHz tone to track. The error signal may then be used as an indicator of frequency variation in the tape recorder speed, in other words, the wow and flutter component.

The phase locked loop is largely immune to variations in the amplitude of the tone from the tape, and it's capable of acquiring lock even when the incoming tone and the internal VCO are quite different in frequency.

To calibrate the wow and flutter meter, a 3 kHz tone is fed directly into the meter. This is varied by some known amount (a digital frequency counter is handy here) and the corresponding change in error voltage measured.

In the circuit shown, the blocking capacitor C1 removes a large DC component from the error voltage signal, passing signals above about 0.5 Hz. The CA3140 op amp (IC1) buffers this network and supplies the outside world with the error signal at a low impedance level. As it stands, the error signal, due to wow and flutter signals, is quite small: in the region of 5 millivolts for a typical tape deck. IC1 could be wired as an amplifier if the monitoring device needs a larger signal.

The error signal also contains a certain amount of 6 kHz frequency; the op amp low pass filter circuit (IC2) removes this residual signal. The wow and flutter waveform is monitored on an oscilloscope, the cutoff frequency of the low pass filter being reduced by adjustment R2 until the 6 kHz is seen to disappear. ■



AUDIO POWER METER

An accurate way to determine whats watt in your hi-fi system, with our true reading power meter

POWER IS PROBABLY the least understood and most misrepresented quantity in the electric measurement system. This is especially so in the area of audio amplifier and speaker specifications when terms like peak, peak to peak, music and RMS are related to power.

Power is simply the rate at which energy is being used. It is expressed in watts and the value may vary from femtowatts (10^{-12} W), as in the input power of a FET, to thousands of megawatts in the power generation field. The term thousand megawatts is generally used in preference to the more correct term, gigawatts.

Power can be calculated simply by multiplying voltage and current:

$$P = EI$$

In a DC circuit where both voltage and current remain constant no problem arises. However in an AC or a DC circuit where the voltage is not constant with time, this formula only holds for instantaneous power as the power varies with time. Power as we usually use the term is the time average of this. If the load is resistive, i.e. contains no inductance or capacitance, and we can measure the RMS value of the voltage, we can still use this simple formula. However measuring the RMS voltage is not easy as most voltmeters measure the peak or average rectified voltage with a suitable scaling factor built in to give a correct result when measuring a sine wave signal.

Reactive Reaction

If the load is reactive the current and voltage will no longer be in phase, i.e. the peaks do not occur at the same point in time. The difference can be expressed either by the phase angle in degrees or by the cosine of this angle (known as the power factor). The current waveform can either be ahead of the voltage (leading) or behind it (lagging). Capacitive circuits give rise to a leading power factor while inductive circuits lag.

If working with a sine wave, and if the power factor is known, the formula for power can be expressed as:

$$P = EI \cos \phi$$

where ϕ is the phase angle. In a DC circuit $\cos \phi$ is unity so the formula holds for this case as well. An example is a 40 W fluorescent light which takes 430 mA from the 240 V mains. At first sight, this implies a ▶

power consumption of over 100 W, until it is realised that its powec factor is about 0.45 lagging. The formula above, using cos ϕ=0.45, thus gives a power consumption of only 46.4 W. (The additional 6 odd watts is dissipated in the ballast.) The product of voltage and current is known as the VA rating and is used when calculating the currents in a circuit. If a capacitor is connected across a sine wave AC circuit the current taken can be calculated by dividing the voltage by the reactance of the capaitor. While this circuit draws current, it has a power factor of very near zero (90° phase lead) and therefore takes no power! By adding the correct amount of capacitance to an inductive circuit (i.e. the fluorescent light) the power factor can be altered, reducing the current drawn (but not the power).

Confused yet?

Ample Reason

Getting back to audio amplifiers and their ratings, the problem lies in the complex nature of the music waveform and how to specify the amplifier's rating. As the waveform is far from a constant sine wave with the peak power being anything up to 20 times the avêrage, numerous methods such as peak power, peak to peak power, music power, etc. evolved. However, for a long time there was no set standard, and one amplifier advertised with a 50 W (music) rating was in fact a 5 W stereo amplifier. The situation got so out of hand that the US Government brought down legislation on how amplifiers were to be tested. This is with a continuous sine wave signal with level set so that the distortion is at a specified level and power calculated from the RMS output voltage: hence the term RMS power. Note however that the term RMS refers to the method of measurement, i.e. the use of RMS voltage, and it is not the RMS value of the power waveform. It is, in fact, the average of the power waveform.

Speakers are just as confusing. They are normally specified not in terms of the power they can dissipate, but the maximum power of amplifier they are suitable for. This is due to the fact that music is never (well, rarely) a continuous sine wave and the average power in the speaker may be only 10% of the RMS rating of the amplifier, even with the amplifier clipping.

To measure the power actually being delivered to the speaker under music conditions, a wattmeter must be used.

Design Features

To multiply current and voltage together we had the choice of analogue or digital techniques. Unfortunately while digital is the 'in' thing, offering versatility and accuracy, it is not fast enough to calculate the instantaneous power on high frequencies. We therefore chose the analogue method.

Looking around the ICs, the only ones with reasonable price and availability were the MC1494, 1495 and 1496. The 1496 (or 796) is the cheapest and most readily available, but has the disadvantage of not being able to multiply DC signals or AC signals with a DC offset. The 1494 and 1495 are about the same price, and of the two, the 1494 was more linear and easier to use.

We chose not to use any input buffer on the voltage input but had to pay the penalty of having a lower input impedance than normal with voltmeters.

Using the Power Meter

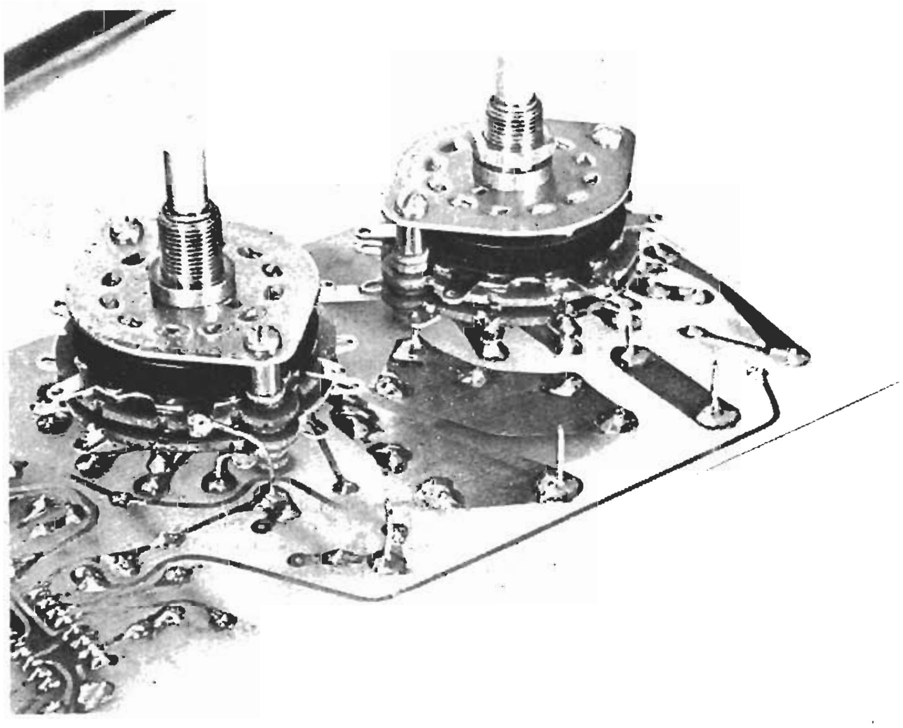
To use the meter we must measure both voltage and current. There must be a common point for these measurements. The current connection can be in either of two ways as shown in the drawings below. One measures the power out of the supply and the second the power into the load. The difference? The current shunt in the wattmeter drops one volt when working at the full range value and this may or may not affect the reading. At 10 A this accounts for 10 W which, if the power being measured is only 100 W, is a 10% error — although if the measured power is 2400 W the error is only 0.4%.

The range of the meter is the product of the individual ranges, i.e. on 30V and 1 A the fsd is 30 W, while 30 V and 3 A gives 100 W FSD. To help give a reading reasonably high on the scale, the voltage range can be overvoltaged by a factor of 2. Due to power dissipation problems this should not be attempted on the current ranges The peak voltage or current can be as high as three times the range value.

Construction

We mounted all the components associated with the meter and the switches on a single pc board and if the same or similar case is to be used this is recommended.

Except for the meter and the switches the components are mounted on the 'normal' side of the pc board. These should be mounted ▶



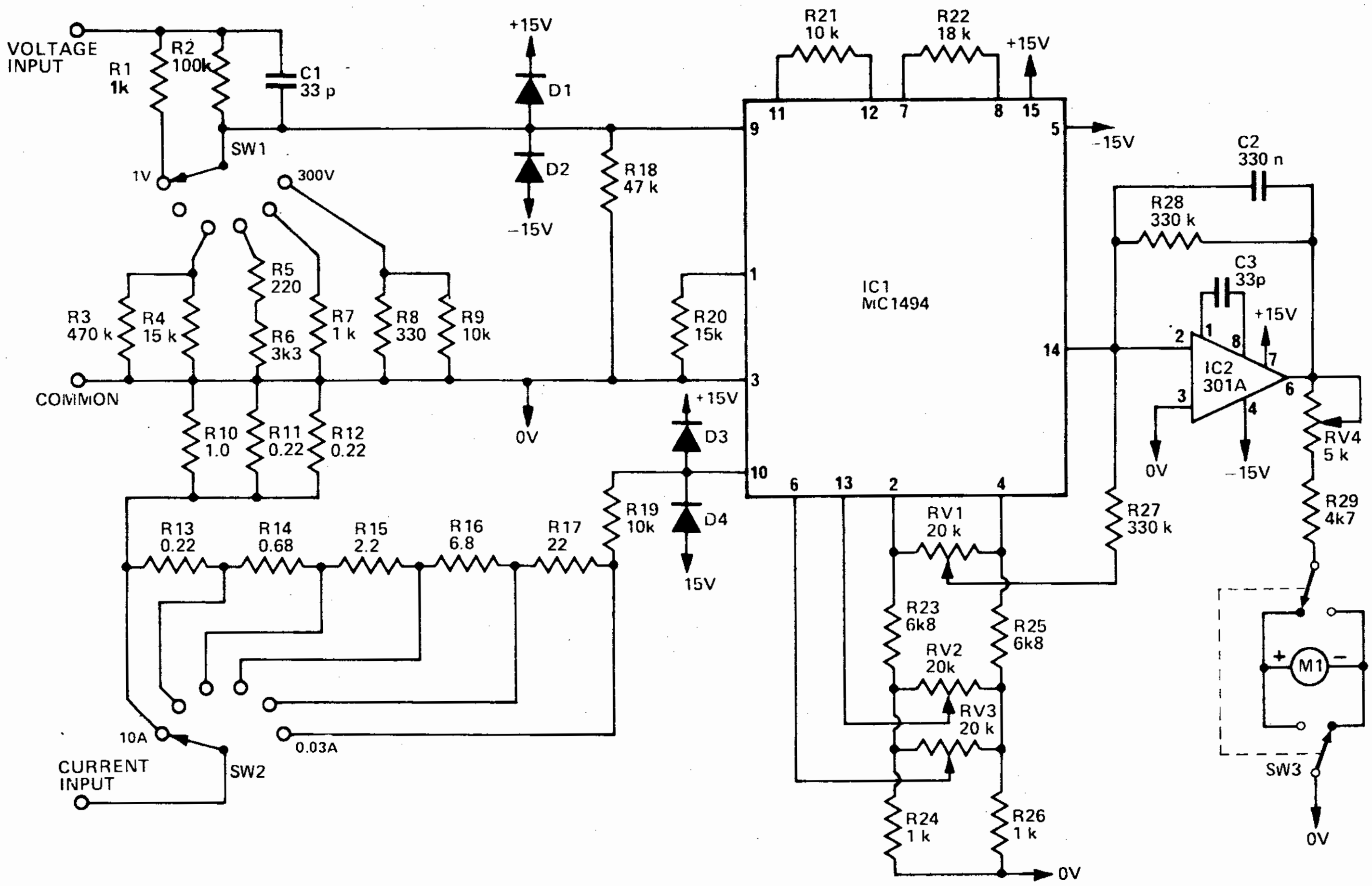


Fig. 1. The circuit diagram of the audio power meter.

HOW IT WORKS

Power is the product of current and voltage. This holds irrespective of the nature of the load, provided you are talking about instantaneous power. By multiplying current and voltage together and then taking the average of these instantaneous values we find the true power. Again this works irrespective of the load.

In this circuit the multiplying is done by IC1 (MC1494), the output of which is a current proportional to the product of the inputs. For more detailed notes on this IC, see the separate section. The current output of this IC is converted to a voltage by IC2 with C2 providing the averaging. The meter is then simply wired across the output of this IC with a meter reversing switch provided. This reversing switch is needed not to measure negative power, but to correct for reversed readings due to differing external connections.

The power supply is a full wave bridge with a centre tap giving about ∓20V DC which is then regulated to the ∓15 V required by IC1.

Adjustments for zeroing the voltage and current inputs are provided by RV2 and RV3 while RV1 compensates for offsets in the output. These are supplied by a stable ∓4 V reference in IC1. Range switching is done by SW1 and SW2. Protection against overvoltaging the IC is provided by D1 — D4.

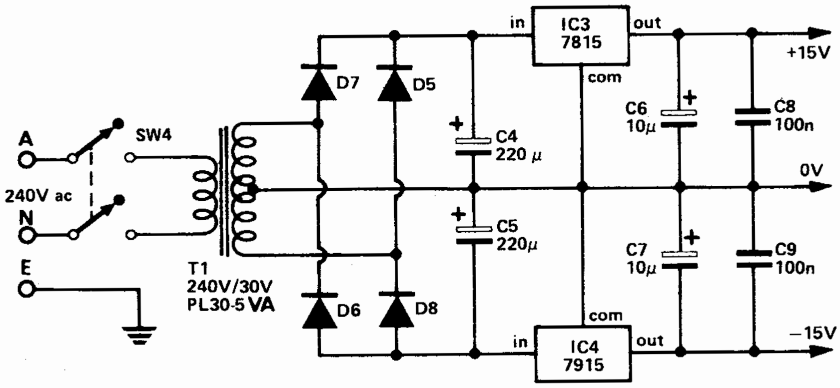
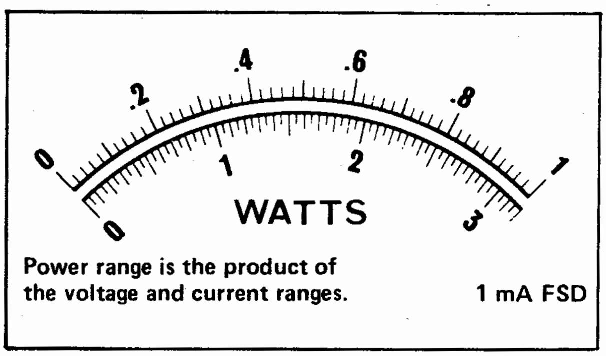


Fig. 2. Power Supply Circuit.



Right: meter scale designed for a 1mA FSD meter. These scales may need to be altered for differing meter units.

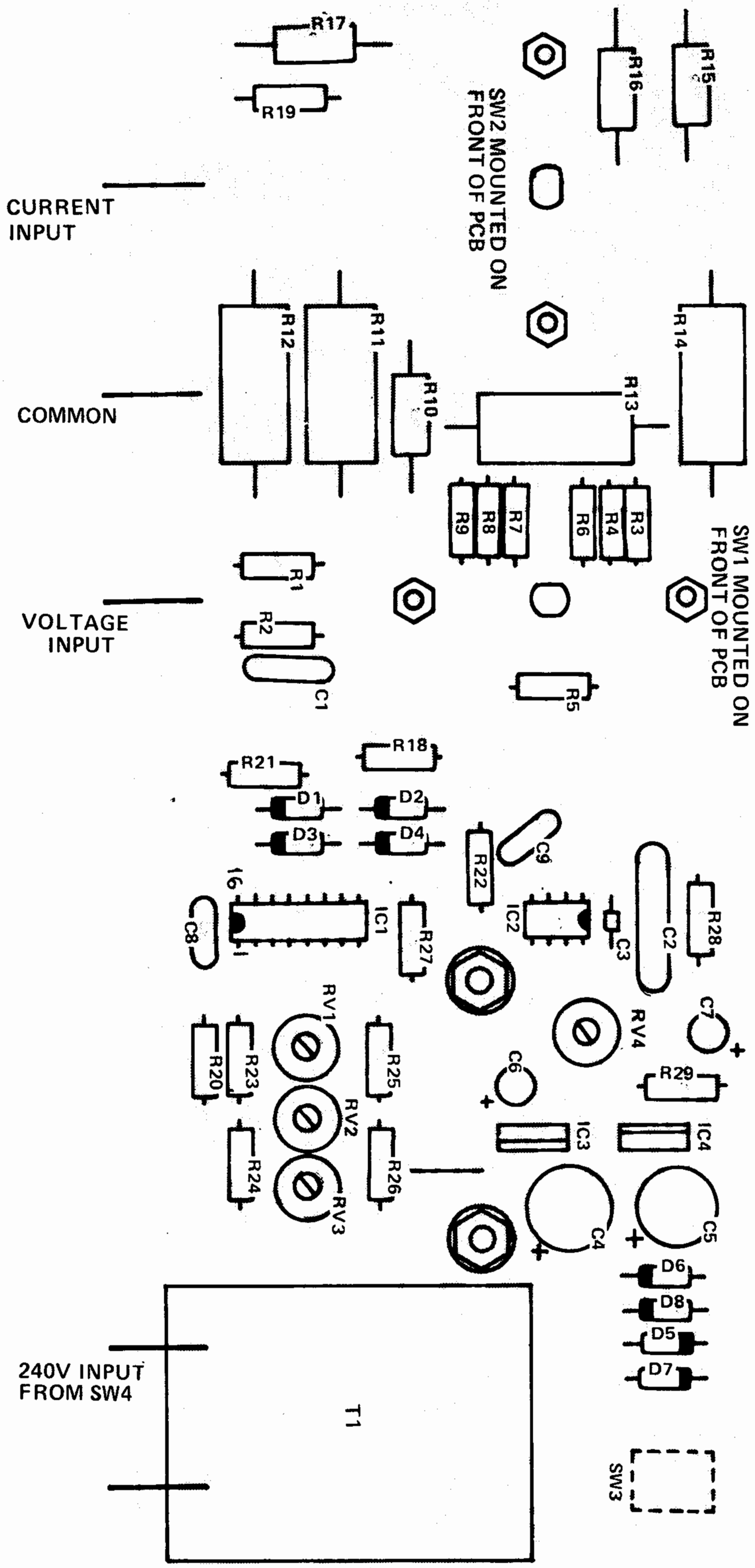


Fig. 3. Overlay for the Power Meter.

PARTS LIST

Resistors all 5% ½W unless stated

R1, 7, 24, 26	1k
R2	100k
R3	470k
R4, 20	15k
R5	220R
R6	3k3
R8	330R
R9, 19, 21	10k
R10	1R 1W
R11-13	0R22 5W
R14	0R68 5W
R15	2R2 1W
R16	6R8 1W
R17	22R
R18	47k
R22	18k
R23, 25	6k8
R27, 28	330k
R29	4k7

POTENTIOMETERS

RV1-3	20k trimmer
RV4	5k trimmer

CAPACITORS

C1	33p 500V ceramic
C2	330n polyester
C3	33p ceramic
C4, 5	220u 35V electrolytic
C6, 7	10u 25V electrolytic
C8, 9	100n polyester

SEMICONDUCTORS

IC1	MC1494
IC2	301A
IC3	7815
IC4	7915
D1—D4	1N914
D5—D8	1N4004

MISCELLANEOUS

PCB
SW1, 2 two pole 6 position 10A rotary Radiospares
SW3, 4 two pole toggle switches
Transformer 15-0-15, 5VA
Meter 1mA FSD
Three binding posts
Instrument case 255 x 100 x 205mm
Power cord and clamp
Two knobs
Front panel

BUYLINES

Most of the parts for this project are readily available. Two things which may cause trouble are the switch assemblies and the quadrant multiplier itself.

The switch is an RS unit and as such can be obtained from any of their stockists. As for the IC, Tamtronik — who advertise on page 32 of this issue — can supply this and by the time you read this they will be able to sell you all the rest as well!

first with the only critical part of the assembly in the area of the range switches. Here the high powered resistors should be spaced at least 5mm from the PCB as they run hot at maximum current. Also the leads of all the reistors in this area should be cut off close to the pc board after soldering. This is to give adequate clearance to the rotary switches. We used two self tapping screws into the plastic of the transformer case to help fix it onto the board. We have made ▶

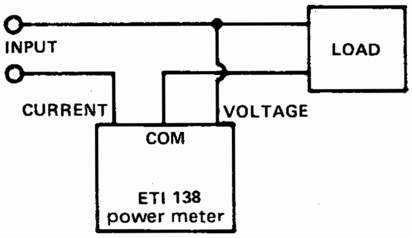


Fig. 4. This connection measures the power into the load.

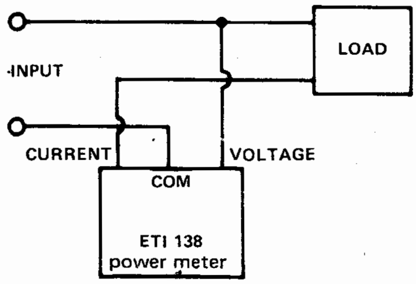


Fig. 5. This connection measures the power out of the supply.

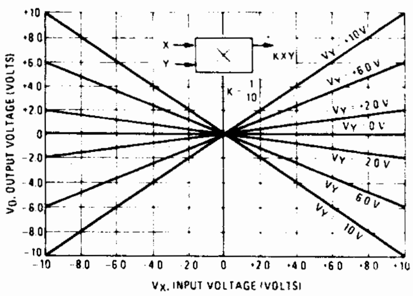


Fig. 6. Transfer characteristics of the IC.

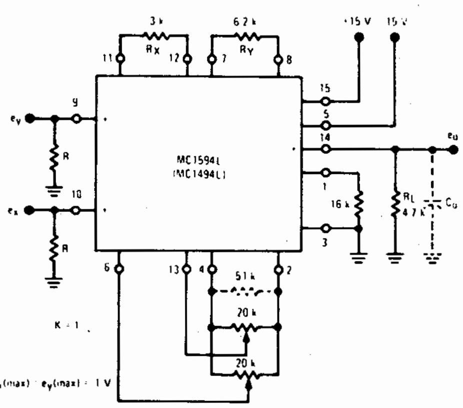


Fig. 7. Typical connections for a wide band multiplier or balanced modulator.

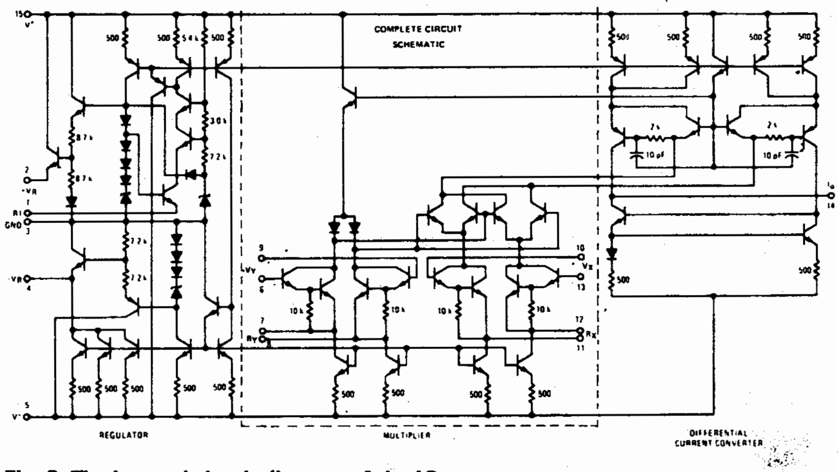


Fig. 8. The internal circuit diagram of the IC.

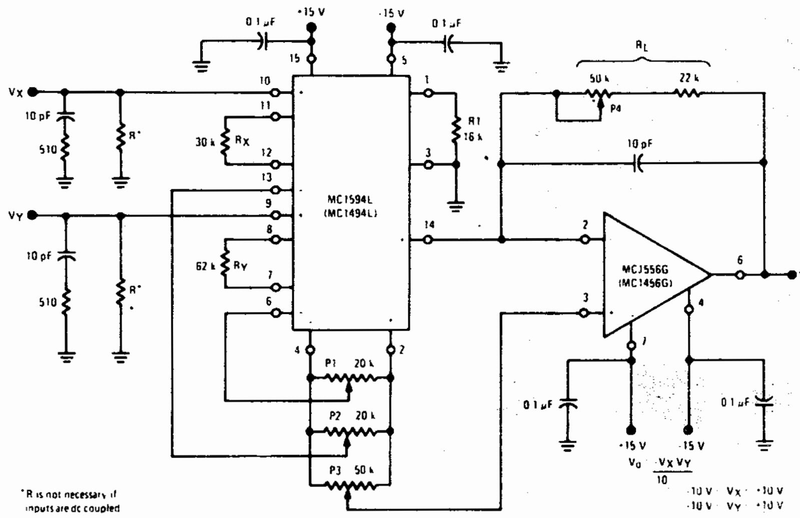


Fig. 9. Typical connection of a low frequency multiplier. For a squaring circuit simply parallel the two inputs. In this case pin 6 can be connected to 0V and P1 deleted.

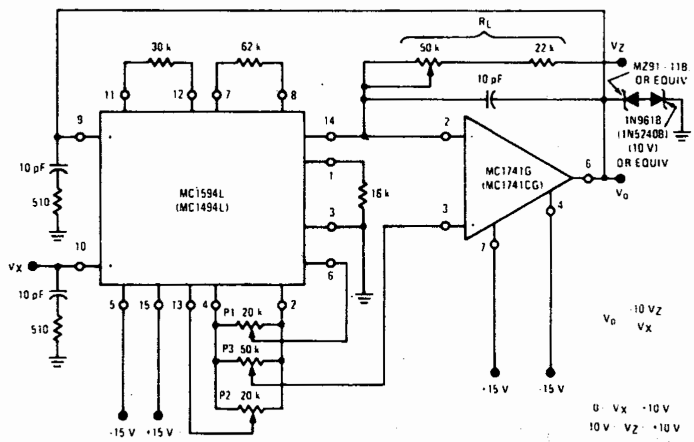
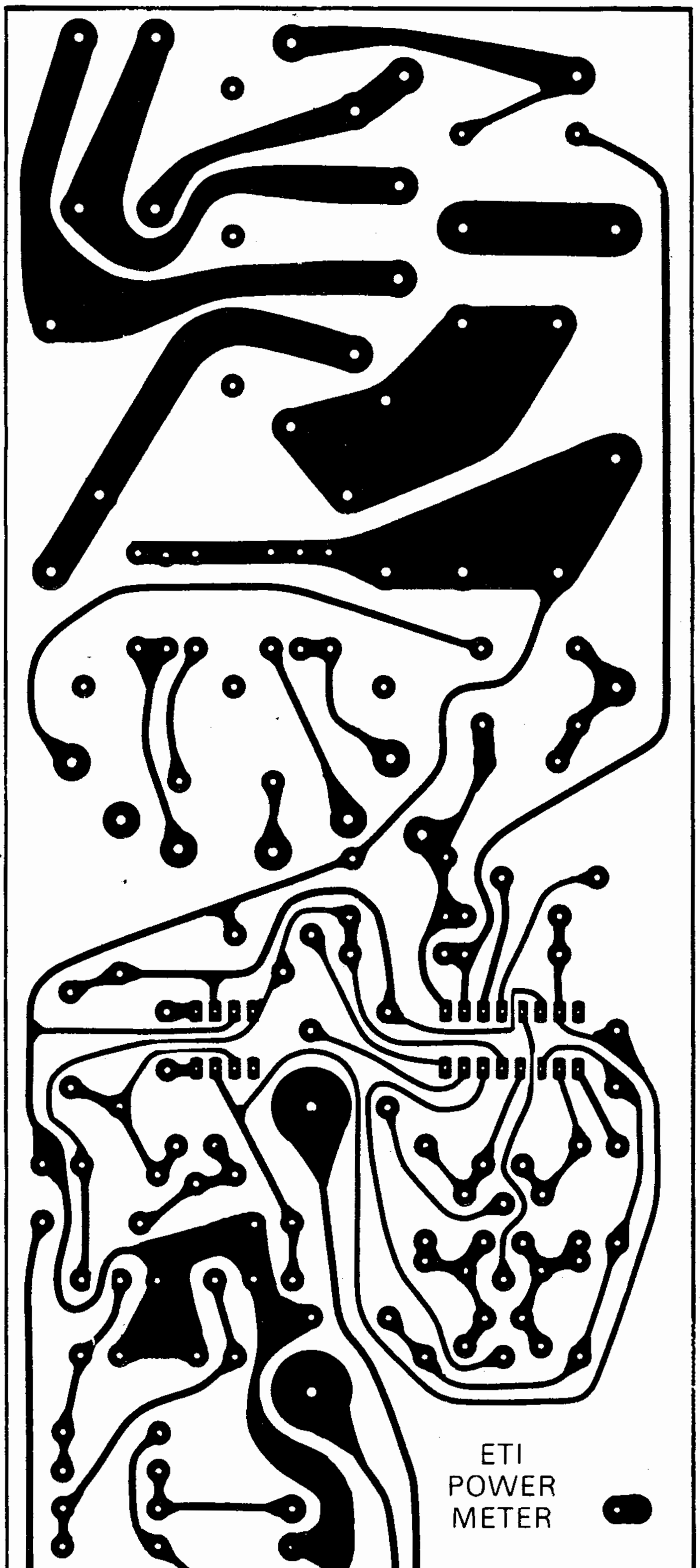


Fig. 10. Typical connection of a divide circuit. For the square root joins pin 9 and 10. Like the squaring circuits pin-6 can be connected to 0V and P1 deleted.

Full size foil pattern for the power meter.



allowance for either the cermet (VTP) or the normal carbon trim pontentiometer.

Calibration

Four adjustments are required, which are performed as follows:

Select the 1 V and 0.03 A ranges and switch on. If the meter reads in reverse, toggle SW3. Don't worry about the reading unless it is off scale. If it is, adjust RV1 to bring it back towards zero. Now apply a voltage of about 1V DC to the voltage input and note the meter deflection. Adjust RV2* until there is no deflection when this voltage is applied. Now apply the voltage to the current input (it will take about 30 mA) and adjust RV3 until there is no deflection. Recheck the voltage input and readjust if necessary.

Now with no voltage applied adjust RV1 to give zero output. Apply exactly 1 V to both current and voltage inputs and adjust RV4 to make the meter read FSD.

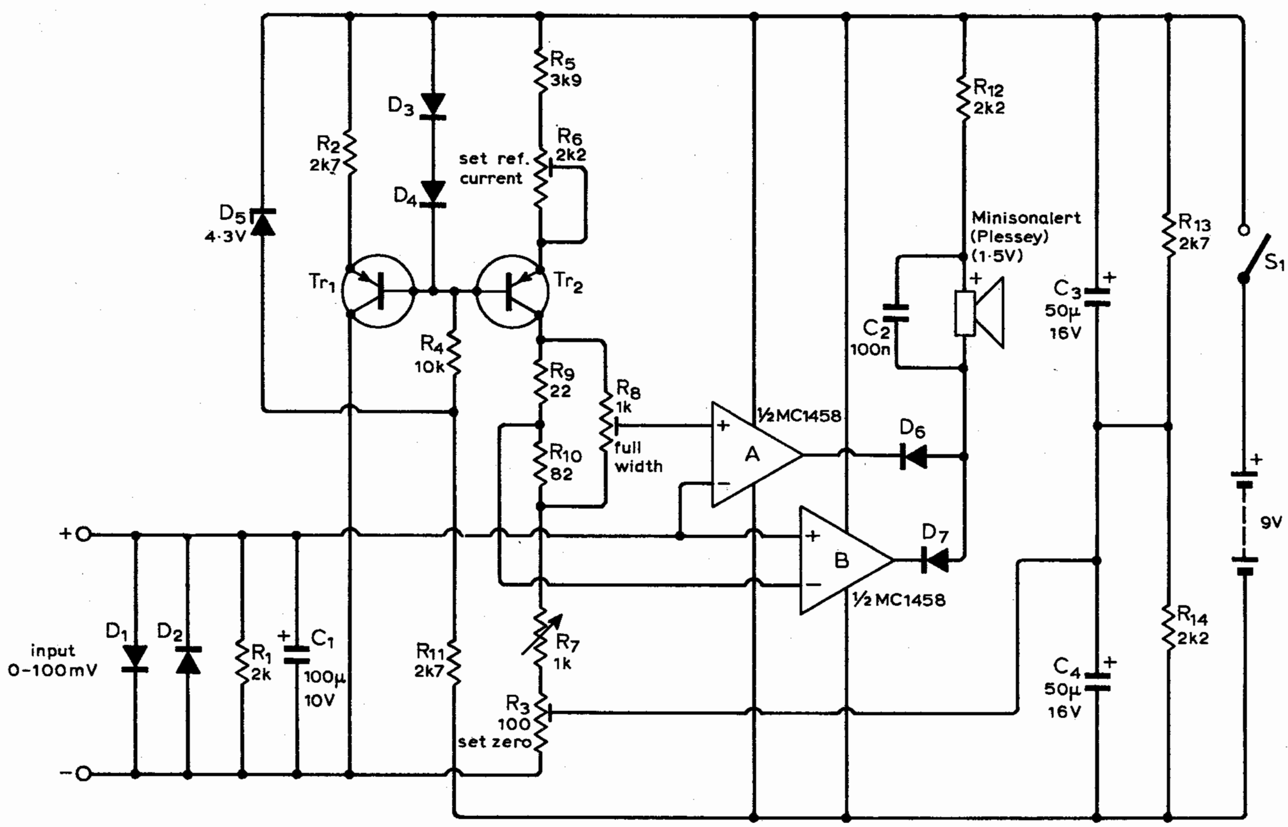
This is all the calibration that should be necessary.

About the 1494

The 1494 is a variable transconductance multiplier with a bidirectional current source output. What this means is that it looks at the voltage on the two points and gives an output current potential to the product of the two. Typical applications include: multiply, divide, square, square root, phase detection, frequency doubling, balanced modulation/demodulation and electronic gain control. An internal circuit diagram is given for those interested.

Values and Limitations

1 For best temperature coefficient R1 (pin 1 to 0V) should be 16k (we used 15k as it is easier to obtain). This sets the value of all the current sources inside the IC (I1 = 8/R1)
2 The value of Rx (pin 11 to pin 12) should be ⩾3x peak input voltage(X) expressed in k ohms.
3 The value of Ry (pin 7 to pin 8) should be ⩾6x peak input voltage(Y) expressed in k ohms.
4 Choose the scaling factory required ie Vout = K.Vx.Vy.
5 Load resistance (pin 14 to 0V) can be calculated by RL = (K.Rx.Ry.)I1)/2
6 If RL is connected between pin 14 and 0V without an inverting amp. the frequency response is limited by the output capacitance of 10pF.
7 For best temperature coefficient the load between pins 2 and 4 should be 8.6k. ETI



Circuit of multimeter for blind students (G. P. Roberts). Tr_1 and Tr_2 are silicon p-n-p types, e.g. BC177, BC187. D_5 is a 400mW, 4.3V zener diode, e.g. BZX79/C4V3, and other diodes are small signal silicon types, e.g. BA100, IN914A.

diodes. D_6 or D_7 is able to sink the 3mA required to operate the "minisonalert" which produces an audible signal at 3500Hz. However, the network consisting of R_8, R_9 and R_{10} is arranged so that for values of reference voltage very nearly equal to the input voltage, the outputs of both amplifiers go positive, producing an audible null. The width of the null—a compromise between accuracy and ease of setting of R_7—is adjusted by means of R_8. In practice, the high open loop gain of the Motorola 1458 dual op-amp ensures that the "edges" of the null are quite sharply defined, allowing the null width to be made as small as 0.2mV.

Where an attenuator raises the source impedance as seen by the input, the capacitor C_1 allows the instrument to be used to measure d.c. quantities containing moderate amounts of a.c. ripple. Although slowing the response time, this facility is useful when poorly regulated mains supplies are involved. The zener diode D_5 is included to further stabilize the current sources Tr_1 and Tr_2 against changes in battery voltage.

The instruments were built in diecast aluminium boxes measuring $8\frac{1}{2}$in × $5\frac{1}{2}$in × 2in. A simple in-built attenuator provides push-button selection of three voltage ranges (1V, 10V and 100V) and three current ranges (10mA, 100mA and 1A). A standard linear wirewound potentiometer was used for R_7 and this, in conjunction with a large pointer and embossed scale of 2.2in radius, was found to be easily read to accuracies of within 2% of full scale. Front panels were made from plastic laminate board. Braille dots were made by pushing ordinary dress-making pins through tight fitting holes drilled in the board, and cutting off their stems flush with the other side. The panel was also engraved for the benefit of sighted teachers.

The materials and metalwork were provided by the School of Mathematics and Physics, Macquarie University, with help particularly from Mr Ingram Paterson.

G. P. Roberts,
Cheltenham,
N.S.W., Australia

An audible voltmeter and bridge-indicator

"Bellbird" — an aid for the blind

by R. A. Hoare, B.Sc. *Manurewa High School, New Zealand*

Most of us will realise how much we owe to our eyes in the pursuit of our electronics profession or hobby. We may feel that blindness would completely end our participation in such activities, but this is wrong, as many blind radio amateurs have proved. However, there are many difficulties and a major one is the making of measurements. Various methods have been devised to enable the blind to read moving-pointer instruments, and most of these use photocells and buzzers. There are also null-type instruments with large dials labelled in Braille, but these are slow and inconvenient. Modern digital methods seem to offer the answer.

Clive is a seventh-form physics student who has been blind from birth. His hobby is electronics and he builds all sorts of things, relying on written descriptions of the circuits. Sometimes he is helped by having an integrated circuit mounted on a larger printed circuit board as a sub-mount, but apart from that he is self-reliant. It was with Clive in mind that a rather old digital voltmeter (using r.t. logic) was bought with the idea that it could be used to give an audible output. The meter could be used not only in his hobby activities but also in his 7th form physics experiments and later in his university studies.

The voltmeter is conventional in that it has three digits with an over-range 1 and an automatic polarity indication. Overloading results in blanking and an X display on a Nixie tube. It seemed that the problem was that of converting the parallel display, where all the digits are seen at once, to a serial presentation, where only one figure is seen at a time. I believe that a device which announces the digits orally is available, but this is expensive and difficult for the lone worker to make. The information inside the voltmeter is, of course, in binary form and there seemed to be no reason why this should not be suitable for direct communication with the user, if translated into suitable sound and presented bit by bit. There are several possibilities: changing note length, note pitch, or note quality. The first method is used in Morse code but would possibly be confusing in this new application, unless the five-bit Morse numerals themselves were used. These are rather cumbersome and would present difficulties in decoding. The translation of a binary 1 into a high pitched note and a 0 into a low pitched note seems natural and has proved acceptable in practice. If the X, +, — and over-range 1 can be combined into one digit then we have four 4-bit digits to convey, in addition to the decimal point, which suits the capability of a 16 to 1 multiplexer very well. This, as many will know, is an i.c. with an output which can take up the state of any one of sixteen inputs, as selected by the binary number on four address pins. It was found possible to invent a simple code for the prefix digit, thus:

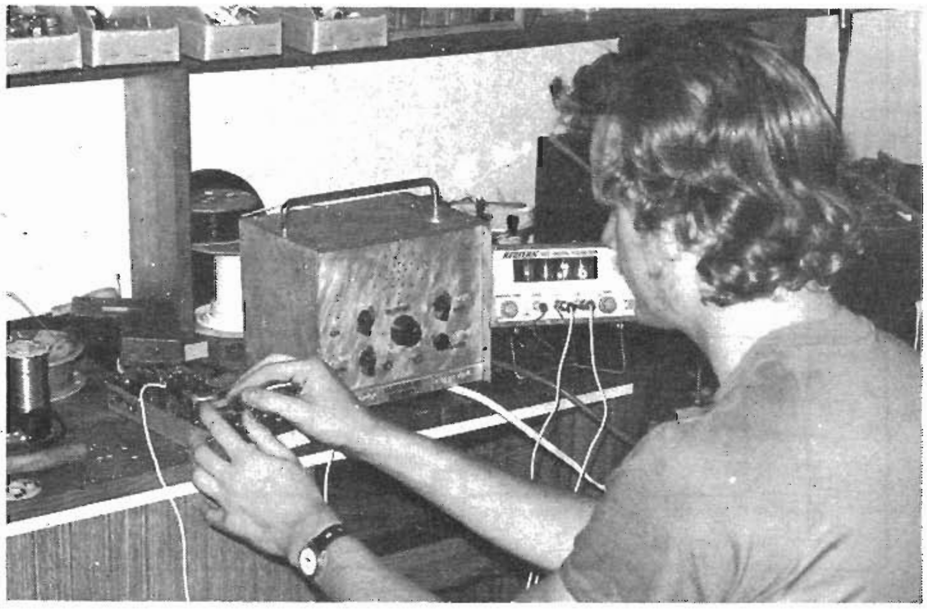
+0	1	1	0	0
−0	1	0	1	0
+1	1	1	0	1
−1	1	0	1	1
X	1	1	1	1

All of these are binary numbers greater than 9, so there is no possibility of confusion with other digits.

The multiplexer is made to select each of its 16 inputs in turn by means of a binary counter connected to its address pins. The counter is operated by a multivibrator (two monostables) working continuously.

We now come to a point where two distinct design approaches are possible. There must be pauses between the digits, a long pause at the end of each reading, and extra pauses to enable a brief "decimal point" pulse to be inserted at the correct point between digits. These delays can be provided either by monostables, which switch off the counter for a period, or they can be arranged to span a given number of counter pulses. The latter method uses a fully digital system. There are advantages and disadvantages to each system, but I was attracted by the simplicity and flexibility of the monostable method because it was difficult to know in advance the exact time intervals required, and monostables offer simple and almost infinite adjustment.

The tone frequencies are provided by an LM566 voltage-controlled oscillator, the output from which is amplified by



Measuring instrument being used by Clive, an electronics hobbyist who has been blind from birth.

an LM380 2½ watt amplifier. The frequencies generated by this device depend upon the input direct voltage and the value of a capacitor connected to the i.c. It seemed to be wasteful merely to use this versatile component to give two pitches, and a circuit was devised so that the instrument could also be used for an entirely different purpose, as a bridge null-detector. In the latter application the suitably amplified out-of-balance voltage from almost any bridge circuit is used to alter the frequency. (A bridge rectifier circuit gives a rise in pitch for both positive and negative input voltages.) In the digital application a fixed voltage is switched in and the multiplexer is used to change the capacitor values.

Clive had no difficulty in learning the code. A speed control was fitted, and it was not long before new timing capacitors had to be provided to allow him to work faster. The rather strange warbling note gave rise to the name "Bellbird," bearing some resemblance to the song of that New Zealand bird.

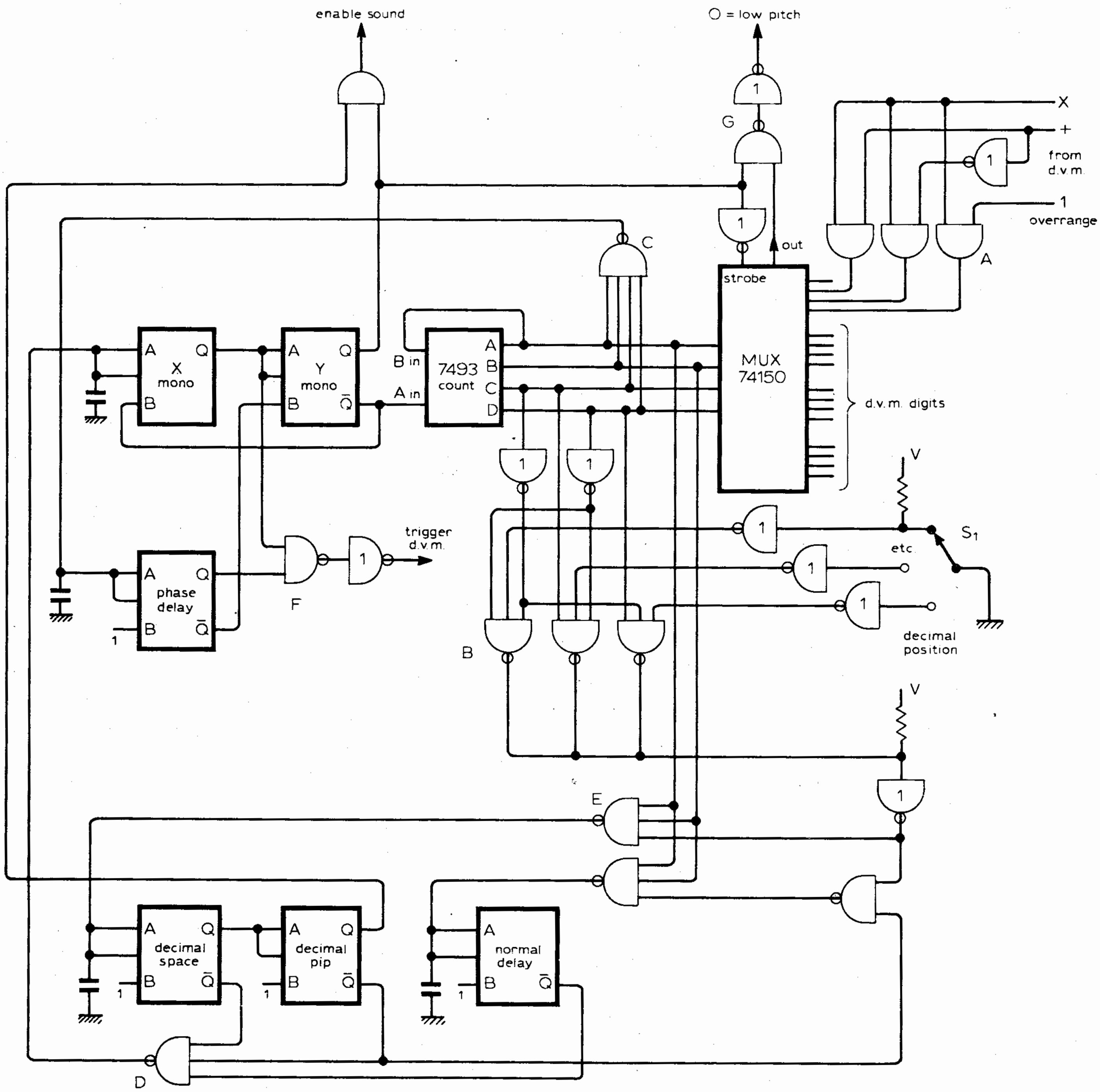
I consider that sighted workers may have a use for such a machine, as it enables one to concentrate on the circuit under test instead of dividing attention between that and a meter reading.

Circuit description

For clarity, the digital circuit diagram (Fig. 1) omits power supplies, irrelevant connections, and timing resistors and capacitors. The various delay i.c.s are all type 74121, though dual devices could be used. Doubtless improvements could be made, not least of which would be the avoidance of "glitches" caused by race conditions, which are the reason for small capacitors on the A inputs.

Monostables X and Y, operating as a multivibrator, cause the binary counter 7493 to select each of the voltmeter digits in turn to be presented at gate G. The OR gates A provide the prefix code already discussed. The action of the multivibrator is interrupted by gate D,

Fig. 1. Digital circuits for translating a measured voltage, represented by the digits within the digital voltmeter, into a series binary code suitable for triggering a circuit (see Fig. 2) to produce audible low/high pitches. Spaces between numbers and readings are also produced by these circuits, together with a decimal point when required.



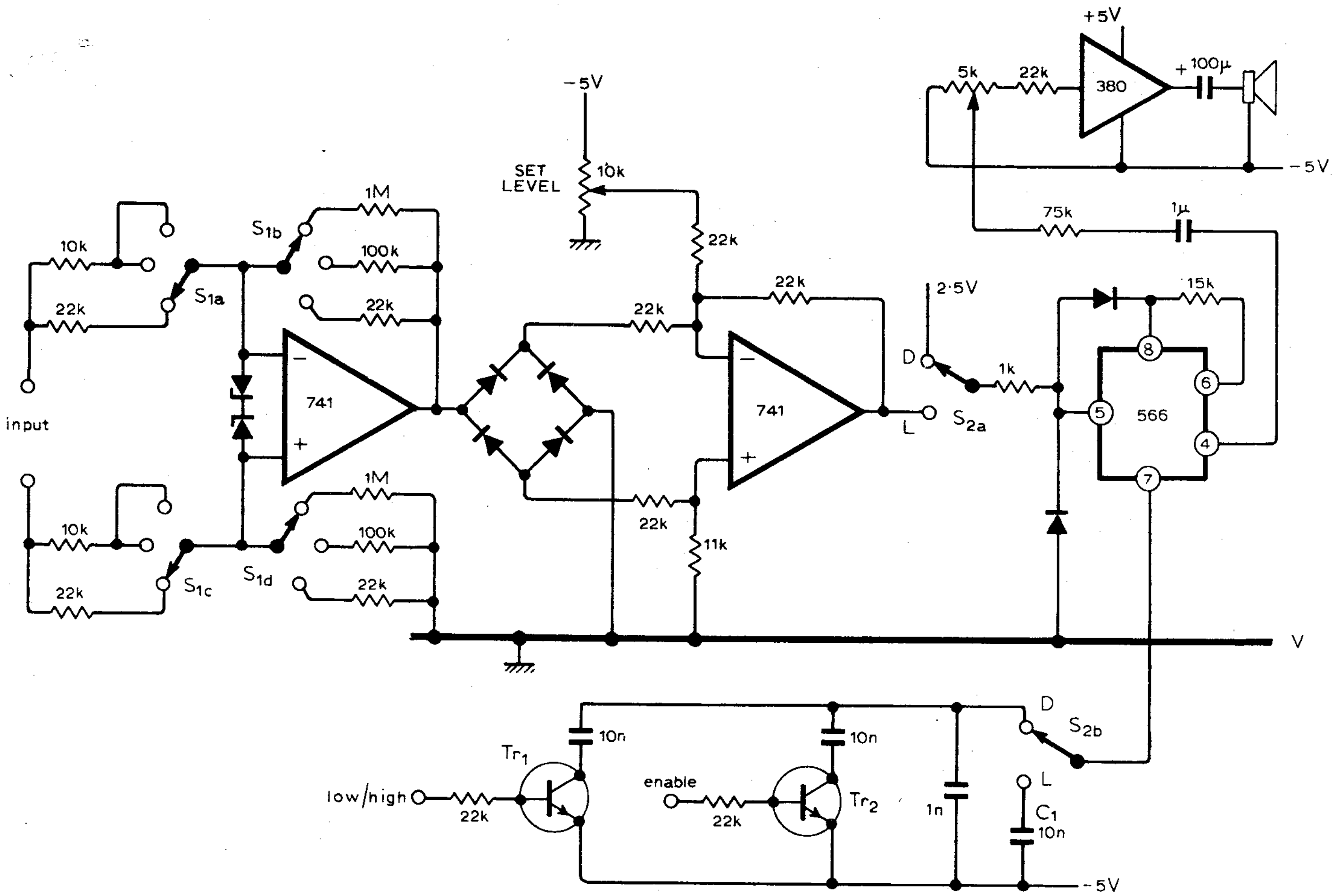


Fig. 2. Bridge-indicator and sound circuit. Switch positions D enable the circuit to be used in the digital voltmeter mode, the v.c.o. LM566 producing a low pitch for an input binary "0" and a high pitch for an input binary "1". Switch positions L enable the circuit to be used as a bridge null-detector, the null point being at the point of lowest pitch.

which operates whenever the "decimal space" "decimal pip" or "normal delay" monostables fire. Gates B control these in the following way: with S_1, the decimal point switch inside the voltmeter, in the position shown, the wired-OR output from gates B goes to logical 0 on a count of C = 0, D = 0. In the mid position the output goes to 0 on C=1, D=0 and in the bottom position on C=0, D=1. The A and B lines of the counter output are connected, as well as the inverted B gates output, to gates E. It will be seen that the upper gate fires the decimal-point delay on a count of 0011, 0111 and 1011 in the three switch positions mentioned. This will put the decimal delay after the prefix, first or second digits respectively. When the decimal delay monostable finishes its pulse it triggers the decimal-pip monostable, which takes over the job of arresting X and Y and sends a signal to the sound enable gate.

The lower gate E is operated, giving normal delay, whenever the A and B lines are high and the B gates output is also high, provided that the decimal-pip monostable has not fired. This means that it is triggered at the end of digits when the decimal point signal is not given. At the end of the whole cycle or phrase a count of 1111 forces the output of gate C low, thus firing the phrase-delay monostable and operating gate F to send a trigger signal to the voltmeter, which luckily has provision for this external control of its cycle. It is possible that spurious readings would be obtained with some meters if the outputs were multiplexed while they were half way through their cycle. In the Bellbird system the digital voltmeter measures only between phrases.

It will be noticed that if there is no connection at S_1 no decimal point indication is given.

Bridge indicator and sound circuit

Referring to Fig. 2, S_{2a} connects the input of the voltage controlled oscillator to either the amplified out-of-balance voltage or to a fixed 2.5V potential, to give the two modes of operation. At the same time S_{2b} connects the oscillator to either C_1, for linear operation, or a circuit controlled by Tr_1 and Tr_2 for the voltmeter application. Safety diodes protect the LM566 from negative inputs, which could result from failure or wrong connections in the previous circuits. Two type 741 operational amplifiers perform the tasks of amplifying input signals with gains of 1, 10 and 100, selected by S_1, and adjusting the d.c. output level to give a suitable range of tone. The input to the first 741 is protected by two 5V zener diodes. A rectifier bridge ensures that when the output from the first 741 goes either positive or negative from zero the oscillator input voltage, and therefore the tone, will rise. This bridge should be constructed with germanium diodes to avoid a large "dead zone" caused by the higher forward voltage of silicon types.

The switching circuit enables one or two capacitors to be connected to the oscillator, giving high and low audible tones.

The audio circuit is as simple as possible: the a.c. component from LM566 is attenuated by a volume-control potentiometer before being fed to the LM380 amplifier. The loudspeaker coupling capacitor need not be of a high value since no low tones are required.

In use no special difficulties were noted. In linear operation the null point was obtained by listening for the point of lowest pitch. Greater sensitivity was obtained by adjusting S_1 and 1mm discrimination could easily be obtained on a metre bridge experiment. In the digital mode, as already noted, the learning process was fast. Sighted pupils were also interested, and as they had already some knowledge of binary code they were able to translate for themselves, though at a slower rate. After practice, of course, recognition of the "digit pattern" occurs, as with Morse code.

Digital meter for the blind

A three-decade instrument with audible balance indication

by T. C. R. S. Fowler, B.Sc.
University of Bristol

A combined recording level indicator and d.c. voltmeter built in 1972 for a blind man was the forerunner of the instrument to be described here. In the voltmeter mode the original instrument indicates the magnitude of the voltage by the frequency of an audible output note, and a 12-way switch enables any one of twelve reference voltages to be switched in for comparison. The reference voltages rise in constant-ratio ("3dB") increments from 50mV to 2.26V, an external attenuator being used at the input to provide higher voltage ranges as required.

At a later stage, the requirement arose for a wider range of reference voltages and it was decided to select the references by means of rocker switches in a binary-coded decimal array. Resistance measurement is now made possible, the complete instrument becoming a 3-digit volt-ohmmeter costing around £15.

Instead of giving a continuously-variable output frequency controlled directly by the input voltage, the digital instrument includes a comparator which causes the audible output frequency to change abruptly, generally from a steady high note to a steady low note, as soon as the reference voltage is made to exceed the voltage being measured; only if the two voltages are very nearly equal does the output frequency "dither" in the intermediate range. All three types of output sound are clearly identifiable, even by the non-musical!

The 12 rocker switches are arranged in three columns of four, with weightings of 4, 2, 2 and 1 in each column, most significant digit column on the left. In addition, a lever bar like the spacer bar on a typewriter actuates a microswitch and enables one further least significant digit to be added and removed repeatedly and easily. Thus the full-scale reading is "10.00" rather than "9.99", and the stabilized voltage is set to 10.000V when calibration is carried out.

To take a measurement, all rocker switches are initially set to "off" (left-hand side down); a high note output should result. The switches are then operated in order, starting with the most significant "4", each being left on if the note does not change, but switched off again if the note changes. Finally a stage is reached when operating the lever bar causes the note to change. The state of the rocker switches, representing the numerical value of the quantity being measured in 3-digit decimal form, is then read by touch by the operator.

Circuit description

Fig. 2 is the circuit diagram of the instrument with the reference voltage generator network shown in block form; the circuit of this network is given in Fig. 3, in which S_4 to S_{15} inclusive are the 12 rocker switches, and S_{16} the micro-switch actuated via the lever bar.

An equivalent circuit for a reference voltage generator of this type comprises a direct voltage generator variable in steps between the terminal voltages of the stabilized power supply used and of a constant output resistance R_0 equal to that of all the digital-to-analogue network resistors in parallel. Here the equivalent generator is variable in 10mV steps from 0 to +10 volts, and $R_0 = 0.4R$, where R is the resistor associated with the most significant "4", viz. R_1 in Fig. 3. (It may be of interest to note that where the extra least-significant-digit facility is provided, as by R_{13} and S_{16} here, the relationship $R_0 = 0.1nR$ holds for any number of decades, where R is the "most significant" (lowest) resistor of the network and n is the numerical value associated with it.) To keep power consumption reasonably low yet the values of R_0 and the highest resistors R_{12} and R_{13} not unduly high, a value of approximately 5000 ohms was chosen for R. In practice, a 4.7kΩ (nominal value) high-stability resistor, measured and found to be of 4720 ohms, was used in the prototype for R_1 and as the basis for all the other resistors in the reference voltage network, giving R_0 = 1888Ω and R_{12}, R_{13} = 1.888 MΩ. In the prototype, a "main" resistor close to the required value was used in combination with one or more "auxiliary" resistors, in parallel, in series, or in series-parallel configurations; for example, for R_2 and R_3 a high-stability resistor of nominal value 10kΩ was used as the "main" resistor; shunting it with two resistors, nominally 470kΩ and 10MΩ respectively, gave the required value in both cases. Where, as in this example, the parallel resistors were of relatively high value, carbon film types were considered adequate.

On the 10V and 100/1000V settings, switch S_{2A} connects the output of the reference voltage generator directly to the inverting input of the comparator IC_1 (a "741" operational amplifier). On the other ranges, resistive attenuators are used, the constant output resistance R_0 of the reference generator giving attenuation.

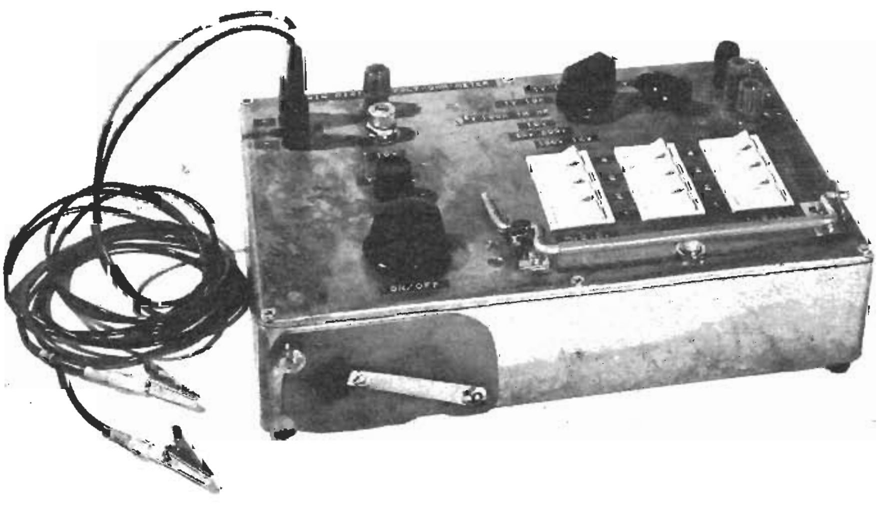


Fig. 1. The complete instrument. The multiplier lead may be substituted for one of the crocodile clip leads.

The voltage present between the input terminals is applied, virtually unattenuated on the 1V, 5V and 10V ranges, and attenuated by a factor of ten on the higher ranges, between the "0V" line and the non-inverting input of the comparator IC_1, switch S_{2B} being used to make the necessary connections. On the former group of ranges the 10kΩ resistor R_{18} and diodes D_1, D_2 and D_3 give overload protection, and R_{18} with C_2 provide smoothing; on the higher ranges R_{18} is replaced by the 1:10 attenuator network R_{17}, R_{19}, R_{20}, R_{21} and R_{22}, the last two resistors providing a small positive bias voltage which compensates for the effect due to the input current of IC_1 and the rather high output resistance (about 200kΩ) of the attenuator network.

The input resistance on the 50V and 100V ranges is very approximately that of R_{17} and R_{19} in series; viz. 2.22 MΩ. For the 500V and 1000V ranges the multiplier lead is used, adding approximately 20 megohms. Thus the instrument has a resistance of approximately 22,000 ohms per volt on the 100V and 1000V ranges, and 44,000 ohms per volt on the 50V and 500V ranges.

The output state of the comparator IC_1 is indicated by the audible output of the instrument, which is generated by Tr_1, Tr_2 and associated components forming an astable circuit, the emitter follower Tr_3, and the loudspeaker. Generally, of course, the output voltage of the comparator is "hard over" at either the upper or the lower limit, and the values shown for R_{23}, R_{24}, R_{35} and R_{36} were chosen to give a very distinct, though not extreme, frequency change when the comparator switches. The values of the capacitors C_4 and C_5 in the astable circuit were made unequal for power economy. The tonal quality of the output sound is considered not unpleasant, although the provision of a jack socket for headphone operation as an alternative to the loudspeaker might nevertheless be a worthwhile addition.

The connection of R_{23} to the "B+" line is used to give a rough audible indication of the positive battery voltage—if this is low, both frequencies from the astable circuit are appreciably lower than with the nominal 18 volts, though still remaining clearly distinguishable from each other. If the "B−" voltage is very low, the astable

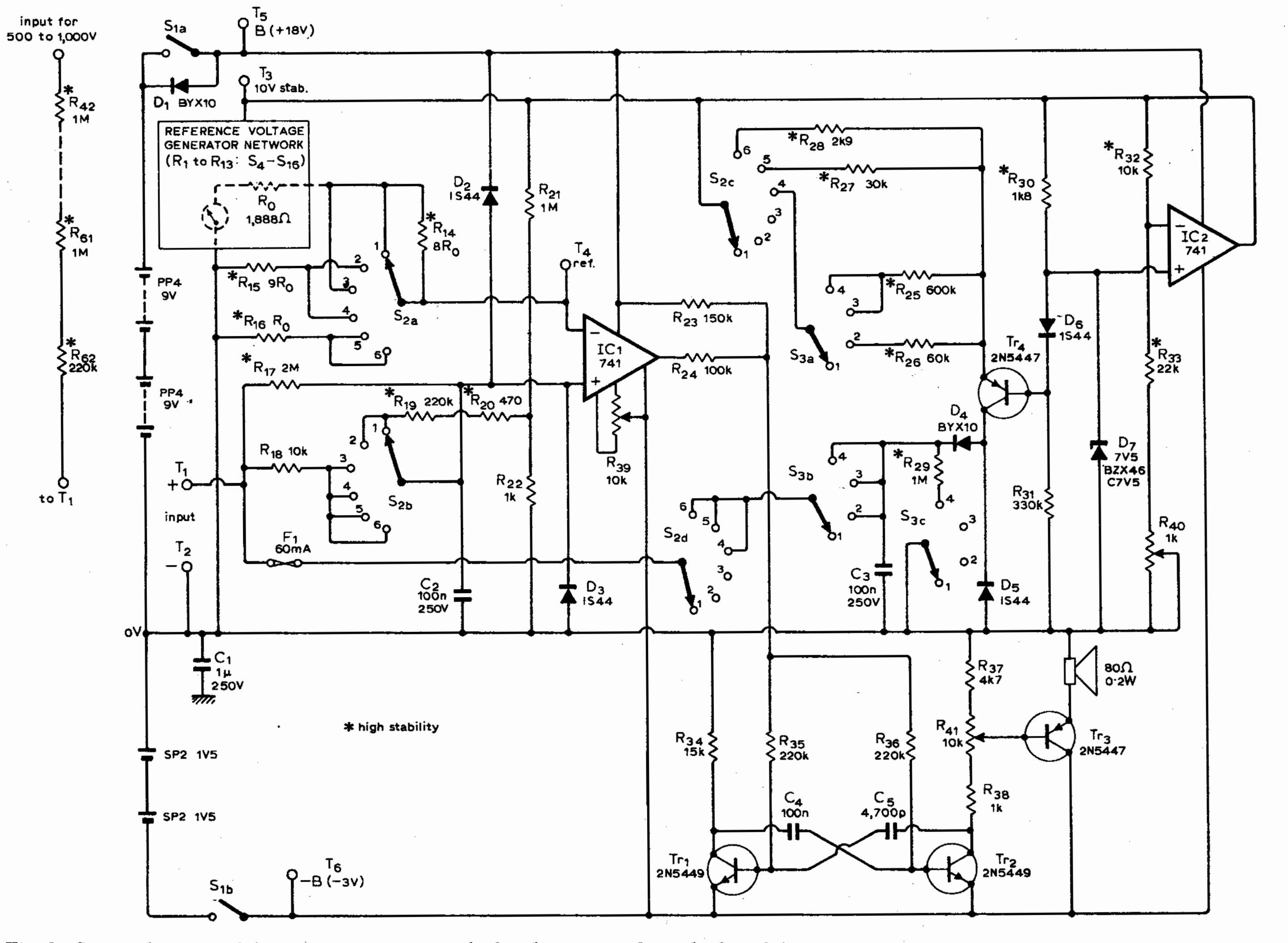


Fig. 2. Circuit diagram of the meter. Resistors marked with an asterisk are high-stability.

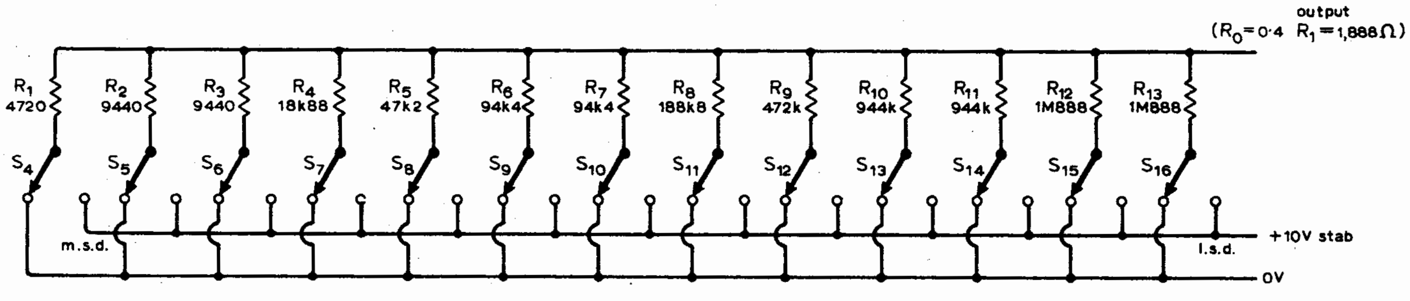


Fig. 3. Reference voltage generator network.

circuit will not operate. A further case of non-operation of the astable circuit should be mentioned here; it will not start if the B− voltage is switched on after the B+ voltage; this trouble can be obviated by the use, as in the prototype, of a two-position rotary wafer switch for S_1, the wafer contact being filed if necessary to ensure that the B− supply is always switched on first. No such filing proved to be necessary in the prototype, and non-starting has not been a problem.

The comparator offset adjustment potentiometer R_{39} is set so that on all the voltage ranges, with all the rocker switches set at "off", and the input leads short-circuited, the high output note results, but pressing the bar to add one least significant digit produces the low output note. (Once set, R_{39} should seldom need to be readjusted, so this trimmer is mounted on a circuit board rather than on the control panel.) The relevant reading is thus that obtained from the rocker switches when these have been set so that pressing the bar causes the output frequency to change; the extra least significant digit added by pressing the bar should not be included in the reading.

The necessary 10-volt stabilized supply is generated by means of a second 741 operational amplifier IC_2. Reference diode D_7 is nominally a 7.5V type, but in the interest of power economy it is run in this circuit at under 2mA and gives a reference of about 6.9V. The accuracy and stability of the "10-volt" line are nevertheless reasonably satisfactory, variations being typically less than ± 5 mV after an initial settling period. If a suitable digital voltmeter is available, the monitoring terminal T_3 and fine adjustment facility provided by R_{40} on the control panel make it easy for a sighted person to check and if necessary readjust the line to 10.000V.

The reference diode voltage also controls, via temperature compensator D_6, the base voltage of TR_4 which provides at its collector the selection of "constant" currents required for the various resistance ranges. These currents are 1mA, 100μA and 50μA, respectively, for the 1kΩ, 10kΩ, and 100kΩ ranges, and 5μA for both the 1MΩ and the "HR" range. Thus the full-scale voltage is 1V on the first two of these ranges (switch S_2 at positions 6 and 5 respectively, S_3 at "R"), and 5V on the other three (S_2 at position 4, S_3 at "R", "IM" and "HR" respectively).

Since the current from Tr_4 on each range is not completely independent of collector voltage, it is best to trim resistors R_{25} to R_{28} inclusive to give the correct current when the maximum resistance for the particular range is being measured, viz. 1MΩ, 100kΩ, 10kΩ and 1kΩ respectively, using a digital current meter between the resistor and the negative ("0-volt") input terminal. Once the correct current has been set up, the current meter should be removed from the circuit (or short-circuited) to eliminate the voltage drop introduced by it; accurate measurements of resistance should then be possible. The high-stability resistor R_{29} is switched in on the "HR" range; when R_{25} has been trimmed to give the requisite 5μA on the 1MΩ range, switching to the "HR" range should give a full-scale reading with no external resistor connected, i.e. with input terminals open-circuited. In the prototype it was preferred to set the current marginally on the low side, giving an open-circuit reading on the "HR" range of 0.995, so that the open-circuit reading would remain measurable even if there were a slight upward drift subsequently; using this method the quantity x, the reading obtained as a fraction of full-scale, referred to later, is strictly the reading divided by 0.995, but a good approximation for x when close to unity is obtained by subtracting the reading from 0.995, and subtracting the result from unity, e.g. a reading of 0.975, giving a difference of 0.02, would give $x = 0.98$. (If the open-circuit reading changes after a time, say to 0.997, this new value should of course be substituted for 0.995 in calculating x.) It might be worthwhile to add a front-panel trimmer for R_{25}, to enable the open-circuit reading to be set to full-scale each time the "HR" range was used, enabling x to be read directly.

Diodes D_4 and D_5 and the 60mA fuse are included to protect Tr_4 in the event of voltages being applied when the instrument is set to measure resistance; a high positive voltage will be "isolated" by the resulting high-resistance state of D_4, while if a negative voltage is applied D_5 clamps the collector of Tr_4 to the "0-volt" line, the fuse blowing if the current through the diodes is excessive. The current-generating circuit is disconnected from the input terminals in all cases except those in which the two selector switches, S_2 and S_3, are correctly set for a resistance measurement.

Choice of code

The 4,2,2,1 code was selected for use in this instrument in preference to the binary 8,4,2,1 code to make it impossible to set up a number greater than 9 in any decade, as this might be confusing. The digits 2 to 7 inclusive in each decade can thus each be set up in two different ways, which fact can be used for occasional "cross-checks", and for short cuts in getting a reading, e.g. if the first "2" has been found too large, the "1" in the same decade should next be tried, "by-passing" the second "2" switch, to save time in getting the reading.

It would appear that a 4,3,1,1 code and a 3,3,2,1 code would offer similar advantages, and there may be little to choose between these three codes from either mathematical or "hardware" considerations, but perhaps some "human factor" reason why one of the three codes is to be preferred. If it has not already been done, it would seem desirable to resolve this choice as soon as possible, so that an optimum standard code for this type of instrument can be made widely known.

Mains power supplies

Mains-derived d.c. supplies of similar voltage, viz. nominally +18V and −3V, may be used in place of the batteries; as mentioned above, there is considerable latitude on these values, +12V and −2V being adequate. The "B+", "B−" and "−" input terminal may conveniently be used for the connections, and as long as the on/off switch S_1 is left at "off" the batteries need not be removed; but if the positive supply voltage to be used exceeds the battery voltage the positive battery lead should be unclipped to prevent unwanted conduction through diode D_1; diode D_2 will of course continue to give overload protection as long as the positive rail remains connected to a low-impedance voltage source.

Construction

Fig. 1 is a photograph of the instrument, for which a die-cast box approximately 10.75 × 6.75 × 2.25 in. has been used.

All the signal terminals, switches and controls are mounted on the lid of the box, with labelling as shown for the benefit of sighted users or helpers. On the underside of the lid are the two electronic circuit boards —one carrying the d–a resistor network, the other all the active devices and associated components—most of the interconnecting wiring, and the fuseholder. The two boards have flexible lead connections and may be pivoted away from the underside of the lid for servicing, if necessary.

Only the battery housing and the loudspeaker are fitted to the lower part of the box. These are connected by flexible leads to the relevant points on the underside of the lid, so that the instrument can be operated with the lid removed and inverted for servicing, as shown in Fig. 4. A pivoted flap locked by a single knurled nut gives access to both battery housing tubes, one of which houses two PP4's for the positive supply, the other two SP2's for the negative supply. The loudspeaker is mounted facing downwards, a number of holes drilled in the base of the box, and the rubber feet which support the unit, providing an adequate air path for the sound; the speaker thus occupies no top panel space, and should also remain relatively dust-free.

For all external connections 4mm socket terminals are used. The two terminals on the left in Fig. 1 are the signal input terminals. For these two plain crocodile clip leads are provided, together with a third lead (not shown in Fig. 1) with an insulated probe clip and a multiplier box a few inches from the 4mm plug end, containing a series resistor chain which gives the multiplier lead a resistance of approximately 20 megohms. This multiplier lead is used, in either the positive (upper terminal) line or negative (lower terminal) line, as appropriate, in the measurement of positive and negative voltages in excess of 100 volts: it converts the 50-volt and 100-volt ranges to 500-volt and 1000-volt ranges. (Neither input terminal is connected to the case. Either can be so connected, as required.)

A little to the right of the input terminals are a monitoring terminal and screwdriver adjustment potentiometer for the 10-volt stabilized supply, while a fourth terminal, in the upper right-hand corner, enables the reference voltage present at the comparator to be monitored or connected to other equipment.

The two remaining terminals, also on the right-hand side of the panel, monitor the positive and negative battery voltages respectively. The instrument can be used to measure its own battery voltages, the positive one directly; the negative voltage can be measured by connecting to it the negative terminal of a 4.5-volt battery, measuring the net positive voltage and subtracting it from the external battery voltage.

Operation

The rocker switch array and l.s.d. microswitch bar are clearly shown in Fig. 1. To the left of them are the on/off switch and volume control, and above them (i.e. further from the operator) are the two range selection switches. The right-hand range selection switch is set to the most anti-clockwise position ("V") for all voltage readings, and to the next position ("R") for all resistance ranges except "1M" (1 megohm), for which the third position of the switch is used, and "HR" (high resistance), for which the fourth, most clockwise, switch position is used.

With the right-hand switch set at "V", the left-hand range selection switch gives voltage ranges of 100V, 50V, 10V, 5V, 1V, and 1V (again), starting at the anti-clockwise limit. The first two of these ranges are converted to 1000V and 500V respectively by using the multiplier lead in place of one of the two direct input leads, as described above. On the 500V, 50V and 5V ranges the numerical value obtained from the rocker switches must be divided by two (or the switch weightings thought of as 2,1,1,$\frac{1}{2}$ instead of 4,2,2,1), so that the other ranges, on which the numerical value obtained is directly relevant, seem likely to be preferred except where maximum resolution is required.

With the right-hand range selection switch set at "R", resistance ranges of 1000 ohms, 10,000 ohms and 100,000 ohms are obtained from the left-hand switch set at positions 6 (most clockwise), 5 and 4 respectively. For the 1 megohm and high resistance ranges the left-hand switch is set at position 4 and the right-hand switch at "1M" or "HR" as required.

Measurement of resistance is straightforward on all but the "HR" range, i.e. the rocker switches having been set and read, and the position of the decimal point determined according to the range in use (as in voltage measurements), the numerical value obtained is a reading in ohms of the resistance connected between the input terminals. On the "HR" range the same constant current supply is used as on the 1 megohm range but, as R_{29} is included in the circuit, the reading obtained represents the resistance of this and the external resistor in parallel, from which the value of the latter can be calculated ($R = x/(1-x)$ megohms, where R is the external resistor being measured and x the reading obtained as a fraction of full-scale). This method is used to avoid the necessity for an even lower constant current supply on the "HR" range than the 5 microamps required on the 1 megohm range, as ideally the current used should be large compared with possible variations in comparator input current for the latter to cause negligible inaccuracy.

x	0.50	0.667	0.75	0.80	0.833	0.875	0.90	0.95	0.98	0.99
R (megohms)	1	2	3	4	5	7	9	19	49	99

Using the expression $\frac{1}{1-x} - 1$ megohms for R may be found to simplify the mental arithmetic; the "-1" may be ignored with little error at very high resistance values. Where only a rough measurement of a high resistance is required, it may be found expedient to do a rough interpolation between a few memorized spot values on the "HR" range as shown in the table.

For the blind user wishing to avoid mental arithmetic completely, a comprehensive conversion table in braille for the "HR" range could be prepared, and might be worthwhile if many high resistance measurements to maximum accuracy were required. The "HR" range is the one rather complicated range to use on this instrument, but the facility it gives was considered worthy of inclusion.

In setting up the rocker switches according to the frequency (high or low) of the audible output, the operator performs manually a sequence of a type which is carried out electronically in certain a–d converters; this may sound laborious, but in practice proves quite quick and acceptable. A blind user has reported, after only a few days' use, that setting the switches takes very little time indeed, and has pointed out that when using his moving-coil multimeter he must allow a little time for the needle to settle before clamping it; and he considers the few seconds involved in setting the switches a small price to pay for the accuracy of the voltage and resistance measurements, which was just not available to him previously.

The readout from the rocker switches is of course of a clear and positive nature, requiring no braille labelling, and is hence readily usable also by blind people not familiar with braille, as well as by sighted people.

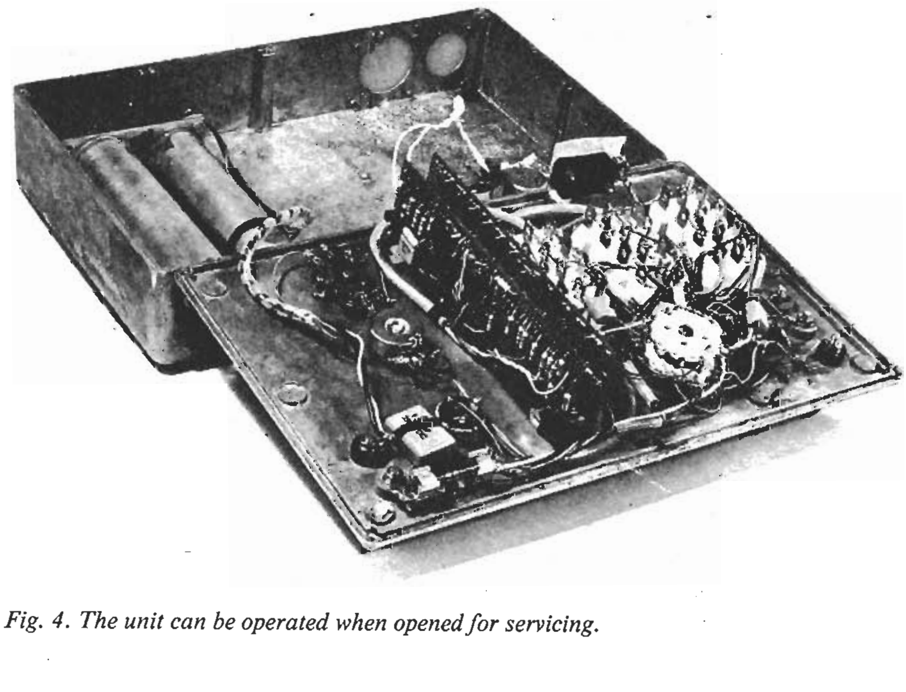
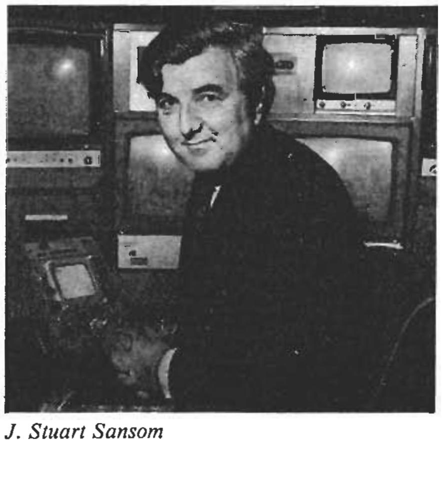


Fig. 4. The unit can be operated when opened for servicing.

About people

J. Stuart Sansom, O.B.E., F.I.E.R.E., has been appointed director of studios and engineering at Thames Television. Mr Sansom began his career in television in 1953 with a firm working on the development of high-quality telerecording systems, joining Television Wales and West in 1957 to become head of maintenance. He later went to ABC Television as head of development, in which capacity he was involved in the investigations into colour television systems prior to the final choice of PAL. He became chief engineer of Thames in 1966, and was appointed technical controller in 1970. Mr Sansom has served on many committees, including the EBU Technical Committee.



J. Stuart Sansom

The Measurement of Loudness

Harry F. Olson*

THE ULTIMATE significant subjective destination of original or reproduced sound and noise is the human ear. Therefore, the varied responses of the auditory system are particularly important factors in the reproduction of sound. One of the response functions of the human hearing mechanism is loudness. The purpose of this paper is to describe a loudness meter based upon the fundamental principles of the loudness response of the human hearing mechanism.

When a sound or noise of any quality or structure impinges upon the human ear, the magnitude of the resultant sensation is termed the loudness. It is the intensive attribute of an auditory sensation in terms of which sounds may be ordered on a scale extending from soft to loud. Loudness depends primarily upon sound pressure but it also depends upon frequency and waveform of the stimulus. The units on the scale of loudness should agree with common experience estimates about the magnitude of the sensation. The measurement of loudness is a significant part of the audio art because the loudness of a sound or noise plays an important role in the reproduction of sound.

Loudness is functionally related to sound pressure level, frequency, and waveform. Turning this around, the sound pressure level as measured by a sound level meter does not indicate the loudness of a sound. However, a conversion can be made in the readings of a sound level meter employing octave band pass filters to obtain the loudness. This is indeed a long and tedious process, as the exposition in this paper will show. What is required is a loudness meter that indicates the loudness of a sound in real time. Furthermore, the loudness indication should agree with the loudness as perceived by the listener.

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There are many uses for a loudness meter. For example, the loudness meter can be used to monitor the loudness of an audio program so that the peak permissible levels of all manner of audio program material will provide the same loudness to the listener. In the production of contemporary recorded music one of the objectives is to obtain the maximum loudness. For a certain maximum amplitude level of the signal, which is determined by the constraints of the record, a loudness meter can be employed to obtain the maximum program loudness by modifications of the frequency balance and timbre present in the signal. There are many other uses for the loudness meter in the measurement of sounds and noises.

Loudness Scale

The establishment of a loudness scale is a very complicated matter. A large number of investigators in many countries have carried out research on the loudness of a complex sound. A detailed description of the work is beyond the scope of this paper. Therefore, only the basic data on loudness required for the development of a loudness meter will be presented.

The unit of loudness is the *sone.* A sone is defined as the loudness heard by typical listeners when confronted with a 1000 Hz tone at a sound pressure level of 40 phons.

The loudness level of a sound is given by

$$P = 20 \log_{10} \frac{p}{p_o}$$

where p = loudness level, in phons,
P = measured sound pressure, in microbars, and
p_o = a sound pressure of 0.0002 microbars

The loudness level [1] of a sound or noise is expressed as n phons, when it is judged by normal listeners to be equally loud compared to a pure tone of frequency 1000 Hz consisting of a plane progressive sound wave radiating to the observer, the sound pressure of which is n (decibels) above the standard ref-

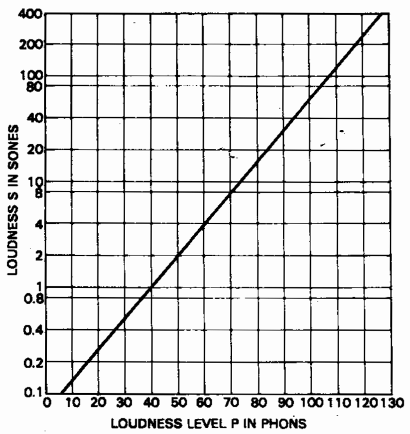

Fig. 1—The relation between the loudness in sones and the loudness level in phons.

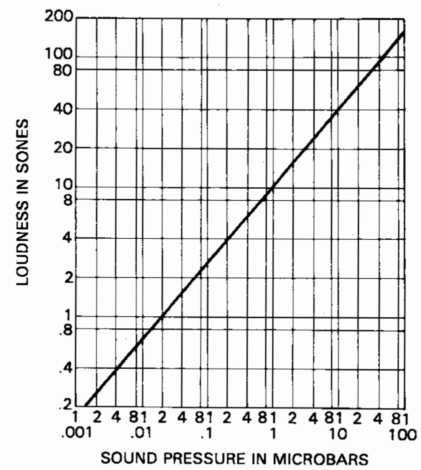

Fig. 2—The relation between loudness in sones and the sound pressure in microbars.

erence sound pressure of 0.0002 microbars.

The relation[1] between loudness in sones and loudness level in phons is given by

$$S = 2^{(P-40)/10}$$

where S = loudness, in sones and
P = sound pressure level, in phons, given by equation 1.

The relation between the loudness in sones and the loudness level in phons is shown by the graph of Fig. 1.

The relation between loudness in sones and sound pressure in microbars, shown by the graph of Fig. 2, indicates that there is a nonlinear relationship between the loudness in sones and sound pressure in microbars.

Measurement Of Loudness

In order to provide a measure of the loudness for the complex sounds of speech, music, and noise, there must be a means to separate the complex sounds into manageable segments. In particular, to establish the loudness of a complex sound, at least three specifications must be available as follows:

1. A scale of subjective loudness. This is termed the *sone scale* described in the preceding section.
2. The equal loudness contours for discrete frequency bands of the complex sound.
3. The rule by which loudness adds as the discrete frequency bands of the complex sound are added.

If specifications 1, 2 and 3 can be established, then the loudness of the

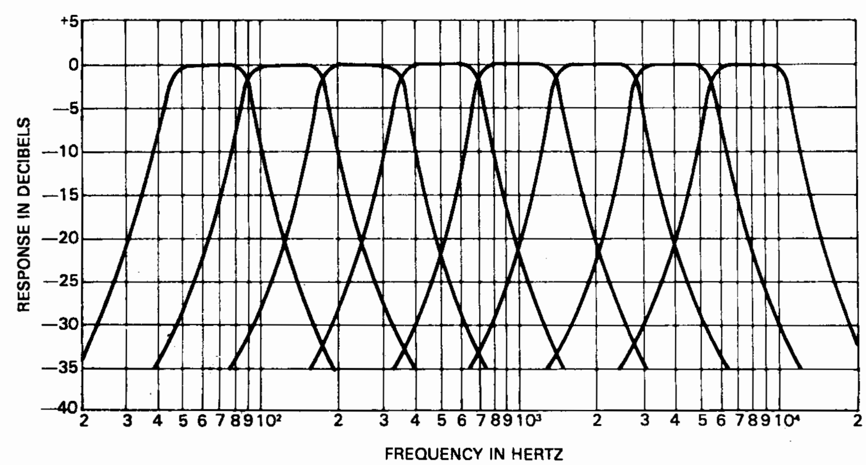

Fig. 3—The frequency response characteristics of the octave band pass filters.

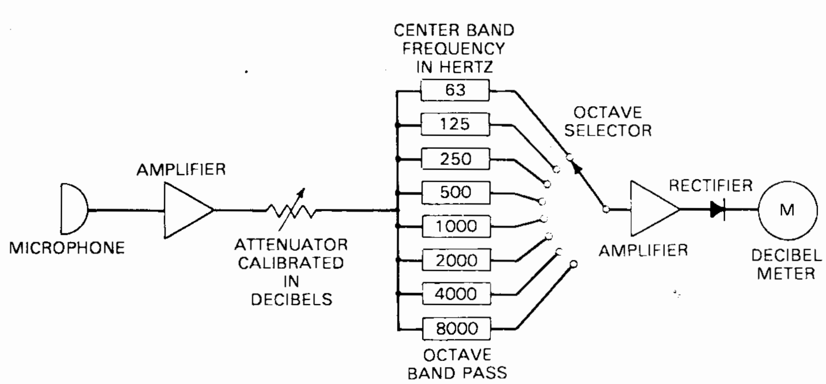

Fig. 4—Schematic diagram of a sound level meter for measuring the sound level in an octave.

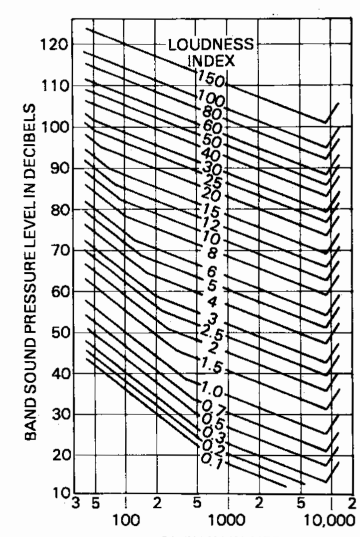

Fig. 5—Contours of equal loudness index for octave bands in the audio frequency range.

complex sounds of speech, music, or noise can be determined. The objective and subjective information[2,3,4] relating to the specifications of items 1, 2 and 3 have been established by investigators concerned with the subject of loudness. Furthermore, these investigators have shown that the loudness of a complex sound can be determined from the physical data on the complex sound in conjunction with the specifications of items 1, 2 and 3.

The specific method for determining the loudness of a complex sound is to split the audio frequency range into frequency bands. This is a complex procedure in which the complexity increases with the number of frequency bands. From a practical standpoint there should be as few frequency bands as possible without sacrificing frequency selectivity. A suitable frequency band appears to be the octave. The frequency response characteristics of the octave band pass filters employed in this development project are shown in Fig. 3.

The system for determining the sound pressure level in the eight different octave bands in the audio frequency range is shown in Fig. 4.

When the sound pressure level in each octave band has been measured, the next step is the proper summation of these data to provide the total or overall loudness of the complex sound. In this investigation and development, the procedure selected for calculating the loudness of a complex sound is the one developed by S.S. Stevens.[2] This is also the standardized procedure[5] as given in ISO-R532 Method A. In accordance with this Standard, the relation between the total loudness and the loudness index in each octave band is given by

$$S_T = 0.7\, S_M + 0.3\, \Sigma\, S$$

where S_T = total loudness of the complex sound, in sones,
S = loudness index in each octave band, and
S_M = greatest of the loudness indices.

The loudness index is obtained from the graph of Fig. 5. The sound pressure level in each octave band is determined by means of the system of Fig. 4. Employing the geometric mean frequency for each octave band, the loudness index for each octave band is determined from Fig. 5. Then the total loudness of the complex sound in sones is computed by means of equation 3.

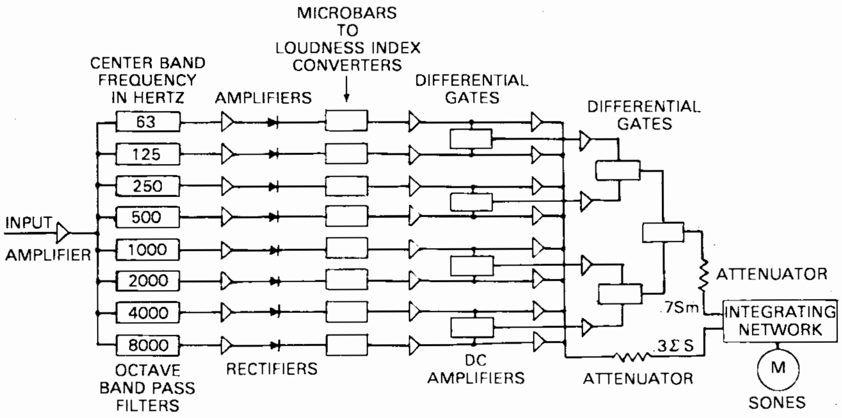

Fig. 6—Schematic diagram of the elements of a loudness meter.

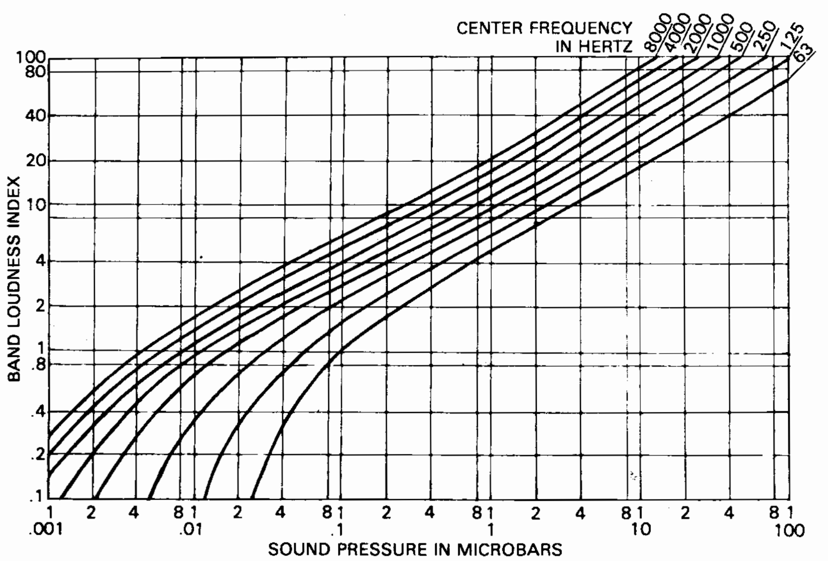

Fig. 7—The relation between the loudness index and the sound pressure in microbars for the octave bands of Fig. 3.

Loudness Meter

To provide a loudness meter requires an automated instrumentation of Fig. 4 incorporating the data of Fig. 5 and the procedures of the preceding section operating in real time. Specifically, equation 3 shows that the loudness meter must provide the following: the measurement of the loudness index in each channel, the summation of the loudness indices in all the channels, the selection of the channel with the greatest loudness index, the proper relation and summation of the sum of the loudness indices and the highest loudness index, and an indicating meter with the proper dynamics to depict the loudness of the complex sound in sones from the summation input.

A schematic diagram of the loudness meter[6] is shown in Fig. 6. The signal input is fed to eight octave band pass filters. The frequency response characteristics of the filters are shown in Fig. 3. The output of each band pass filter is followed by an amplifier coupled to a rectifier, which in turn is followed by a microbar-to-loudness-index converter. The input-output characteristics of the eight microbar-to-loudness-index converters are shown in Fig. 7. These converters are in the form of nonlinear active elements as a part of operational amplifiers and their output is fed to a d.c. amplifier. The output of these amplifiers provides the loudness index for each octave channel, and the loudness index outputs from the eight channels are added by means of separate direct current amplifiers and fed to the attenuator coupled to integrating network and the sone meter. In order to determine the channel with the highest output, the eight microbar-to-loudness-index converters are fed to differential electronic gates in the form of a net-

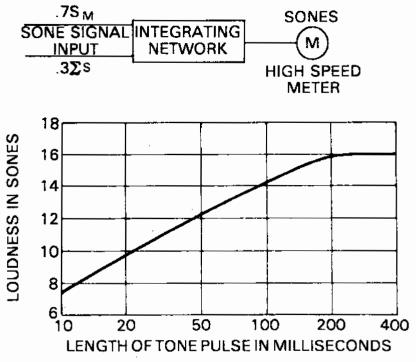

Fig. 8—The integrating network and high speed meter of the loudness meter. The graph depicts the relation between the loudness in sones and the length of the tone pulse in milliseconds.

work tree, the output of which is fed to an attenuator coupled to the integrating network and the sone meter. The two attenuators are adjusted to obtain the correct values of 0.7SM and 0.3ΣS. Under these conditions the sone meter will indicate the loudness in sones of an audio signal input to the loudness meter.

The remaining and very important subject is the dynamics of the amplitude characteristic of the indicating meter. The amplitude response of the indicating meter system should correspond to the ear response to individual, repetitive and overlapping short, medium, and long time pulses of sound and continuous sounds. Since most sounds of speech and individual musical instruments are of short duration, this then becomes a very important problem. The basic question is the loudness of a complex sound as a function of the time the sound persists. Obviously, a short pulse of sound of amplitude equal to a long pulse of sound will exhibit a lower loudness level. From published data and data obtained from this development (loudness as a function of the time length of the sound pulse), the graph shown in Fig. 8 was drawn. As would be expected, the loudness of a relatively short time pulse of sound decreases with the duration of the time of the pulse. This data was used to develop an integrating network in conjunction with a high speed indicating meter. A block diagram of the integrating network and the high speed meter for indicating the output of the loudness meter in sones is shown in Fig. 8. The integrating network consists of active growth and decay networks applied to an operational amplifier. Since the main intended application for this meter was the determination of the loudness of speech and music, the integrating network was tailored to provide the correct indication of loudness for this type of program material.

The signal input to the loudness meter should correspond to the level of the reproduced sound. For example, the average listener prefers a loudness level of the reproduced sound of 80 phons. The input to the loudness meter should be adjusted so that a level of 80 phons will give an indication of 16 sones.

Performance Of The Loudness Meter

A large number of subjective tests have been carried out to determine the performance of the loudness meter employing reproduced speech and music. A few of the tests and results will be described.

Test No. 1. The reproduced sound level of a *musical* program was varied over wide limits. The observers agreed that the indication of the loudness meter agreed with their sensation of loudness.

Test No. 2. The reproduced sound level of a *speech* program was varied over wide limits. The observers agreed that the indication of the loudness meter agreed with sensation of loudness.

Test No. 3. The same musical program was reproduced in highly compressed and uncompressed conditions. The compressed program was reproduced at a level of 2 dB lower than the uncompressed program as read on a conventional volume indicator (VU meter). The loudness meter indicated a level 3 sones higher for the compressed program. Here the two meters indicated a reversal in the readings. The subjective evaluation by the observers agreed with the loudness meter. This shows the conventional volume indicator does not indicate loudness.

Test No. 4. Speech was recorded at a low speaking level and at an almost shouting level. The two were reproduced at the same top level of 80 dB as indicated by a conventional volume indicator (VU meter). The shouting speech indicated a higher loudness on the loudness meter. Again the observers agreed with the loudness meter.

Test No. 5. Employing a contemporary musical program, the loudness meter was used to provide the maximum loudness for the same peak level as the unchanged program. The main operations were compression and changes in the frequency distribution. Employing a reproduction peak level of 85 dB as indicated on a peak reading level meter, the modified program indicated a loudness 6 sones higher than the unmodified program. This is an increase in loudness level of 6.7 phons which is indeed a considerable increase in the sensation of the loudness.

Summary And Conclusion

A loudness meter has been described which indicates the loudness of an audio signal. Since the ultimate significant subjective destination of all original or reproduced sound is the human ear, a meter which indicates the loudness as perceived by the ear is an important audio instrument. For example, the loudness meter will become a very useful tool for determining the loudness of any simple or complex sound or noise, for monitoring the maximum permissible level of all manner of audio programs, for obtaining the maximum loudness of an audio program for a certain maximum peak level, etc.

The author wishes to express his appreciation to R. A. Hackley, D. S. McCoy, and D. G. Murray for contributions to the development work of the loudness meter. Æ

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A Sound-Level Meter

How to Build and Use a Sound-Level Meter with One-Third Octave Filters

DAVID GRIESINGER

SPECIFICATIONS	
Sensitivity	30 to 120 dB SPL when used with microphone having sensitivity of −60 dB re 1V/μbar
Frequency Range	20 to 20,000 Hz in three ranges
Frequency Accuracy	±10%
Input Impedance	100 k ohms at Mic input 1.0 Meg at high-level input
Power Required	9 V at 2.5 mA from transistor radio battery. Life about 10 hours.

Most serious recorded-music listeners are aware that the sound they eventually hear depends to a large extent on the acoustics of their listening room. Even after determining the best possible placement for the speakers, and perhaps adding some drapes or a rug, the system as a whole can sound muddy, or lack the lifelike quality of the more-fortunate systems. True realism depends on having a flat frequency response where you listen, and most rooms will not cooperate. Fortunately, frequency response is easy to vary electronically. Several devices have appeared recently which are intended to deal with this problem, and simple electronic filters are really quite easy to make. Unhappily, adjusting such a filter can be frustrating. It is difficult to determine just what corrections to make by merely listening to music. One of the most meaningful ways of measuring the frequency response of a hi-fi system is to measure the response of the system to narrow bands of filtered noise. The little instrument described in this article, when combined with a high-quality omnidirectional microphone, is capable of making this kind of measurement as well as many others. The sound level meter is not tricky to build, and should cost less than fifty dollars.

All the meter needs to measure room response is a calibrated preamplifier, a tunable active filter, and a VU meter. To measure the sound levels of music and machines, a bass filter is included to give the standard "A" weightings. The result is a true sound-level meter, capable of measuring sounds down to 20 dB SPL and third-octave noise down to 0 dB SPL. With a microphone sensitivity of −60 dB re 1V/μbar, this corresponds to a pure tone input of only 0.18 μV rms, making it possible to measure the noise output of almost anything.

Before describing the circuit, I think I should say something about the way the instrument is to be used, and how that use determines the choice of microphone. Figure 1 shows the response of my living room to pure tones and to third-octave noise, with the mike placed where I do most of my listening. The wide variations in the pure-tone response exist in all rooms, and correspond to resonances or standing waves similar to the ones in organ pipes. The number of possible frequencies for these resonances or standing waves is very large, since the sound wave can bounce around the room in many directions, using combinations of the walls, floor, and ceiling as reflectors. We hear the combination of all the reflecions. For some frequencies, these reflections add up in phase, creating a peak, and for others they cancel each other out, giving us no sound at all. The ear is much more sensitive to peaks in the response than to dips.

Above 200 Hz the response to pure tones looks similar, except that it becomes rather difficult to measure. The variations in the response get very close together on the dial of the oscillator, and unless you tune very slowly it is easy to miss them entirely.

Fortunately, music sounds much better in such a room than this curve would seem to predict. Musical tones are continuously changing, and the individual resonances do not have time to build up. The response that we subjectively hear is the average of nearby resonances, and is a lot smoother than the pure-tone curve. Unfortunately, it is still not very flat. To measure accurately how a room sounds to music, one must use tones which resemble music in the way they waver. Although music generally wavers about a semitone, or a twelfth of an octave, noise filtered through a third-octave filter gives good results. Figure 1 shows that the third-octave curve follows the response to pure tones. The noise curve may look smooth, but it must be treated with respect. A dip or a peak of three decibels in a third-octave-noise curve is almost always audible, especially if it occurs between 300 and 2000 Hz.

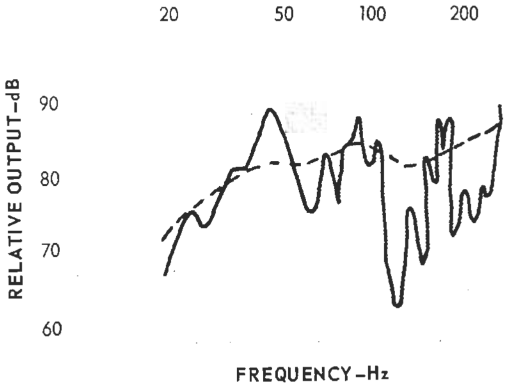

Fig. 1—Low-frequency response of the author's room with AR-5 placed near a corner and microphone at room center. (- -⅓ octave noise;—pure tones)

The basic principle of this discussion has been that sound in a room does not simply come from the speaker to you, but forms standing waves which surround you. To measure the intensity of the sound correctly you must use an omnidirectional microphone. Cardioid microphones have a predictable response in a very large room or in an anechoic chamber, but when put into a standing wave their output can vary widely.

Microphone Selection

The microphone used must be at least as good as the system you are trying to measure. Fortunately, there are many high-quality omnidirectional condenser microphones available. These microphones have been designed for recording, but they can easily be used for sound measurement, especially if you do not need to know the absolute level of the sound.

There is a complication to using a recording mike. Any omnidirectional microphone is only really omnidirectional at low frequencies. When the wavelength of the measured sound becomes close to the diameter of the microphone, diffraction effects tend to reduce the response to any wave which does not come from the front. If the sound in your room comes from all directions, some roll-off to the measured sound will be noticed at high frequencies. To compensate for this effect some manufacturers build a treble rise into the response of microphones designed to measure noise. These mikes, such as the B&K 4134, are flat to randomly incident noise (Fig. 2). Recording mikes are usually closer to the B&K 4133, with a flat on-axis response. Most people would shudder at a treble rise in the on-axis response, although when using an omnidirectional mike in a large live hall some treble boost is often necessary.

Notice from Fig. 2 that diffraction effects depend strongly on the diameter of the microphone. B&K types 4131 and 4132 are one-inch microphones, 4133 and 4134 are half-inch, and 4135 and 4136 are quarter-inch. For a microphone of ¾-inch diameter or less, these diffraction effects occur mostly above 4 kHz, and are serious only above 10 kHz. The sound in a room is usually not omnidirectional at those frequencies. Some loss of treble will be noticed due to the nature of the microphone, but if it is pointed directly at the speaker this roll-off is not too serious. In any case it is not possible to predict exactly how much it will be. When making measurements at these frequencies it is better to look for the smoothness of the response. If you want an exact number, the best way to get it is to point the microphone 90 deg. from the speaker, and correct the resulting measurements for the response of the microphone to randomly incident noise. Some microphone manufacturers supply such a curve with the mike. For those which don't, it tends to be close to the response to pure tones at 90 deg. incidence.

The condenser microphone chosen should have a well known on-axis response, and be ¾-inch diameter or less. If you intend to make a large number of speaker measurements, a ½-inch sound-measurement microphone might be a good investment. Otherwise it would be wise to choose your microphone for its use in other applications, such as recording. Remember that good omnidirectional microphones of similar diameter and flat on-axis response will sound identical. Cardioid microphones are currently popular, but I prefer to use an omni wherever I can. They tend to add echo in a natural way, and lack the increase in bass when close to the source of sound.

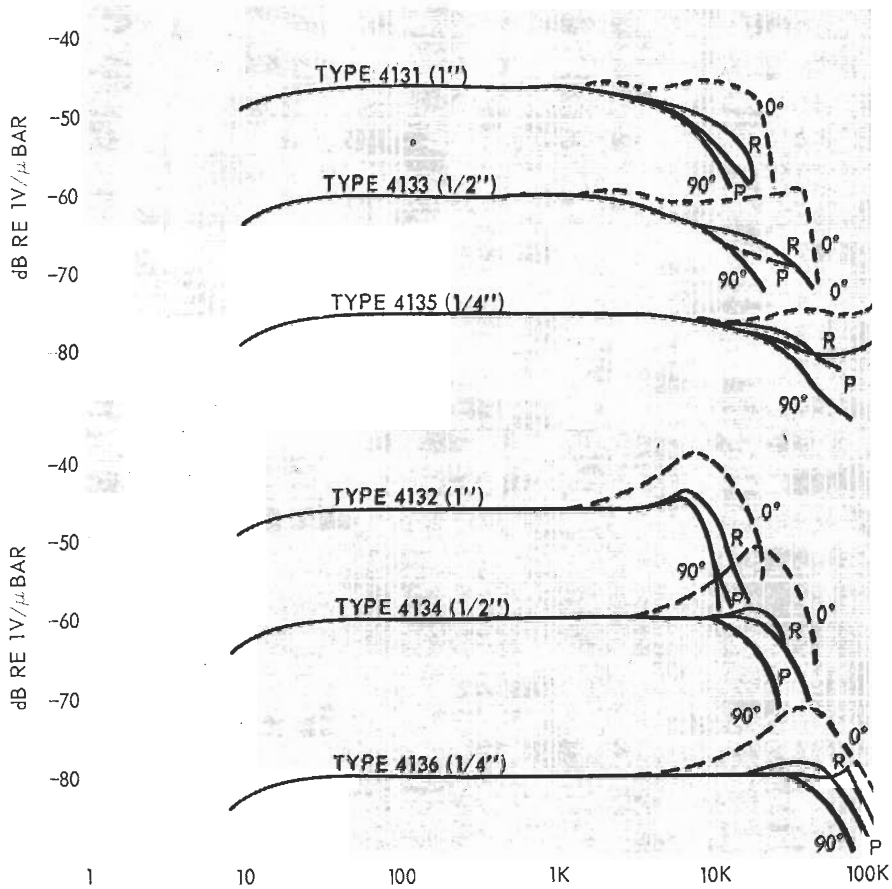

Fig. 2—Typical frequency responses of some B & K condenser microphones. 0° = free-field response at 0° (normal incidence); 90° = free-field response at 90° (grazing incidence); R = random incidence response, and P = pressure response.

Making the Measurements

Making speaker measurements is delightfully simple. You just feed some source of broad-band noise into the speaker system, and measure the resulting sound with the meter. Noise which sounds uniform in frequency content and has equal energy content per third-octave band is called pink noise. If you have such a generator, use it. Otherwise, use the interstation noise from an FM tuner. You may have to disconnect the antenna to get noise which sounds uniform enough. This noise is not pink, but you can correct your curves by measuring it electrically with the meter.

The microphone should be placed in the position you use for most of your listening. If you get too close to the speakers, interference between the drivers can give misleading results, although you can test for high-frequency dispersion this way. You should occasionally turn off the speakers to make sure that the background sound in the room is below the level you are measuring. Measure each speaker separately, since they usually require different compensation. If you insist on having a composite curve for the whole system, be sure to set the tuner on stereo to get independent noise in both channels. Otherwise interference effects can occur.

The measured response is usually less than wonderful! What can you do about it? Try moving the speaker first. The standard things to try are lifting the speaker off the floor or away from the wall by a foot or so, to the long side rather than the short side of a room, or vice versa. You should avoid having two reflective surfaces facing each other. I prefer to make these changes by ear, using a choral recording or a record of a familiar voice.

Your response will probably still show various slopes in the midrange. These slopes are very important. Try varying both the settings of the speaker balance controls and the amplifier tone controls. If you can't do much this way, you must add some compensating networks between the preamplifier and the amplifier. If your preamp has low output impedance when compared to the input impedance