# Low distortion oscillator

## 1 — An improved Wien bridge design

by J. L. Linsley Hood

**This instrument uses a modified circuit to reduce the typical output harmonic distortion of a Wien bridge oscillator by a factor of up to 10. A sine wave output from 10Hz to 100kHz in four switched ranges is available together with a square wave. A constant impedance output attenuator is also provided with four switchable levels from 1V to 1mV. The complete unit can be powered from two 9V batteries.**

The Wien bridge arrangement shown in Fig. 1 is one of the most convenient circuit configurations for use in a wide range variable frequency oscillator circuit because the operating frequency can be made continuously variable by means of a twin-gang potentiometer for $R_x$, $R_y$, or ganged capacitor for $C_x$, $C_y$. However, most of the conventional Wien-bridge oscillators using this type of circuit, such as the example in Fig. 2, have a minimum distortion figure of

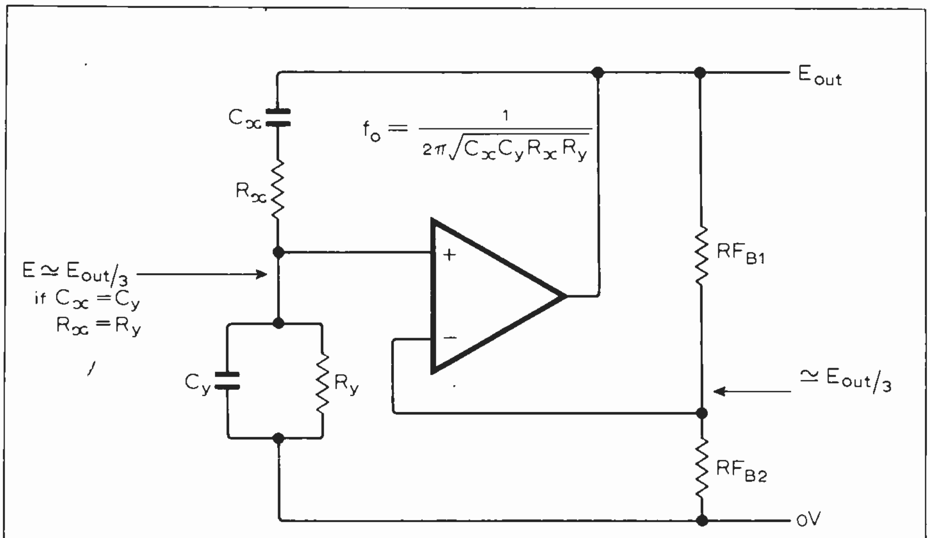



*Fig. 1. Basic Wien bridge oscillator circuit. For frequency variation either $R_x$ and $R_y$ are used as a twin gang potentiometer or $C_x$ $C_y$ as a ganged capacitor.*

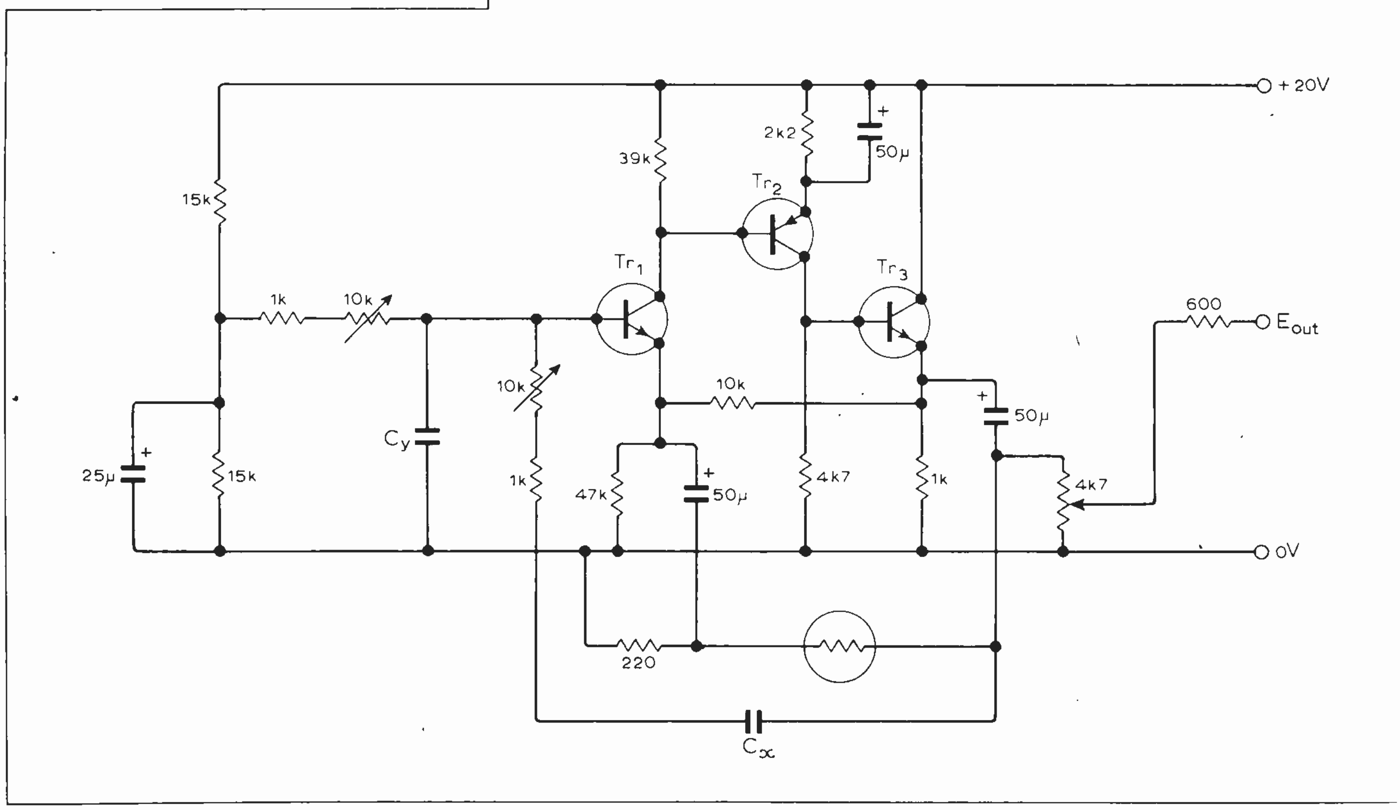
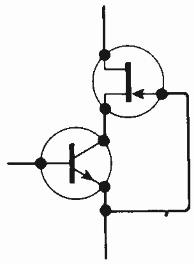
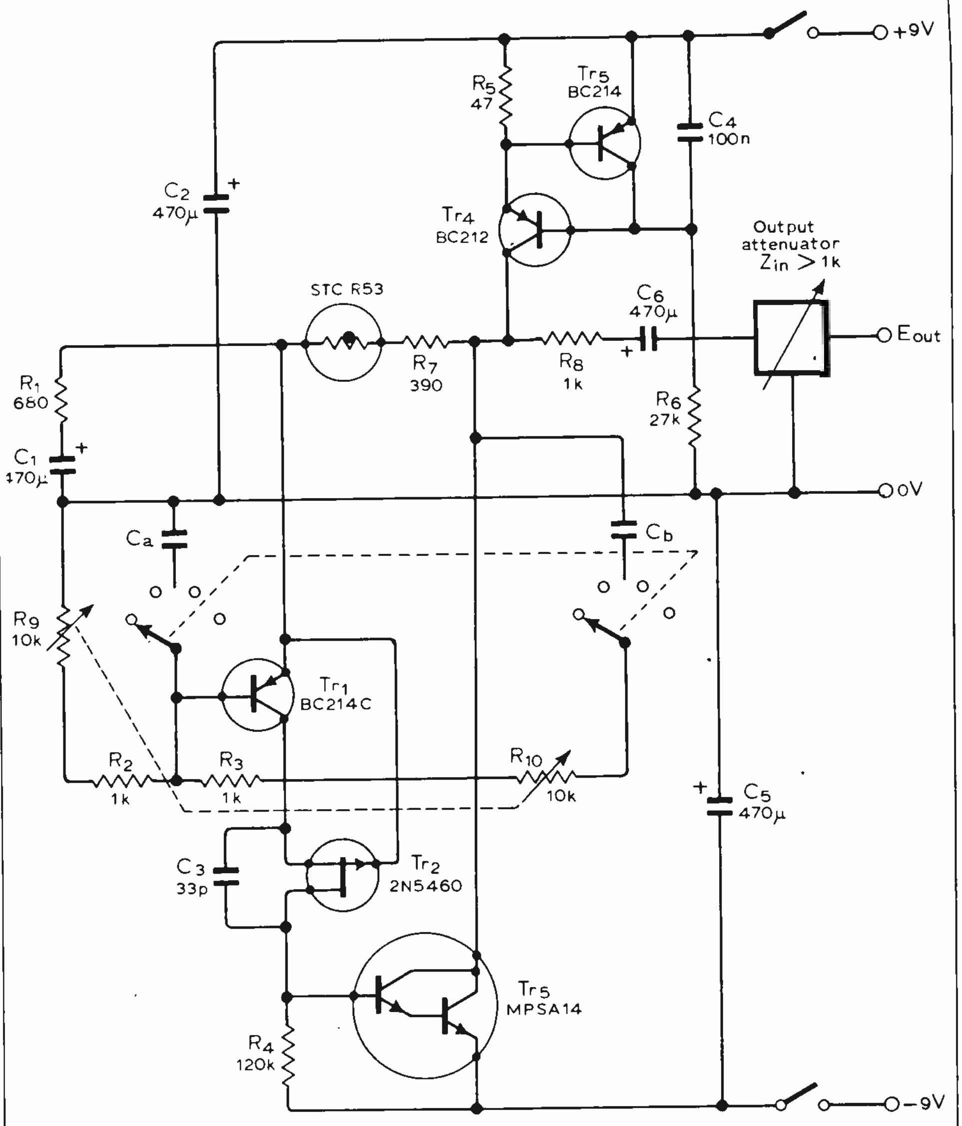



*Fig. 2. Conventional Wien bridge circuit which produces around 0.01% harmonic distortion.*





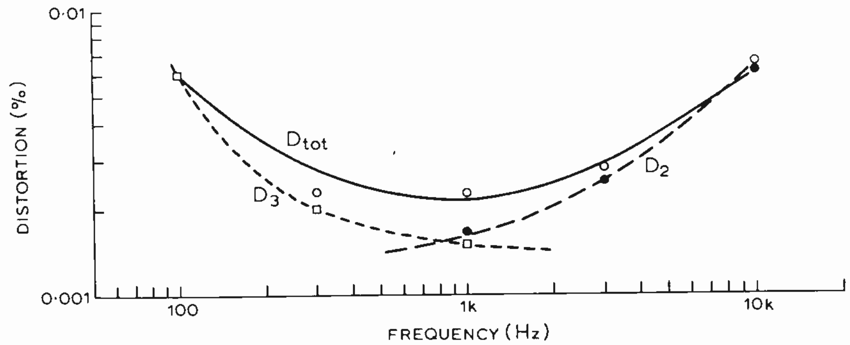




*Fig. 3. Bipolar f.e.t. cascode arrangement.*

*Fig. 4. Modified Wien bridge oscillator circuit. To reduce surface recombination noise a p.n.p. bipolar device is used in the input.*

*Fig. 5. Spot frequency distortion measurements of improved circuit.*

around 0.01 to 0.02% which is barely adequate for test purposes with modern amplifier designs. This has encouraged the more widespread use of the less convenient parallel T oscillator arrangement for producing very low distortion reference signals.

On analysis it is apparent that the major cause of residual harmonic distortion in the conventional thermistor stabilized Wien-Bridge oscillator circuit, at frequencies high enough for thermal modulation of the thermistor to be unimportant, is due to common mode failure[2,3] in the first stage amplifying device. Here the peak signal voltage applied in common mode to the base and emitter of $Tr_1$ is approximately $2\sqrt{2}/3\ V_{out}$ r.m.s., which can be a significant proportion of the available $V_{ce}$ in $Tr_1$. Improved performance can be achieved in three ways; by reducing the ratio of $V_{out}$ to $V_{cc}$, which may not be convenient. By reducing the magnitude of the signal voltage fed back to $Tr_1$, or, finally, by reducing the sensitivity of the input stage to common mode malfunction. In view of the high independence of output impedance and drain current with respect to drain voltage in most junction f.e.t.s, the use of an f.e.t. as the input device is attractive. The straight substitution of an f.e.t. for the input bipolar transistor, however, results in a large reduction in loop gain. The use of a bipolar device in cascode with an f.e.t. as shown in Fig. 3 overcomes this problem and offers a gain which is characteristic of the bipolar device together with an output impedance and common mode rejection ratio typical of a junction f.e.t. Moreover, the collector-emitter voltage of the bipolar input device is maintained at a constant potential, appropriate to the drain current drawn from the f.e.t., and as such provides a bootstrap action.

A practical circuit using this type of input configuration is shown in Fig. 4. Some small additional improvements in this circuit are the use of a p.n.p. input device, which produces less surface recombination noise in the junction, and the use of a constant current load for the output amplifying stage which gives greater output linearity and improved independence of $V_{cc}$. The typical t.h.d. of this design is shown in Fig. 5. Over the frequency range 200Hz to 3kHz, for an output of 1.5V r.m.s. into a 2kΩ load, the distortion content is between 0.0015 and 0.003% associated with a settling time of less than 2 seconds. This is independent of $V_{cc}$ in the range 13 to 20V or ±6.5 to 10V if a split supply is used as shown in Fig. 4.

Because most of the residual distortion arises in the output stage, somewhat lower values can be obtained for a given output load if the current, determined by $R_5$, is increased. For the values shown this is about 10mA.

The settling time of low distortion oscillators has been examined by Oliver[4,5] with the general conclusion that this will lengthen as the t.h.d. becomes lower, especially at lower frequencies because this is related to the number of cycles of signal applied to the thermally sensitive element. However, this is less of a problem with a Wien-bridge system compared to feedback networks which produce a transmission null at the operating frequency.

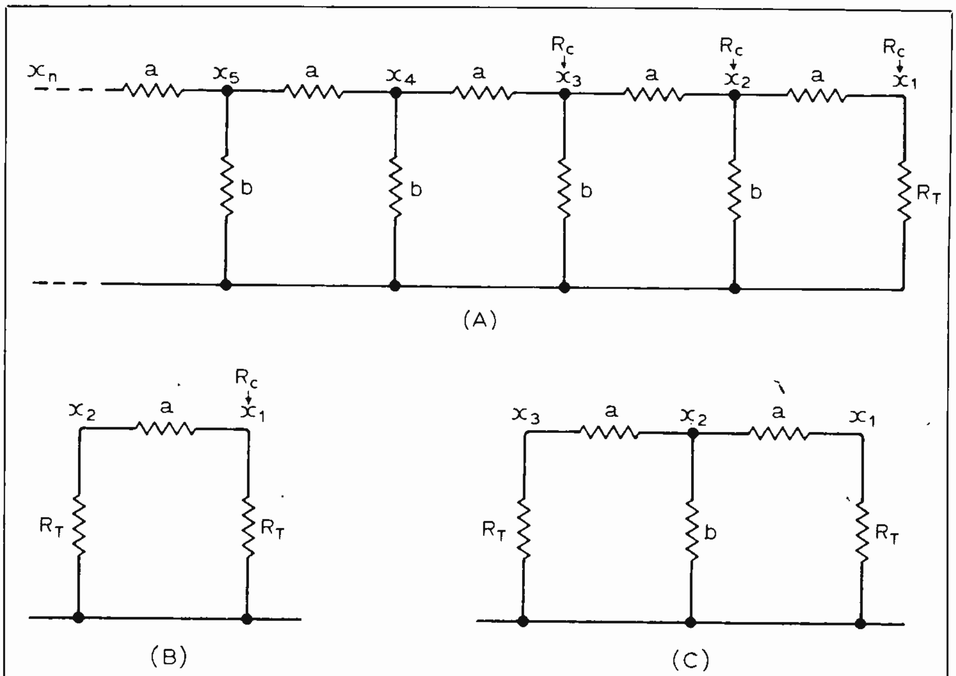



*Fig. 6 Basic resistive transmission line attenuator and the sections which are considered when calculating the resistor values.*

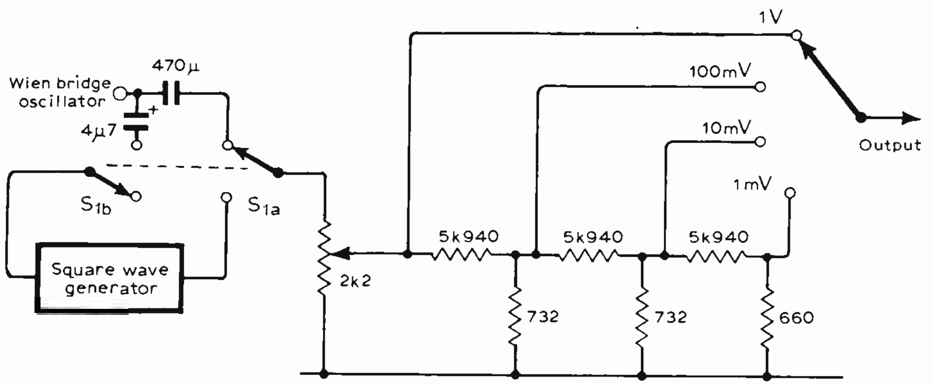



*Fig. 7 Practical attenuator. The 5k940Ω resistors can be formed by a 6k8//47kΩ, the 733Ω by a 6k8//820Ω, and the 660Ω by a 22k//680Ω.*

## Output attenuator

It is accepted as a practical convention that low frequency signal sources should have an impedance of 600Ω. The easiest method of achieving this is to take outputs from tapping points along a conventional resistive transmission-line attenuator as shown in Fig. 6 (A). Resistor values can be calculated for any desired characteristic impedance and attenuation factor, provided that the line is either of infinite length or is correctly teminated at both ends by resistor $R_T$.

The attenuation of the line from $x_2$ to $x_1$ is $R_T/\ a+R_T$ and if this is defined as $1/K$ then $K=a+R_T/R_T$ where $K$ is the reciprocal of the attenuation factor. If this definition is correct it must hold true for the shortest element of transmission attenuator as shown in Fig. 6 (B). The characteristic impedance of this line, as seen at $x_1$ and $x_2$ ($R_C$) is $R_C=R_T//(a+R_T)$, so

$$\frac{R_C}{R_T}=\frac{R_T}{R_T}//\frac{(a+R_T)}{R_T}$$

therefore $R_C/R_T=1//K$ which equals

$$\frac{K}{(1+K)}.R_T$$

therefore $R_T=\dfrac{(1+K)}{K}.R_C$

This defines the terminating resistors.

For calculation of the series resistor a, if the characteristic impedance of the line is specified and the attenuation characteristic is known,

$$K=a+R_T/R_T \text{ or } K.R_T=a+R_T$$

therefore $a=K.R_T-R_T$, which equals $R_T(K-1)$. As already shown,

$$R_T=\frac{(K+1)}{K}R_C$$

therefore $a=\dfrac{(K+1)(K-1)}{K}R_C$

so $\quad a=\dfrac{(K^2-1)}{K}R_C.$

To calculate the shunt resistor b, consider a line with these elements as in Fig. 6(C).

The impedance at $x_2$, as defined by $R_C$, is

$$b//\frac{(a+R_T)}{2}$$

or $\dfrac{1}{b}=\dfrac{1}{R_C}-\dfrac{2}{(a+R_T)}$

therefore $\dfrac{1}{b}=\dfrac{1}{R_C}-\dfrac{2R_T}{(a+R_T)R_T}$

$$=\frac{1}{R_C}-\frac{2}{R_TK}$$

$$=\frac{R_TK-2R_C}{R_CR_TK}$$

therefore $\dfrac{1}{b}=\dfrac{\frac{(K+1)}{K}.KR_C-2R_C}{\frac{(K+1)}{K}.KR^2_C}$

which equals $\dfrac{K+1-2}{(K+1)R_C}$

therefore $b=\dfrac{(K+1)}{(K-1)}R_C$

which allows the value of b to be calculated.

If a step attenuation of × 10 or greater is used, the influence of the source impedance to the line can be ignored. In the practical circuit of Fig. 7 the attenuator is fed from a potentiometer to give amplitude variation between ranges. The non-standard resistor values can be produced by the parallel combinations detailed in the caption.

## Printed circuit board

A p.c.b. which accommodates the Wien bridge oscillator, frequency range capacitors, square wave generator and output attenuator will be available for £3.00 from M. R. Sagin at 23 Keyes Road, London, NW2. The board follows the authors complete circuit to be published next month.


**References**

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