# Spot-frequency distortion meter

## Measures very low (0.00001%*) levels of harmonic distortion.

**by J. L. Linsley Hood**

This article describes a spot frequency distortion measuring instrument which uses a bootstrapped notch filter technique to avoid typical parallel T problems of 2nd and 3rd harmonic attenuation. Oscillator amplitude stabilization is achieved by a Darlington-based Wheatstone bridge arrangement with a thermistor controlling currents in each limb. The final combination of oscillator, notch filter and wide bandwidth millivoltmeter offers marked improvements in noise factor and linearity, permitting the resolution of much lower levels of harmonic distortion than is normally possible.

THERE IS NOW considerable interest among engineers in the use of distortion measuring systems as a general tool for circuit performance analysis. While this can most conveniently be done by the use of a spectrum analyser, giving rapid identification of the nature of the harmonic impurities, with equipment of this type the lower level of detectable distortion is usually about —80dB or 0.01%, while the areas of current interest are 10 to 100 times less than this. For these applications therefore, the somewhat laborious methods of notch filtering for a single measuring frequency are still required.

As it is feasible, with relatively simple circuits, to measure waveform impurities below 0.0001% (—120dB) and to generate sinewaves with impurity contents of about 0.0002%, the methods employed here may be of interest to those engaged in circuit analysis, as a means of attaining a more detailed view of non-linearity. In order to reduce the complexity of construction, the equipment was designed to operate at five 'spot' frequencies within the audio band — 100Hz, 300Hz, 1kHz, 3kHz and 10kHz.

### Measuring apparatus

The most straightforward way of determining the amount of distortion present in a pure sinusoidal waveform is to interpose a sharply tuned notch-filter between the input waveform and a measuring circuit and while there are several suitable filters, the most convenient of these is the 'parallel T' network, shown in a schematic measuring apparatus in Fig. 1. The transmission and impedance characteristics of a simple T network are shown in Fig. 2, which demonstrates the difficulty inherent in the use of a passive 'parallel T' of this type in the signal path. There would be significant attenuation of both the second and third harmonics of the incoming signal, leading to an inaccurate measurement of the level of distortion.

The sharpness of this notch can be increased by the application of overall negative feedback around a loop containing the 'parallel T' and a suitable

*For 10V r.m.s. input signal.

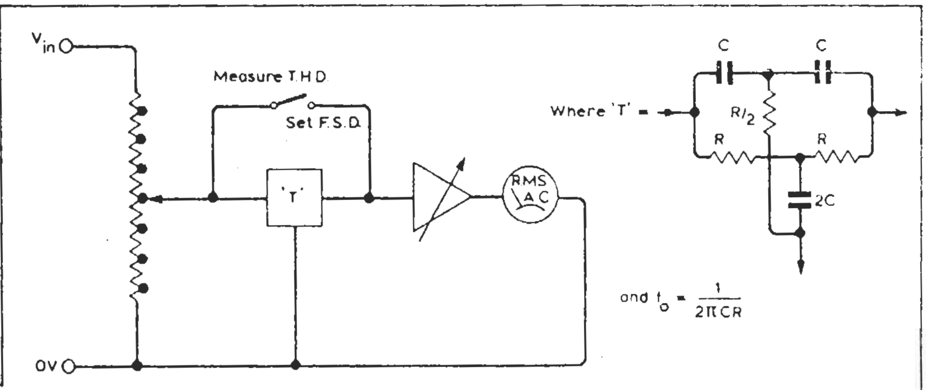



*Fig 1: Conventional parallel T distortion measuring arrangement and the basic T circuit*

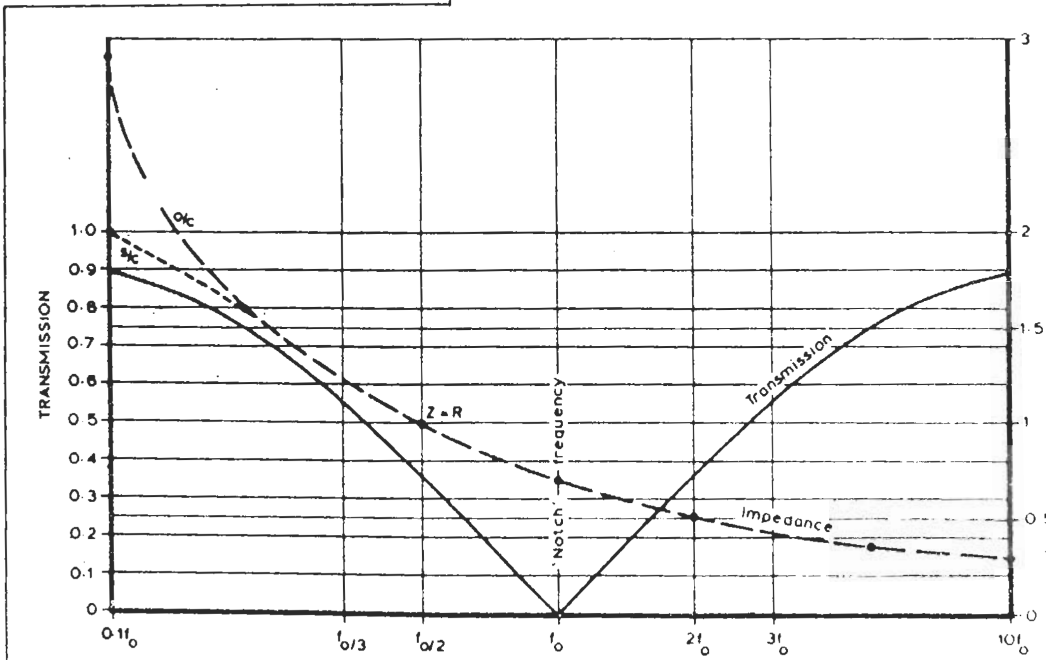



*Fig. 2: Typical attenuation of 2nd and 3rd harmonics in the parallel T signal path*

following amplifier so that for the same attenuation at the notch frequency, the transmission at $f_n/2$ or $2f_n$ can be made substantially identical to that at much lower or much higher frequencies and this is an arrangement which has been employed in commercial 'parallel T' distortion meters[1]. Unfortunately this method suffers from the disadvantages that the input circuit is made more complex and that there is some injection of amplifier noise into the notch filter, lessening the sensitivity of the system.

An alternative approach, which leads to simpler circuit configurations, is to apply positive feedback to the 'common' limb of the T, by means of a 'bootstrap' arrangement of the type shown in Fig. 3. This leaves the input to the T free from other circuit connections, so that it may be taken directly to a low impedance input attenuator. The sharpness of the notch can be controlled by the extent to which the 'bootstrap' voltage approaches that of the input voltage to the amplifier. In general, too sharp a notch will make the equipment less easy to operate, so the proportion of the input voltage applied to the 'bootstrap' connection is chosen so as to achieve a generally flat response in respect of second and higher harmonics.

The characteristics of the notch filter, with regard to both the notch frequency and its equivalent output

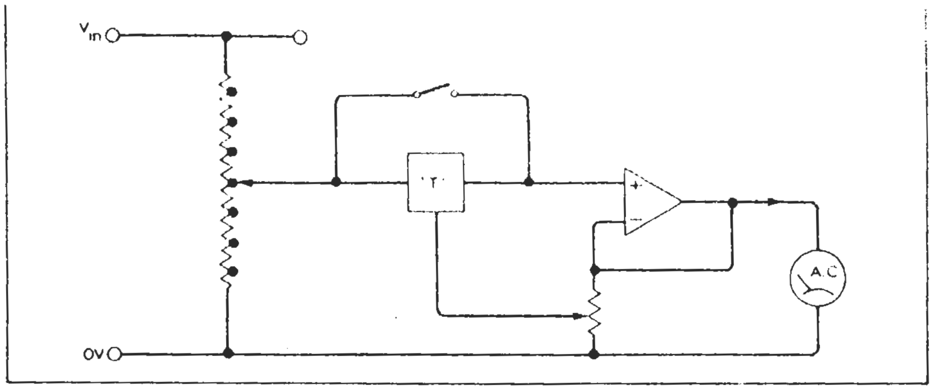



**Fig 3:** *Bootstrapping the network*

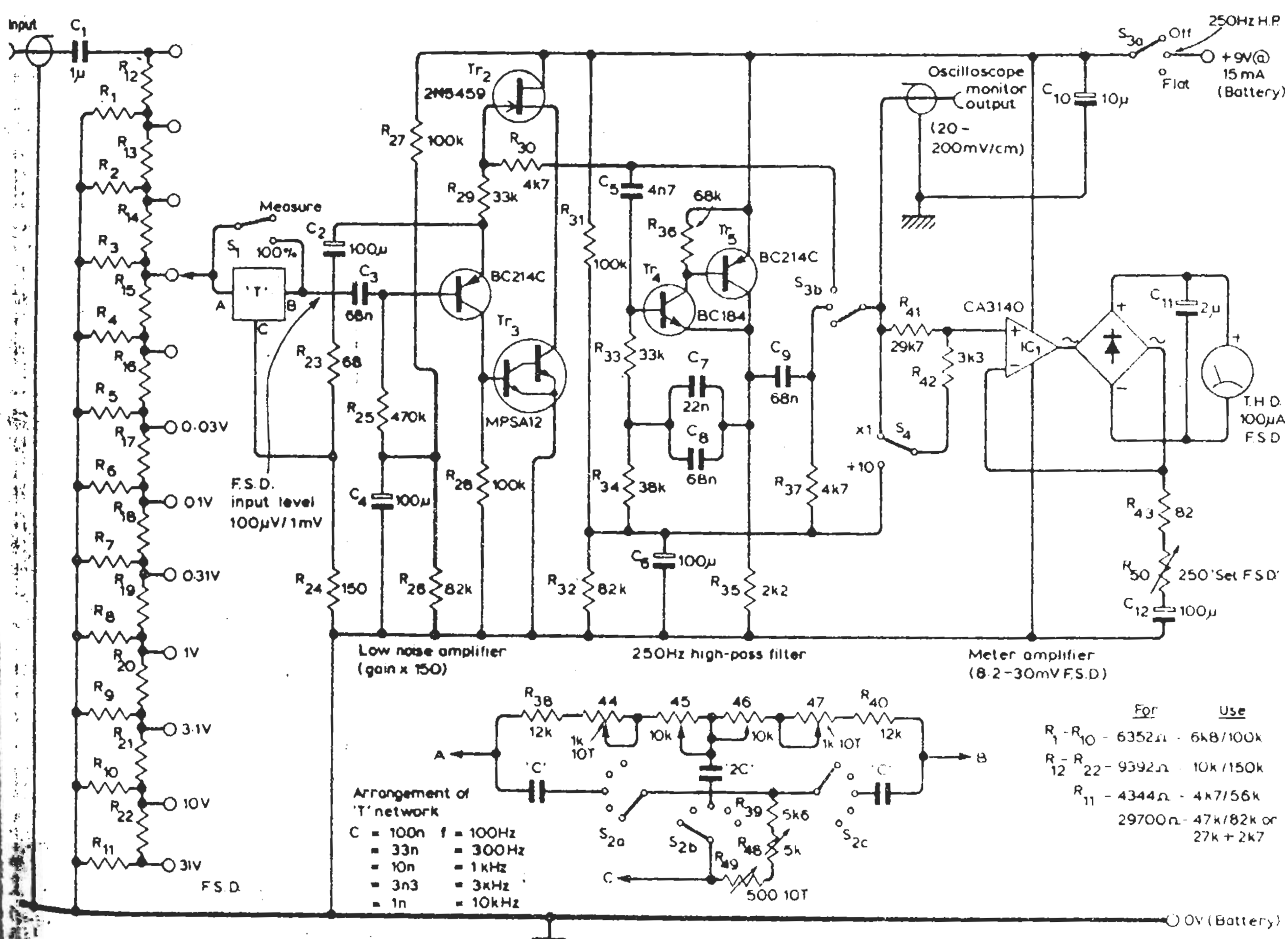
and 'noise' impedances, are influenced by the impedance seen at the input to the T. The input attenuator should therefore be of the constant impedance type. Suitable values for this can readily be calculated[2]. Ideally, the parallel T should be fed from an impedance which is not more than one-tenth of the nominal impedance of the T and the following amplifier should have at least 10 times its input impedance over the frequency range of interest.

**Fig 4:** *Distortion meter circuit. Instructions for making up odd value resistors ($R_1$-$R_{10}$ etc.) mean "use" 6k8 and 100k in parallel*

## Bootstrapped T circuit

A suitable electronic circuit, which employs a bootstrapped T as the notch element, and largely satisfies the circuit requirements is shown in Fig. 4. In this, the T network is fed from an input attenuator having a voltage attenuation of $\sqrt{10}$. (3.162) or 10dB, and a characteristic impedance of 3.3kΩ. The output of the T is taken to a low-noise amplifier with an input impedance of about 300kΩ, and a gain of 150. The effective input noise is mainly determined by the impedance characteristics of the T.

A wide-bandwidth a.c. millivoltmeter



is driven from this amplifier through a two position (x1 and x1/10) attenuator and an optional 250Hz, —20dB/octave, 'bootstrap' filter[3], with a high-pass characteristic. The use of an RCA CA3140 c.m.o.s. operational amplifier allows an effective 100kHz bandwidth, ± 1dB, for the meter circuit. The full-scale sensitivity of the meter circuit is adjustable by the 'set f.s.d.' control over the range 8.2 - 30mV. The complete instrument can be operated from a 9 volt transistor radio battery and the current consumption is approximately 15mA.

Tuning of the notch to the nominal 'spot' frequencies is by means of a 10k twin-gang and 5k single-gang potentiometer. Fine tuning is then accomplished by two 1k and one 500Ω ten-turn potentiometers.

The ultimate sensitivity of the instrument, assuming an adequately low noise component in the input signal under test, is less than 0.0001% for a 1 volt (r.m.s.) input signal, or less than 0.00001% for a 10 volt (r.m.s.) input signal. At these harmonic distortion levels, even assuming adequate freedom from mains-frequency hum — which can be obtained with care in the screening of the instruments and the layout of connecting leads — the effectiveness of the plug and socket connections is extremely important and gold-plated connectors should be used if available.

## Operating the instrument

The method of operation of the instrument is relatively simple, in that the input attenuator is used in two roles, that of adjusting the input magnitude of the signal fed to the instrument, and that of adjusting the f.s.d. harmonic distortion reading. The technique is as follows — assuming an appropriate sinusoidal signal is applied to the input of the instrument, the sensitivity is progressively increased by moving the slider of the input attenuator switch ($S_1$) upwards from the lowest sensitivity (30V r.m.s.) position until a suitable setting is found, at which a full scale deflection can be obtained on the output meter with $S_4$ in the 'x1' and $S_5$ in the '100%' position.

$S_3$ is then switched to the 'measure' position, and $S_1$ is moved upwards towards the maximum sensitivity setting, with each upward step corresponding to a 10dB increase in the meter display sensitivity. In percentage terms, this gives a step sequence of 100%, 31.6%, 10%, 3.16%, 1%, and so on. If an input voltage of 1 volt (r.m.s.) is applied, the maximum sensitivity position will correspond to a f.s.d. value of 0.01%. Since the input noise of the instrument, integrated over the 100kHz measuring bandwidth, gives a meter deflection of less than 1% of the full scale, a reading due to the harmonic residues and other components of the input signal (0.001%) can be seen on a suitable meter. If a 3 volt input signal is available, the maximum f.s.d. input sensitivity setting would be equivalent to 0.00316% and if a 10 volt signal were available, a full scale deflection equivalent to 0.001% would be provided, allowing minimum detection levels of 0.00003% and 0.00001% respectively.

These assumptions have been checked in practice using an oscillator whose t.h.d. at 1 volt (r.m.s.) output was measured at 0.0002% and when amplified to the 10 volt level through the best available amplifier gave a reading of 0.00018% on the 0.001% f.s.d. setting. Once again, at these levels, the fitting of the plug and socket connections is critical and the notching-out of the fundamental is a matter of some skill.

Although the component values for the notch filter of Fig. 4 are those chosen to give an adjustment of ± 30% about the mean centre frequencies, it is obviously practicable to extend this so that the ranges overlap.

The 'scope output point can be used for a visual or instrument analysis of the harmonic structure of the residues and provided that the fundamental has been removed more simple techniques are often adequate such as a phase sensitive rectifier operated from an external oscillator, frequency-locked through a p.l.l. to a simple multiple of the input frequency.

For simplicity, an average-reading millivoltmeter has been employed as the output meter rather than a more complex 'true r.m.s.' (thermal energy equivalent) system.

A minor practical snag in the use of this instrument with the simple constant impedance input attenuator shown is that the capacitive coupling of the input signal to the input to the T across the attenuator switch leads to a small change in the notch frequency as the input attenuation level is changed, with the consequent need for some readjustment of the null frequency. Better input screening could avoid this.

## A low-distortion spot frequency oscillator

A similar, but rather more complicated, 'parallel T' distortion meter was built some ten years ago and used as a test instrument for the assessment of oscillator performance characteristics, — a number of experimental oscillator circuits were examined by this means. This exercise was instructive in many ways, of which the two most vital were the demonstration of the need for a very low noise level (which precludes the use of most integrated circuits) and the need for very high frequency stability, if a fundamental-nulling measuring technique is to be used.

The attainment of a stable operating frequency demands a highly frequency-selective feedback network and of the many forms available, the 'parallel T' offers the best ratio of performance to complexity. If this type of network is included in the feedback path of a high gain inverting amplifier, of the form shown in Fig. 5, the system gain will be very high at the null frequency and will tend to zero at frequencies remote from this. If the T is slightly unbalanced, in the manner proposed by Bailey[4], so that some positive feedback is given at the notch frequency, the system will oscillate without further persuasion, though with some small penalty in waveform purity due to the lesser discrimination in the feedback ratio.

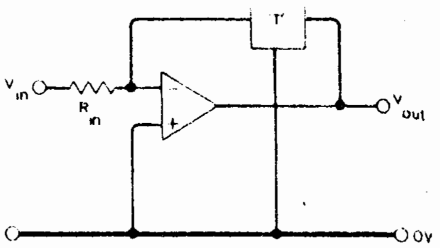



**Fig 5:** *A high gain null circuit.*

Alternatively the T network could be left as a straightforward notch element, having zero transmission — and consequently zero negative feedback — at some specific frequency and a small magnitude sinusoidal signal could be injected into the amplifying device from an external source. Since the amplified signal will have a high degree of monotonicity, because of the frequency/gain characteristics of this circuit arrangement it could be anticipated that the harmonic distortion of the input signal would be lessened by such an amplifier stage.

If this input signal could now be derived from the output of the amplifier through a network which adjusted the magnitude of the fed-back signal in a manner which maintained the output at a constant amplitude, the result would be an oscillator having a waveform purity and signal-to-noise ratio determined almost exclusively by the effectiveness of the amplifier arrangement.

The design problem therefore simplifies into that of providing a circuit block of low input noise level, overall good linearity (especially in respect of the output stage which has to handle the greatest signal level) coupled with as high an open-loop gain as is practicable, and some means of deriving a feedback signal from the output of the amplifier whereby its magnitude and phase can be made to be dependent on the output signal level.

In view of the requirement that the amplifier stage should be phase inverting, the practical choice of amplifier configuration is limited to that of a single gain stage or a two-stage amplifier with the input devices arranged as a long tailed pair. The difficulty of obtaining overall loop stability in a feedback amplifier having three or more gain stages connected in cascade discourages the consideration of this alternative where a wide bandwidth is necessary.

Initial exploration of the first of these two possibilities showed that it was possible to obtain stage gains in the range 50,000 to 100,000 from a single transistor in a Liniac[5] configuration if the amplifying device was isolated from its load by an f.e.t. in the manner shown in Fig. 6. However, the need to couple the amplifying stage to an output point having an impedance of some 600 ohms required an impedance transformation circuit which added considerably to the component count and detracted from the original simplicity of the concept.

If a two-stage design is chosen, it is essential that the gain of the first stage is sufficient to ensure that the noise contribution of the second can be ignored. In general this implies that a relatively high first stage load is necessary, which in turn indicates the choice of either a field-effect device as the second stage amplifier, or a compound configuration of junction transistors. One of the monolithic small-signal Darlington devices meets this requirement admirably and has an input impedance which is sufficiently high to have little effect on the impedance of the collector load of the preceding stage. Also, the stage gain of such a device feeding a constant-current source has been shown to be of the order of 2000-3000[6].

Taylor shows[7] that the use of an input long-tailed pair, because it is basically a push-pull configuration, leads to the cancellation of even-order harmonic distortions, particularly when the devices are matched in characteristics and operating conditions, but also even when the devices are mismatched. A possible gain stage of this type, using an input long-tailed pair and a Darlington transistor second stage is shown in Fig. 7. This has a low-frequency open-loop gain of the order of 200,000 or geater, which allows a substantial measure of loop feedback to be applied and avoids the pitfall demonstrated by Baxandall[8] that low levels of negative feedback may exchange a small measure of non-linearity for a whole host of high-order distortions.

**Fig 6:** *Exploratory high gain amplifier*

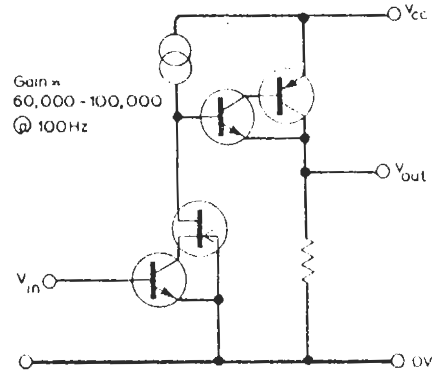
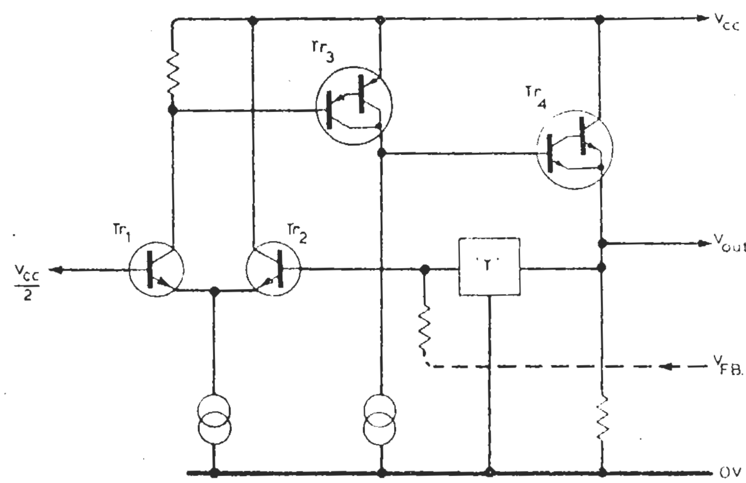



**Fig 7:** *Simplified design which approaches the shape of the final version*



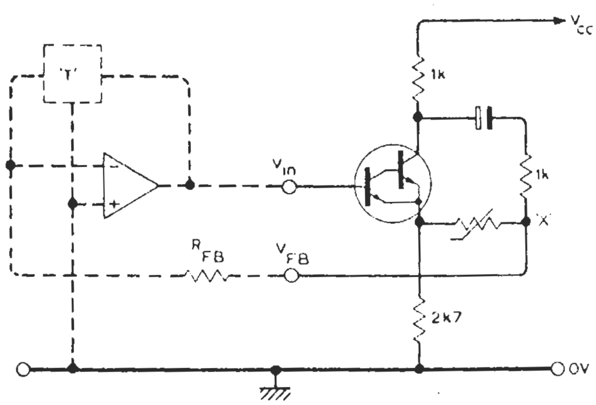



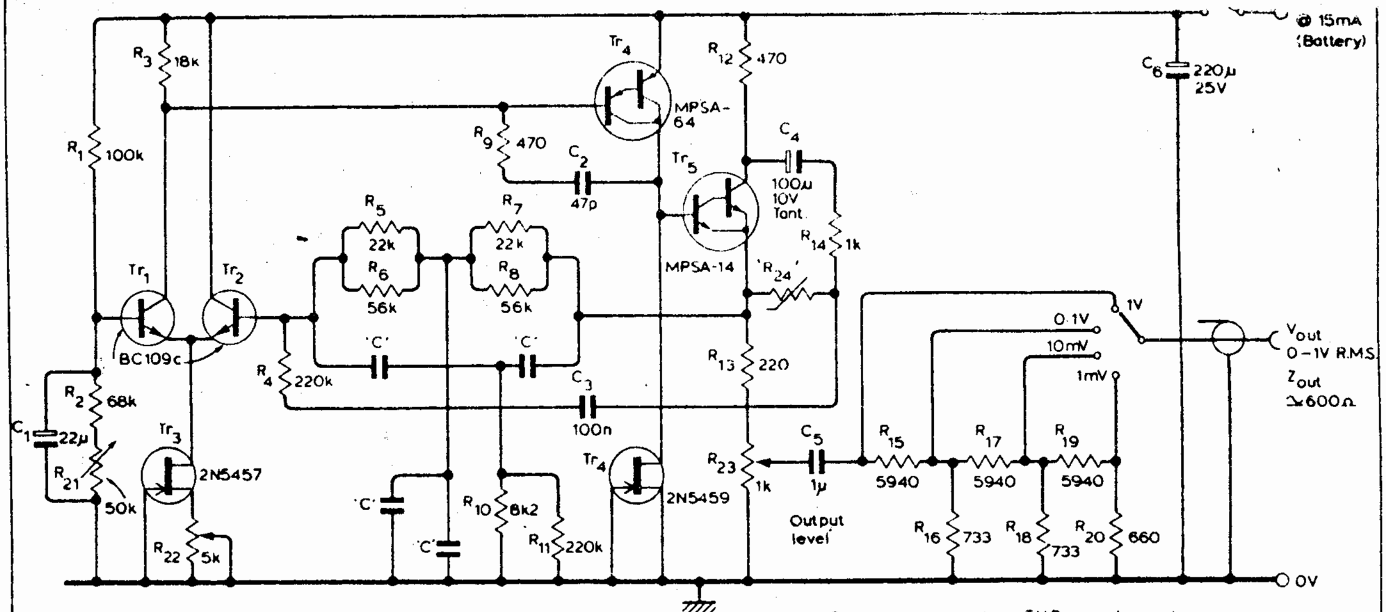

**Fig 8:** *Thermistor-controlled "Darlington bridge" – this part of the oscillator helps to stabilize output signal amplitude*

## Output amplitude stabilisation

The stabilisation of the amplitude of a low-distortion oscillator is a difficult problem, for reasons explained previously[9] and this difficulty is exacerbated by any requirement that the amplitude stabilisation circuit should contribute as little as possible to the overall distortion figure. In this case, the technique adopted is that shown in Fig. 8. This takes advantage of the fact that in a Darlington transistor, the collector and emitter currents are substantially identical and this allows the thermistor to be operated as one limb of a Wheatstone bridge type configuration. Since the ratio of the collector to emitter limbs is 1 : 2.7 the bridge will be balanced (for zero output at point 'X') when the thermistor achieves an internal resistance value, due to the heating effect of the circulating current, of 2.7kΩ. If the applied voltage to the control circuit falls, the thermistor will cool somewhat, which will cause the phase of the feedback signal to be positive, thereby increasing the magnitude of the output. If the magnitude of the signal input to the control circuit increases, then the resistance of the thermistor will fall and the phase of the feedback signal will become negative, causing the output of the oscillator to decrease.

In operation, the total magnitude of the signal present at the base of $Tr_2$ is very small, so that the non-linearity contribution due to the curvature of $I_c/V_b$ characteristics of the input devices is also very small. This demonstrates one of the reasons for the superiority in performance of this (parallel T) type of circuit over the conventional Wien bridge system, in which there is normally one-third of the output signal present at the base of the input transistor with consequently greater contributions from the input device to the overall non-linearity of the circuit.

The final circuit of the oscillator is shown in Fig. 9 and the measured distortion characteristics are shown in Fig. 10. Loop stabilisation is achieved by



Small improvements in T.H.D. may be made by adjustments to $R_{21}$, $R_{22}$ & $R_{11}$. These are best done at 10kHz

**Fig 9:** *Complete schematic of the spot-frequency oscillator*

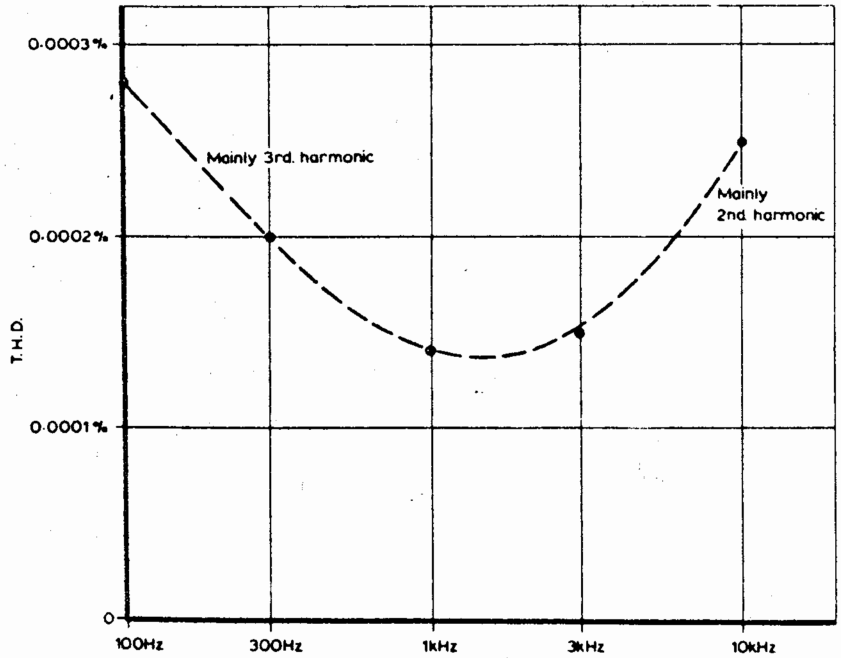



**Fig 10:** *Measured distortion characteristics*

adding a dominant lag capacitor between collector and base of $Tr_4$. The values shown have proved adequate to prevent squegging in three experimental models of this oscillator, but in two of the three cases a 3pF capacitor was quite adequate, with consequent improvement in the h.f. open-loop gain and rather lower t.h.d. figures at 10kHz than those shown in Fig. 10.

While the author's own model of this unit operates only at the five spot frequencies shown, obtained by switching the capacitors in the T (polystyrene foil types) there is no reason other than complexity of switching for the restriction of its operating frequencies to those shown.

The first unit of this type was built using resin dipped carbon film resistors, and this is still in service. A subsequent unit employing metal film resistors throughout showed a small improvement both in t.h.d. and background noise level. Unfortunately for the conclusiveness of this experiment, a similar improvement in the prototype was obtained by replacing the Motorola 2N5089 input devices with Motorola BC109Cs. The f.e.ts are also preferably Motorola types.

Thermistor "R24" should be an STC R54 or equivalent type. Resistance at 20°C should be approximately 50k falling to about 270Ω in operation. This makes other items such as the GM473 or VA3410 suitable.

The measurement of the residual harmonic distortion and noise is greatly facilitated by the monitoring of the signal on an oscilloscope. Even so, the operation of any notch filter with a rejection ratio of 120dB or so requires a certain delicacy of touch and conditions of reasonable tranquillity.

### Printed circuit boards

Two glass fibre p.c.bs are available for the distortion meter and oscillator at a cost of £5 the set from M.R. Sagin, 23, Keyes Rd., London, N.W.2.

### References


1. Radford type A.71 distortion measuring set.
2. Linsley Hood, J.L., Low distortion oscillator, *Wireless World*, September 1977, p.42.
3. Linsley Hood, J.L., The "H" or "bootstrap" l.f. circuit filter, *Electronic Engineering*, July 1976, pp.55-58.
4. Bailey, A.R., Low-distortion sine-wave generator, *Electronic Technology*, February 1960, pp.64-67.
5. Linsley Hood, J.L., The Liniac, *Wireless World*, September 1971, pp.437-441.
6. Idem, p.439.
7. Taylor, E.F., Distortion in low-noise amplifiers, *Wireless World*, August 1977, pp.28-32.
8. Baxandall, P.J. Audio power amplifier design, *Wireless World*, December 1978, pp.53-56.
9. Linsley Hood, J.L., Low distortion oscillator, *Wireless World*, October 1977, p.70.