

A LOW-FREQUENCY OSCILLATOR WITH VERY LOW DISTORTION UNDER NON-LINEAR LOADING

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621.373.421

*This article is the second of a series on electronic circuits for special measuring instruments **). It deals with an oscillator (with fixed frequency) whose output voltage is required to meet exacting demands with regard to amplitude constancy and freedom from distortion under non-linear conditions that cause severe distortion of the current through the load.*

Valve oscillators under non-linear loading

The output voltage of valve oscillators is in most cases reasonably sinusoidal provided the load consists of a linear element. Non-linear loading, on the other hand, can cause severe distortion of the output voltage, and also make it very difficult to meet the condition for oscillation continuously. For certain purposes, however, a voltage is required which, even under severe non-linear loading, should contain only a very small fraction of higher harmonics. This was the case, for example, in a set-up used in this laboratory for measurements on magnetic amplifiers in which ferromagnetic cores undergo varying DC magnetization. For this purpose an oscillator was needed the current from which could show distortion up to 20% of the maximum current taken (r.m.s. value of the total higher harmonics 20% of the r.m.s. value of the fundamental component under full load) while the voltage distortion was not to exceed the very low value of 0.01%. Under varying load, and also under constant loading for longer periods, the amplitude of the output voltage was allowed to vary by no more than 0.1%. The specification of this oscillator was therefore as follows:

Output voltage	U_o	= 50 V approx.
Maximum power	P	= 2 W
Distortion in output voltage at 8 mA distortion current (20% of the nominal current under full load)	d	= max. 0.01%
Variation in output voltage . . .	$\Delta U_o/U_o$	= max. 0.1 %
Frequency	f	= 80 c/s

To this specification should be added that no current forms were to occur with a peak value higher than 120 mA. This limitation rules out those

cases in which the current would contain no more than 8 mA in higher harmonics, but where the voltage distortion requirement would not be met. A case in point would be a current consisting, for example, of a sinusoidal component plus a more or less pulse-shaped component. Keeping to a given distortion percentage one might, by taking the pulse narrow enough, make its amplitude arbitrarily high; in which case it would become increasingly difficult to keep the influence of the pulse on the voltage below a predetermined limit. The distortion requirement can therefore better be formulated by specifying that the *internal resistance* of the oscillator should be particularly low for multiples of the fundamental frequency, namely smaller than $10^{-4} \times 50 \text{ V} : 8 \text{ mA} = \text{approx. } 0.6 \text{ ohm}$. The specification of a maximum voltage variation of 0.1% under fluctuating load also boils down to a low internal resistance, now however at the fundamental frequency.

Interference signals induced in the output voltage should be kept to the same low percentage as the higher harmonics.

In the following we shall discuss an oscillator which meets the conditions mentioned.

Division into an oscillating and a power-output section

The combination of 1) positive feedback for the fundamental frequency such that the oscillation condition is constantly fulfilled, 2) strong negative feedback for the higher harmonics to keep the internal resistance at the low value required, and 3) measures for keeping the amplitude of the voltage constant, entails well-nigh insuperable difficulties in the design of a non-linearly loaded oscillator which is required to deliver energy.

A much more attractive principle is to let the oscillation and the delivery of energy be carried out by two separate sections. The first section, which remains

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**) For the first article see Philips tech. Rev. 24, 275-284, 1962/63 (No. 9).

unloaded, should generate a constant alternating voltage which has the required frequency and a distortion well below the specified limit; we shall call this voltage the *reference voltage*. The second section should closely "follow" the reference voltage. It contains a control system in which a certain fraction of the output voltage is compared with the reference voltage; the difference is amplified and drives the output stage which, even under non-linear loading, must be capable of delivering the required output voltage and power. Here, then, the principle of the stabilized power supply is applied, with this difference that the reference and output voltages are not direct but alternating voltages.

The first section will be called the *reference oscillator*, and the second the *output stage*.

The reference oscillator

When designing an oscillator one can generally choose between the *LC* and the *RC* type. Since the frequency in the present case is low (80 c/s), a coil with a particularly high inductance would be needed for an *LC* oscillator. Since the reference voltage must be just as free from interference voltages of outside origin as from higher harmonics, the screening of such a coil against stray alternating magnetic fields could be difficult. *RC* oscillators however do not require heavy screening, which was one of the main reasons for choosing this type.

The voltage delivered by conventional *RC* oscillators is unsatisfactory as a reference voltage, both in regard to distortion and constancy. The distortion is unsatisfactory because with the conventional *RC* oscillator the selectivity is obtained by using passive *RC* networks, which do little to reduce the distortion introduced by non-linear elements, such as valves. Given a signal around 10 V a distortion of about 1% is therefore normal. To keep the voltage constant, an amplitude-limiter is used — e.g. a thermistor (resistor with negative temperature coefficient, incandescent lamp) — which suppresses the gain as the amplitude increases. Over long periods a constancy better than about 1% is difficult to achieve in this

way. This is primarily due to the influence of the ambient temperature, the operation of thermistors being based on an energy balance.

In the solution adopted a further subdivision is made: the *reference oscillator consists of an amplitude-limiting part and a part which ensures that the oscillation condition is fulfilled*. As will presently be shown, the amplitude-limiting part delivers a voltage which, although its amplitude is very constant, is at the same time severely distorted. Consequently, steps had to be taken at the same time to free the *reference voltage* from higher harmonics.

The limiter used to keep the amplitude satisfactorily constant is a balanced stage (double triode T_1 - T_2 , fig. 1a) with a high cathode resistance R_k (e.g. 0.1 M Ω) in the common cathode lead (known as a "long-tailed pair"). T_1 and T_2 are biased in such a way that half the total cathode current I_k flows through each of them. A slight difference V_d (a few volts) between the voltages on the grids is sufficient to cause the total current I_k to flow to one or the other anode; this is illustrated in fig. 1b, where the two anode currents are plotted as a function of the difference $V_{g1} - V_{g2}$ between the grid voltages. If we superimpose on the grid of T_1 an alternating voltage v_i (fig. 1c) with an amplitude several times that of V_d , then T_1 and T_2 will pass the current I_k alternately. On the anode of T_2 which has a resistance R_a in series with it, a voltage will then appear with a more or less square waveform. The amplitude of the square wave voltage is $I_k R_a$, and is thus independent of the amplitude of v_i . By stabilizing the supply voltages and using

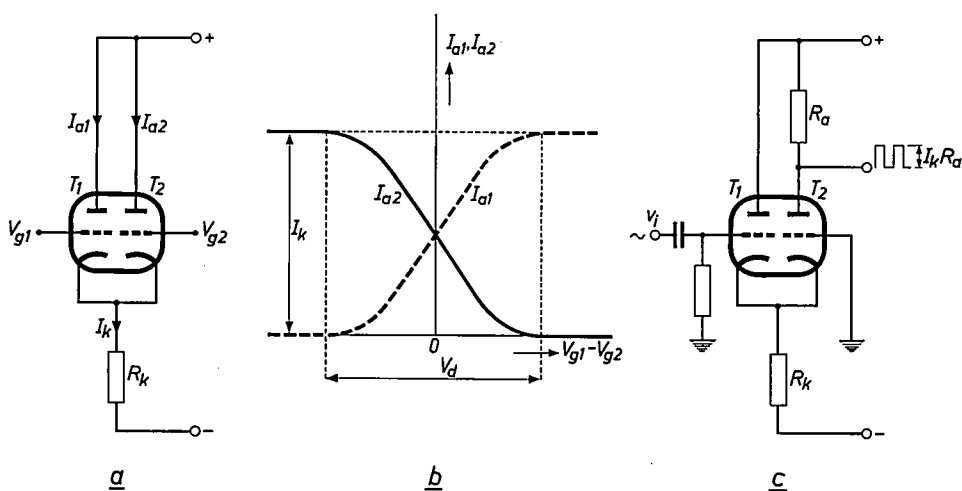


Fig. 1. a) Balanced stage with high resistance — here an ordinary resistance R_k — in the common cathode lead ("long-tailed pair"). b) The anode currents I_{a1} and I_{a2} of the triodes T_1 and T_2 in (a) are plotted versus the difference $V_{g1} - V_{g2}$ of the grid voltages. The difference V_d needed in order for the whole cathode current $I_k = I_{a1} + I_{a2}$ to flow from one anode to the other is only a few volts. c) When an alternating voltage v_i with an amplitude several times that of V_d is applied to the grid of T_1 , a square-wave voltage with amplitude $\frac{1}{2} I_k R_a$ appears at the anode of T_2 , which has a resistance R_a in series with it.

negative feedback through a filter F_1 and an amplifier A_2 . The filter — to which we shall presently return — passes the higher harmonics almost without attenuation but does *not* pass the fundamental component. The higher harmonics therefore undergo strong negative feedback, while the filter blocks the feedback path for the fundamental component. The result is that the higher harmonics are considerably attenuated in proportion to the fundamental. The voltage at point r thus approximates fairly closely to a sine wave; its distortion is about 0.3% — a great deal less than the roughly 50% distortion present in the square-wave voltage at point q .

The distortion at r , however, is still much greater than the permissible maximum of 0.01 % for the reference voltage. For this reason the above procedure is repeated: point r is connected via a resistor R_2 to an amplifier A_3 , which has again negative feedback for the higher harmonics through a filter F_2 (identical with F_1) and an amplifier A_4 .

Since the output s of A_3 is connected to the input p , the whole circuit forms an oscillator. The condition for oscillation is now fulfilled: the loop gain is automatically equal to 1 and the frequency is that at which the total phase shift (of amplifier and filters together) is zero. The phase shift in the amplifiers is negligibly small, and that in the filters (assumed to be ideal) is zero at the frequency at which the transfer is zero (the zero frequency). The whole system therefore oscillates with the zero frequency.

Fig. 2b shows the complete circuit of the reference oscillator. An important feature is that in the "main line", between the points q and s , very little happens that can endanger the constancy of the amplitude of the sine-wave voltage at s . In the main line there are two amplifiers (A_1 and A_3) and two voltage dividers (R_1-T_4 , and R_2-T_8). As can be seen from fig. 2b, A_1 and A_3 are cathode followers (T_3 and T_6 respectively); their gain is therefore a little less than unity ($1-A^{-1}$, where $A \gg 1$) and very constant. The same applies to the voltage divisions; the division ratios are similarly governed by an expression of the form $1-A^{-1}$ with $A \gg 1$. The amplifiers A_2 and A_4 need to have a high gain for suppressing the higher harmonics. They are situated, however, in the negative feedback paths, which are blocked to the fundamental component by the filters F_1 and F_2 ; therefore their gain need not be particularly constant. It would have been much more difficult to obtain a constant output voltage using a perhaps more obvious circuit containing a filter in the main line which passed the fundamental and blocked the

higher harmonics. Again, to avoid induced interfering voltages in coils having a high inductance, such a filter would have to be an RC and not an LC type. The low selectivity of such a filter would have to be improved by considerably amplifying the fundamental component in the main line. It would then have been difficult to get the required amplitude constancy.

In fig. 2b it is seen that the amplifier A_1 consists of a single cathode follower (triode T_3) but that A_3 has a more complicated circuit. The reason lies in the magnitude of the distortion introduced by the valves as a result of the curvature of their characteristic. For A_1 , where the signal — as stated — still shows a distortion of about 0.3%, the distortion introduced by a single cathode follower is relatively insignificant. For A_3 the distortion in the output voltage is required to be better than 0.01%. This makes it necessary to take careful account of the distortion which A_3 itself produces. This distortion is smaller the higher is the impedance in the cathode lead; a limit is set to this impedance, however, by the input impedance of the filter F_2 , which is rather crucial if the low value resistances in this filter are to be metallic resistors. Taking for these the largest values available we find that the distortion (mainly second harmonic) introduced by a single cathode follower at an output voltage of 10 V is such that the distortion in the output voltage comes very close to the specified limit of 0.01%. A single cathode follower might therefore have been sufficient. It was decided, however, that a wider margin was desirable for the reference voltage. This was obtained by using a more elaborate circuit for A_3 , which produces much less second harmonic distortion than the single cathode follower.

As shown in fig. 2b, A_3 consists of a single cathode follower T_7 preceded by a balanced stage T_5-T_6 having a common cathode resistance. The second harmonic in the output voltage of this combination is only 2.5×10^{-3} % of the fundamental component; the other higher harmonics are even weaker.

Distortion introduced by the cathode followers

The distortion introduced by the curvature of a valve characteristic can be calculated by a method described elsewhere¹⁾. Using this method the distortion introduced by the cathode followers employed in this circuit will now be calculated.

The equation of the valve characteristic (anode current i_a as a function of the "total driving voltage" v) can be written as a power series:

$$-i_a = av + \beta v^2 + \gamma v^3 + \dots, \dots \quad (1)$$

¹⁾ J. Rodrigues de Miranda and J. J. Zaalberg van Zelst, New developments in output-transformerless amplifiers, J. Audio Engng. Soc. 6, 244-250, 1958.

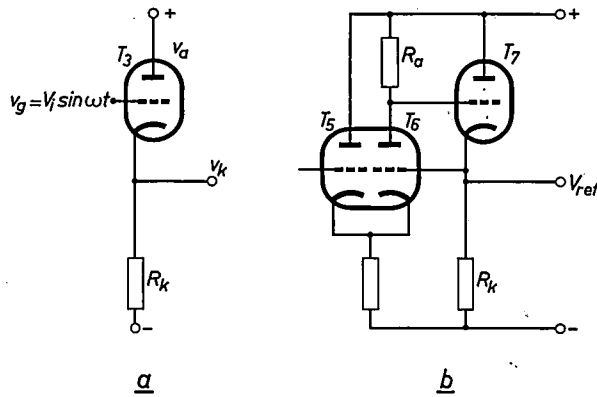


Fig. 3. a) Single cathode follower (like T_3 in fig. 2b): triode T_3 with cathode resistance R_k . b) complex cathode follower (like T_5 - T_6 - T_7 in fig. 2b). The inherent distortion here is much lower than in the single type.

where

$$v = v_g + \frac{v_a}{\mu} - \left(1 + \frac{1}{\mu}\right) v_k. \quad (2)$$

The voltages v_a , v_g and v_k are indicated in fig. 3a, and μ is the amplification factor of the valve.

We can express (1) in the form:

$$v = a i_a + b i_a^2 + c i_a^3 + \dots, \quad (3)$$

where

$$a = \frac{1}{\alpha}, \quad b = \frac{\beta}{\alpha^2}, \quad c = \frac{2\beta^2}{\alpha^5} - \frac{\gamma}{\alpha^4}. \quad (4)$$

For the single cathode follower (fig. 3a), $v_a = 0$ and $v_k = i_a R_k$. Assuming $(1 + \mu^{-1})R_k = r$, we find from (2) and (3) for the single cathode follower:

$$v_g = (a + r)i_a + b i_a^2 + c i_a^3 + \dots \quad (5)$$

If we expand i_a in a power series to v_g :

$$i_a = k_1 v_g + k_2 v_g^2 + k_3 v_g^3 + \dots, \quad (6)$$

the relation between (6) and (5) is the same as that between (1) and (3). By analogy with (4) we can therefore write:

$$a + r = \frac{1}{k_1}, \quad b = -\frac{k_2}{k_1^3}, \quad c = \frac{2k_2^2}{k_1^5} - \frac{k_3}{k_1^4}.$$

Combination with (4) gives:

$$k_1 = \frac{\alpha}{1 + ar}, \quad k_2 = \frac{\beta}{(1 + ar)^2}, \quad k_3 = \frac{-2\beta^2 r}{(1 + ar)^5} + \frac{\gamma}{(1 + ar)^4}, \text{ etc.}$$

If v_g is a purely sinusoidal voltage, $v_g = V_0 \cos \omega t$, this current i_a is given by the following series:

$$i_a = (k_1 V_0 + \frac{3}{2} k_3 V_0^3 + \dots) \cos \omega t + (\frac{1}{2} k_2 V_0^2 + \dots) \cos 2\omega t + (\frac{1}{4} k_3 V_0^3 + \dots) \cos 3\omega t + \dots$$

From this we find the ratio d_2 of the second harmonic to the

fundamental wave of the current i_a (and hence of the output voltage $v_k = i_a R_k$):

$$d_2 = \frac{2k_2 V_0}{4k_1 + 3k_3 V_0^2} = \left[\frac{2\alpha}{\beta V_0} (1 + ar)^2 - \frac{3\beta r V_0}{(1 + ar)^2} + \frac{3\gamma V_0}{\beta(1 + ar)} \right]^{-1}. \quad (7)$$

To determine which of the three terms between square brackets is the most important, we shall fill in some practical values. For the valve ECC 81 at 1 mA and 100 V anode voltage we have: $\alpha = \text{approx. } 1.5 \text{ mA/V}$, $\beta = \text{approx. } 0.80 \text{ mA/V}^2$ and $\gamma = \text{approx. } 1.6 \text{ mA/V}^3$. If r is 50 k Ω and $V_0 = 14 \text{ V}$, we find for the three terms respectively 1550, -0.3 and 1.1, so that in this case the first term is by far the most dominant. To a good approximation we can therefore simplify (7) to:

$$d_2 = \frac{\beta V_0}{2\alpha(1 + ar)^2}. \quad (8)$$

In this case, then, $d_2 = 1550^{-1} = 0.065\%$. The negative feedback further reduces this distortion by a factor which is difficult to calculate but turns out to be of the order of 10. The result is a distortion which remains just below the specified limit. As mentioned above, a certain margin was thought desirable.

From (8) we see that the distortion decreases if r , and thus the cathode resistance, R_k , is increased. R_k consists of the actual cathode resistance in parallel with the input impedance of the filter. The latter sets an upper limit to R_k .

The cathode follower preceded by a balanced stage, as used for A_3 , is represented in fig. 3b (omitting the elements for biasing T_6). Calculation shows that, with good dimensioning and minor simplifying assumptions, the inherent distortion given by (8) is reduced by a factor

$$\left(\frac{1}{\mu'} + \frac{2}{SR_a} \right)^{-1} \gg 1. \quad (9)$$

In this expression μ' is the amplification factor and S the transconductance of the triode T_6 in fig. 3b. With the adjustment used here, this valve ($\frac{1}{2}$ ECC 81) gives $\mu' = 60$ and $S = 2 \text{ mA/V}$. With $R_a = 0.22 \text{ M}\Omega$, the factor given by (9), with which the inherent distortion is reduced, is roughly 50.

The filters

For the same reasons that prompted us not to use an LC oscillator, we also avoided using coils in the filters (F_1 and F_2 in fig. 2) which therefore consists entirely of resistors and capacitors.

As we have seen, the filters block the fundamental component but pass the higher harmonics. A filter network that possesses this characteristic is the twin-T type, composed of resistors and capacitors (fig. 4a). It is so called because it can be regarded

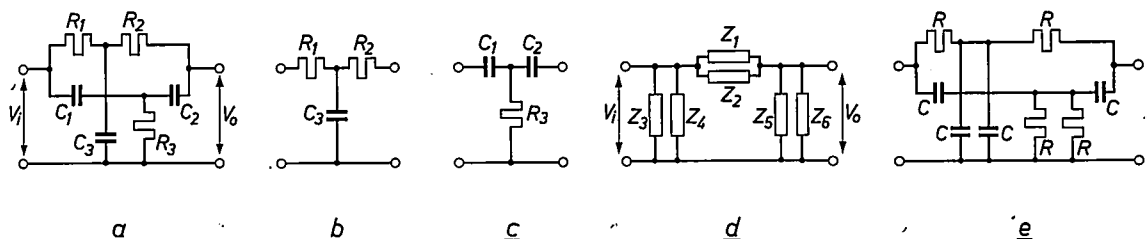


Fig. 4. a) Twin-T filter, consisting of resistors and capacitors. b) and c) The two T sections from which (a) is formed. d) The two T sections of (a) are transformed into II sections. e) In the case of symmetry the network (a) can be formed from four identical resistors R and four identical capacitors C .

as two T sections "one on top of the other" (fig. 4b and c). At one frequency (the zero frequency f_0) the filter passes no signal if the resistance and capacitance values satisfy the following condition:

$$\frac{R_1 R_2}{R_1 + R_2} C_3 = R_3 (C_1 + C_2). \quad (10)$$

In that case

$$f_0 = \frac{1}{2\pi \sqrt{(R_1 + R_2) R_3 C_1 C_2}}.$$

To derive the condition (10) we transform the two T sections (or star networks) of fig. 4b and c into π sections (fig. 4d). The six impedances $Z_1 \dots Z_6$ are easily expressed in terms of $R_1, R_2, R_3, C_1, C_2, C_3$ and the angular frequency ω . The ratio $1/A_F$ of the input to the output voltage of the filter we read from fig. 4d:

$$\frac{1}{A_F} = \frac{V_i}{V_o} = 1 + \frac{Z_1 Z_2 (Z_5 + Z_6)}{(Z_1 + Z_2) Z_5 Z_6}.$$

From this formula it is seen that the only case in which the filter blocks the signal completely ($A_F = 0, V_i/V_o = \infty$) occurs when $Z_1 + Z_2$ is zero. Expressing Z_1 and Z_2 in terms of the resistances and of the capacitances, and putting both the real and imaginary parts of $Z_1 + Z_2$ equal to zero we obtain the following equations:

$$\omega^2 = \frac{1}{(R_1 + R_2) R_3 C_1 C_2} \quad (11)$$

and

$$\omega^2 = \frac{C_1 + C_2}{R_1 R_2 C_1 C_2 C_3} \quad (12)$$

At the frequency f_0 the filter passes no signal if ω in the left-hand members of (11) and (12) is equal to $2\pi f_0$. The right-hand members are then likewise identical, which leads to equation (10).

For filters that exactly fulfil the condition (10) it can be deduced that the transmission-ratio A_F as a function of frequency is given by:

$$A_F = \frac{V_o}{V_i} = \frac{jQ\beta}{1 + jQ\beta}, \quad (13)$$

where β is the relative detuning:

$$\beta = \frac{f}{f_0} - \frac{f_0}{f},$$

and the figure of merit Q is:

$$Q = \frac{\sqrt{(R_1 + R_2) R_3 C_1 C_2}}{R_1 C_3 + (R_1 + R_2) C_2}.$$

If the filter is symmetrical ($R_1 = R_2$, say $= R$, and $C_1 = C_2$, say $= C$; the symmetry also implies: $R_3 = \frac{1}{2}R$ and $C_3 = 2C$), we find $Q = \frac{1}{4}$, so that this kind of filter cannot be expected to have a very great selectivity. With an asymmetric filter a somewhat higher Q can be obtained, theoretically a maximum of $\frac{1}{2}$. The difference is so small as to be

unimportant compared with the practical advantages of a symmetrical filter, whose input and output impedances are more favourable and which can be built with four identical resistors R , and four identical capacitors C (fig. 4e).

Small deviations in the values of C and R cause deviations of the same order of magnitude in the oscillator frequency and voltage amplitude. For this reason, only mica capacitors and metallic resistors are used in the filters.

Instead of the behaviour of the filter itself we shall now consider the behaviour of the filter together with the amplifiers with which it works, i.e. the combination A_1 - F_1 - A_2 in fig. 2a. This active filter, shown separately in fig. 5, has an input voltage V_q and an output voltage V_r (cf. points q and r in fig. 2a). The transmission A_{act} of this active filter can be found with the aid of (13):

$$A_{act} = \frac{V_r}{V_q} = B \frac{1 + jQ\beta}{1 + j(A_2 + 1)Q\beta}. \quad (14)$$

Here A_2 is the gain of the pentode amplifier A_2 , and B a factor, not otherwise relevant to our considerations, which is independent of frequency and differs little from unity. It follows from (14) that

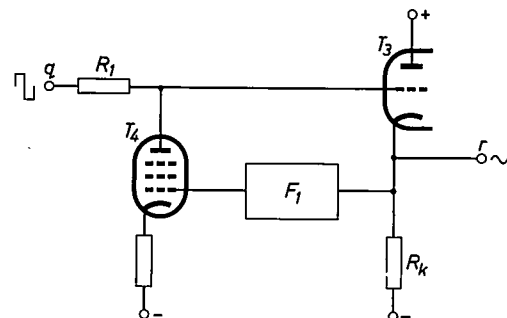


Fig. 5. The "passive" filter F_1 and the amplifiers A_1 and A_2 (fig. 2a) together form an active filter, with input terminal q and output terminal r . The fundamental component of the square-wave voltage at q is passed. The higher harmonics are considerably attenuated owing to the high gain of A_2 .

for the fundamental component ($\beta = 0$) the transmission A_{act} is equal to B (so that the fundamental is passed virtually unattenuated) and that the higher harmonics are better suppressed the higher is the gain A_2 . Given $A_2 \gg 1$ and $Q = \frac{1}{4}$, the absolute value of A_{act}/B at the frequencies $f_0, 2f_0$, etc. is found from (14):

$f = f_0$	$2f_0$	$3f_0$	$4f_0$	\dots	∞
$\beta = 0$	$1\frac{1}{2}$	$2\frac{2}{3}$	$3\frac{3}{4}$	\dots	∞
$ A_{act}/B = 1$	$\frac{2.85}{A_2}$	$\frac{1.80}{A_2}$	$\frac{1.46}{A_2}$	\dots	$\frac{1}{A_2}$

It can be seen that the second harmonic is

suppressed less than the third and higher harmonics. On the other hand, the second harmonic is only weakly represented in the voltage V_q (a perfect square-wave voltage contains no even harmonics at all).

Measurements of the reference voltage

Measurements of the output voltage of the reference oscillator (r.m.s. value about 10 V, frequency 80 c/s) have been made to determine the distortion and the constancy of the amplitude and frequency.

The following values were found for the contribution of the principal higher harmonics to the distortion:

second harmonic	d_2	$= 2.5 \times 10^{-5}$,
third harmonic	d_3	$= 1.5 \times 10^{-5}$,
remaining harmonics	d_{rest}	$< 0.4 \times 10^{-5}$.

To keep the amplitude from varying by more than 0.01%, the supply voltage of + 300 V had to be kept constant to within 0.5 V, the supply voltage of - 200 V to within 0.1 V, and the heater voltage (nominal 6.3 V) to within 0.3 V. These conditions can easily be met.

Although nothing was specified regarding the constancy of the frequency, this too was investigated. In ten independent measurements we counted the number of times the voltage passed through zero in three minutes, and the frequency deviations were found to be no more than 0.02% of the average frequency.

The output stage

The circuit diagram of the output stage and the associated driving stage is shown in fig. 6.

The output stage contains two pentodes (T_{17} and T_{18}) in a single-ended push-pull arrangement²⁾. In common with the ordinary push-pull circuit, this arrangement suppresses the formation of even harmonics, but it has the additional advantage that the direct current does not pass through the load, thus making an output transformer unnecessary. The output pentodes are of the EL 86 type, specially designed for use in single-ended push-pull circuits (high anode current at a relatively low anode voltage). In the output current these valves give rise to distortion of only a few per cent, i.e. an order of magnitude smaller than the distortion caused by the non-linear load itself.

The internal resistance of the output stage is roughly 1000 ohms. As noted at the beginning of this article, the requirement as to the maximum

distortion permissible in the output voltage amounts to specifying that the internal resistance should not exceed about 0.6 ohm. One of the functions of the circuit which drives the output stage is therefore to ensure that the loop gain is a few times 10^3 . For this reason the driving circuit consists of two stages.

The primary function of the driving circuit is to act as a control system, i.e. to drive the output stage in such a way that a variable part kU_0 of the output voltage is kept equal to the reference voltage V_{ref} of 10 V; the difference $kU_0 - V_{\text{ref}}$ is amplified and is used to drive the output stage. The fraction k can be adjusted from 1/6 to 1/4 using a potentiometer P , so that U_0 is continuously variable from 60 to 40 V.

An important question is the *distortion in the driving circuit*. Particularly favourable in this respect are *difference amplifiers*³⁾, i.e. balanced amplifiers having a high resistance in the common cathode lead.

Minimizing the distortion makes particularly severe demands on the first stage of the driving circuit. This stage has at its input an "in-phase voltage" of 10 V; the "anti-phase voltage" is the much smaller difference $kU_0 - V_{\text{ref}}$. To keep the distortion minimum in spite of this relatively high in-phase voltage, it is necessary to minimize the current variations which the in-phase voltage causes in the valves.

This is precisely what a good difference amplifier does, hence the fact that the first stage is a very carefully designed difference amplifier, possessing both a high rejection factor and a high discrimination factor.

To limit the distortion of this stage to 0.01%, the rejection factor should be of the order of 10^4 . This follows from a calculation similar to that given above under the heading *Distortion introduced by the cathode followers*.

The first stage thus consists of the cascodes T_9-T_{11} and $T_{10}-T_{12}$ in a push-pull arrangement; the common cathode lead contains the very high differential resistance of the cascode $T_{13}-T_{14}$ ⁴⁾.

The second stage is a simple difference amplifier, consisting of a double triode $T_{15}-T_{16}$ in a balanced arrangement with an ordinary resistance in the cathode lead.

²⁾ See e.g. J. Rodrigues de Miranda, Philips tech. Rev. 19, 2, 1957/58.

³⁾ In the following section some terms from the technique of difference amplifiers will be used. For an explanation of the terms see G. Klein and J. J. Zaalberg van Zelst, General considerations on difference amplifiers, Philips tech. Rev. 22, 345-351, 1960/61.

⁴⁾ This arrangement is a combination of the circuits shown in fig. 6 and fig. 9 in the article by G. Klein and J. J. Zaalberg van Zelst, Circuits for difference amplifiers, I, Philips tech. Rev. 23, 142-150, 1961/62.

For each of the higher harmonics the internal resistance R_i can be determined by dividing the relevant voltage component by the corresponding current component. For the second to fifth harmonics the results obtained, as the averages of measure-

ments with three values of R , varied from 0.46 to 0.66 ohm.

The lower limit of the distortion in the output voltage is determined chiefly by the distortion already present in the reference voltage (see above); the value found was 3×10^{-5} . The relative amplitude and frequency variations in the output voltage are identical with those in the reference voltage.

The oscillogram in *fig. 8* shows the peaks of the output voltage recorded in a time of roughly 1

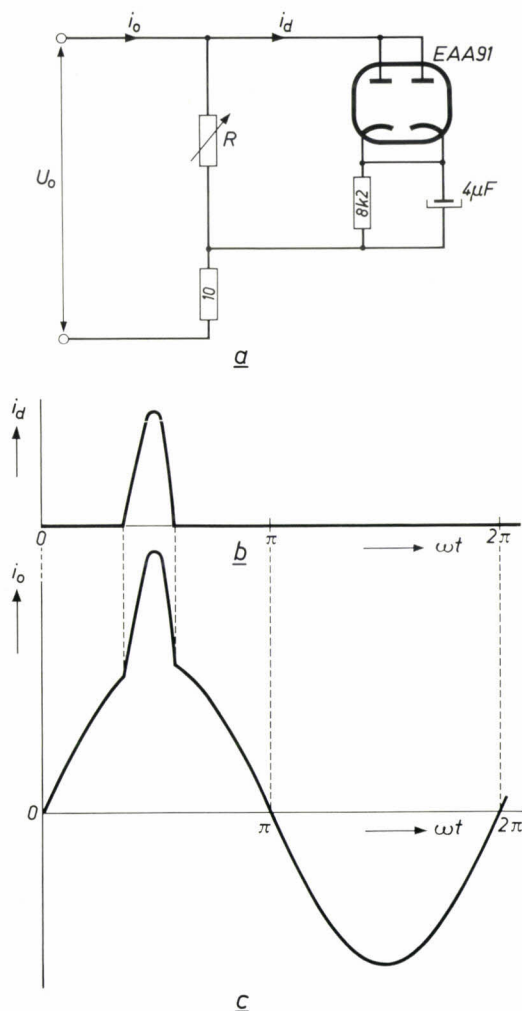


Fig. 7. a) Circuit used as non-linear load for distortion measurements. The load current i_o consists of a linear component, which flows through the variable resistor R , and a non-linear component i_d , which, during a small part of each period, charges up a capacitor of $4 \mu\text{F}$ through the diode EAA 91. The harmonics of i_o were found by analysing the voltage across the 10 ohms series resistance. b) The charging current i_d . c) The total current i_o .

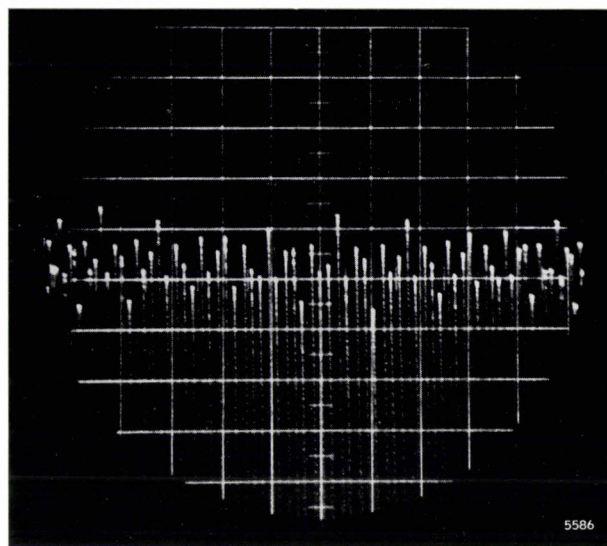


Fig. 8. Oscillogram obtained with a "voltage microscope"⁵⁾, showing the peaks of the output voltage (amplitude approx. 70V). The height of each square corresponds to about 14 mV (the zero line of the sine wave should therefore be imagined at about 35 metres below the peaks!). The maximum fluctuations in amplitude are about 15 mV (0.02%). The width of the oscillogram corresponds roughly to one second.

second⁵⁾. The amplitude of the voltage was about 70 V. The fluctuation is seen to be approximately 15 mV, or 0.02%, which is well below the specified limit of 0.1%.

⁵⁾ G. Klein and J. J. Zaalberg van Zelst, Philips tech. Rev. **23**, 173 ff., 1961/62.

Summary. For measurements on magnetic amplifiers an oscillator was needed that could deliver 2 W at a voltage of 50 V, 80 c/s, a special requirement being that the voltage under severe non-linear loading should show no more than 0.01% distortion and fluctuate in amplitude by no more than 0.1%. This requirement means keeping the internal resistance extremely low (no higher than about 0.6 ohm).

In the solution adopted a division is made into an oscillating section and an output stage incorporating a control system. The oscillating section (the reference oscillator) delivers an almost purely sinusoidal and constant voltage of 10 V, 80 c/s, which is used as the reference voltage for the control system.

The reference oscillator begins by generating a very constant square-wave voltage of 80 c/s, the higher harmonics of which are suppressed by negative feedback via double-T section RC filters, which pass the higher harmonics but not the fundamental component. The result is a reference voltage with no more than 0.003% distortion and 0.02% fluctuation.

The control system of the output stage compares an adjustable fraction ($1/6-1/4$) of the output voltage with the reference voltage and drives the output stage (two EL 86 pentodes in a single-ended push-pull arrangement). The loop gain is so high that the internal resistance of the output stage is reduced to the required low value of about 0.6 ohm. Measures are taken to minimize the distortion in the amplifying stages of the control system.

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¹⁾ Beginning with this volume an abstract of each publication will no longer be given but only the title, with occasional references to other related publications.

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