THREE-PHASE R-C OSCILLATOR FOR RADIO AND AUDIO FREQUENCIES

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(Plate IX)

ABSTRACT. The R-C tuned type of oscillator has received close attention in recent years because of its stability and purity of wave form. The usual procedure is to provide regenerative feedback from the output to the input of an R-C coupled amplifier which may consist of one or more stages. The oscillations are generally limited to the audio frequency range because the choic of components required for producing radio frequencies presents practical difficulties. These have been overcome in the present three-phase oscillator consisting of three identical stages. This oscillator has been used to produce three-phase audio as well as radio frequency oscillations Mathematical formulae have been deduced which are corroborated by actual experiments.

INTRODUCTION

It is well known that an amplifier having a feedback path between the output and the input is liable to produce self-oscillations when the feedback is positive and the overall gain of the amplifier and feedback path is not less than unity. The frequency of the maintained oscillations is that at which the overall phase shift round the loop path is zero. The oscillator may consist of one or more stages of amplification. If a single valve is used and the desired feedback obtained by means of phase-shifting ladder networks, then the unavoidable attenuation in such networks necessitates high amplification and greater excursion of anode voltage with consequent distortion. This can be avoided by using more than one stage of amplification. A three-phase system, due originally to M. Van der and B. Van der Pol(1934), which has been in vogue during recent years, is seen to behave like a selective tuned circuit as the overall phase difference is extremely sensitive to frequency. This is why the waveform of oscillation is nearly sinusoidal. The oscillations are generally limited to the audio frequency range because the choice of components required for producing radio frequencies presents practical difficulties.

In a recent communication by the authors (Rakshit and Bhattacharyya, 1946) it was shown that the conventional circuit of the three-phase oscillator with components selected for producing audio frequency oscillations invariably generates high radio frequencies by virtue of the unavoidable stray and inter-electrode capacities. Audio frequency can be generated by such a system only after making some modifications of the simple circuit. The present paper gives the details of the arrangement.

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THEORETICAL CONSIDERATIONS

The principle of the maintenance of oscillation is to provide the usual regenerative feedback from the output to the input of a three stage R-C coupled amplifier. Fig. 1 shows the diagram of any of the three symmetrical stages

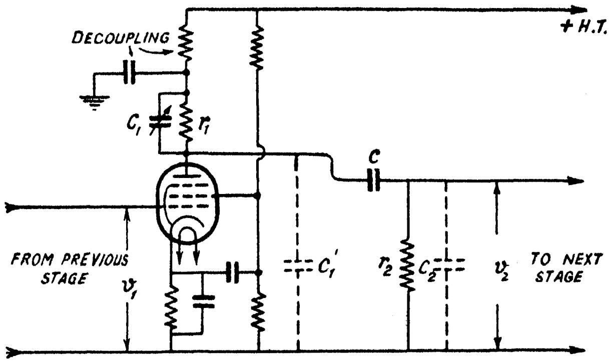


FIG. 1

constituting the feedback chain where pentodes are used to minimise interelectrode capacities and special care is taken to ensure minimum change of electrode voltages due to mains voltage fluctuations.

Assuming a perfect three-fold symmetry let us denote

r_1 = plate load,

r_p = a.c. plate resistance of the valve,

C_1 = shunting condenser across the load,

C = coupling condenser,

r_2 = grid-leak resistance,

C_1' = plate to cathode capacity together with stray wiring capacity to the left of the coupling condenser C,

C_2 = input grid to cathode capacity together with stray wiring capacity to the right of the coupling condenser C, and

g = mutual conductance of the valve.

The exact equivalent circuit of the above is given in Fig. 2(*a*) which reduces to Fig. 2(*b*), assuming $r_1 << r_p$ which is very high for pentodes.

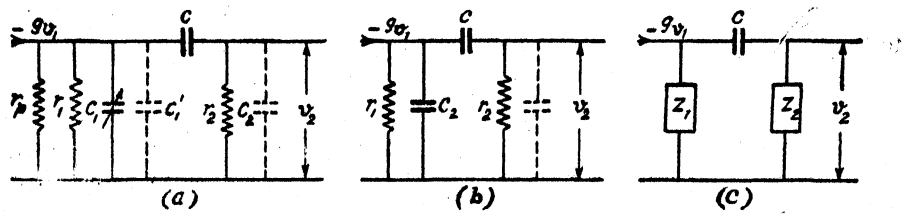


FIG. 2

Fig. 2(*b*) is again equivalent to Fig. 2(*c*) by putting Z_1 for $\dfrac{r_1}{1+j\omega C_1 r_1}$ and

Z_2 for $\dfrac{r_2}{1+j\omega C_2 r_2}$ where ω = angular frequency of the oscillations maintained and $C_t = C_1 + C_1'$.

The alternating current along the branch Z_2 is given by

$$i_{z_2} = gv_1 \frac{Z_1}{Z_1 + Z_2 + 1/j\omega C}$$

Therefore the output voltage which is equal to the input voltage at the next valve is given by

$$v_2 = Z_2 . \, i_{z_2} = -gv_1 \frac{Z_1 Z_2}{Z_1 + Z_2 + 1/j\omega C}$$

On substituting the values of Z_1 and Z_2,

$$v_2 = -\frac{gv_1 r_1 r_2 . \, j\omega C}{j\omega C r_1(1+j\omega C_2 r_2) + j\omega C r_2(1+j\omega C_t r_1) + (1+j\omega C_2 r_2)(1+j\omega C_t r_1)}$$

$$= \frac{-gv_1 r_1 r_2 C}{\Sigma Cr - \frac{j}{\omega}(1-\omega^2 r_1 r_2 \Sigma CC)}$$

where $\Sigma Cr = Cr_1 + Cr_2 + C_t r_1 + C_2 r_2$ and $\Sigma CC = CC_2 + CC_t + C_2 C_t$

Or,
$$v_2 = \frac{gv_1 r_1 r_2 C}{\left[(\Sigma Cr)^2 + \left(\frac{1-\omega^2 r_1 r_2 \Sigma CC}{\omega}\right)^2\right]^{\frac{1}{2}}} . \, \epsilon^{\pi+\theta} \quad \text{... (1a)}$$

$$= |Av_1| . \, /\pi + \theta \quad \text{... (1b)}$$

where
$$\theta = \tan^{-1} \frac{1-\omega^2 r_1 r_2 \Sigma CC}{\omega \Sigma Cr} \quad \text{... (1c)}$$

Hence we see that each stage considered as a separate amplifier produces an amplification given by

$$A = gr_1 r_2 C \Bigg/ \left[(\Sigma Cr)^2 + \left(\frac{1-\omega^2 r_1 r_2 \Sigma CC}{\omega}\right)^2\right]^{\frac{1}{2}} \quad \text{... (2a)}$$

and a phase shift with respect to its input voltage given by

$$\phi = \pi + \theta \quad \text{... (2b)}$$

Now considering the system as a whole it is clear that for the maintenance of oscillation we must have

$$\left.\begin{array}{l}(i)\ A \nless 1 \\ \text{and } (ii)\ 3\phi = 2n\pi\end{array}\right\} \quad \text{... (3)}$$

where n is any integer.

According to condition (ii) the possible values of ϕ between 0 and 2π are $\frac{2\pi}{3}$, $\frac{4\pi}{3}$ and 2π. Of these three values the third one, *viz.*, $\phi = 2\pi$, though mathematically correct, is, however, physically impossible in our system and hence we are left with two alternative cases— $\phi = \frac{2\pi}{3}$ and $\phi = \frac{4\pi}{3}$.

Case I.—High Frequency Oscillations.

For the first case, $\phi=\frac{2\pi}{3}$, *i.e.*, $\theta=-\frac{\pi}{3}$. Writing ω_1 for ω we have from equation (1*c*)

$$\frac{1-\omega_1{}^2 r_1 r_2 \Sigma CC}{\omega_1 \Sigma Cr}=\tan\theta=-\sqrt{3} \quad \ldots \quad (4a)$$

Or,
$$\omega_1{}^2 r_1 r_2 \Sigma CC-\sqrt{3}\,\omega_1 \Sigma Cr-1=0 \quad \ldots \quad (4b)$$

i.e.,
$$\omega_1=\frac{\sqrt{3}\Sigma Cr \pm \sqrt{3(\Sigma Cr)^2+4r_1 r_2 \Sigma CC}}{2r_1 r_2 \Sigma CC}$$

Neglecting the negative value of ω_1 and assuming $r_2 \gg r_1$ and $C_2 \ll C$, so that $4r_1 r_2 \Sigma CC$ is neglected in comparison with $3(\Sigma Cr)^2$, we have,

$$\omega_1=\frac{2\sqrt{3}\Sigma Cr}{2r_1 r_2 \Sigma CC}=\frac{\sqrt{3}(Cr_1+Cr_2+C_t r_1+C_2 r_2)}{r_1 r_2 (CC_2+CC_t+C_2 C_t)}$$

$$\doteq \frac{\sqrt{3}(C+C_2)}{r_1 C(C_2+C_t)} \quad \ldots \quad (4c)$$

Or with more approximation

$$\omega_1=\frac{\sqrt{3}}{r_1(C_2+C_t)}=\frac{\sqrt{3}}{r_1(C_1+C_1'+C_2)}=\frac{\sqrt{3}}{r_1(C_1+C_s)}$$

where $C_s=C_1'+C_2$, *i.e.*, C_s denotes the total stray capacity on either side of the coupling condenser C.

$$\therefore \quad f_1=\frac{\sqrt{3}}{2\pi r_1(C_1+C_s)} \quad \ldots \quad (5a)$$

Applying eqn. (4a) to determine the value of the amplification "A" in eqn. (2a) we get

$$A=\frac{g r_1 r_2 C}{2\Sigma Cr}=\frac{g r_1}{2}\cdot\frac{1}{\left[\frac{r_1}{r_2}+1+\frac{C_t}{C}\cdot\frac{r_1}{r_2}+\frac{C_2}{C}\right]}$$

Hence condition (ii) for the maintenance of oscillation as given by eqn. (3) becomes

$$g r_1 \nless 2\left[1+\frac{r_1}{r_2}+\frac{C_t}{C}\cdot\frac{r_1}{r_2}+\frac{C_2}{C}\right] \quad \ldots \quad (5b)$$

Case II.—Low Frequency Oscillations.

For the second case, $\phi=\frac{4\pi}{3}$, *i.e.*, $\theta=\frac{\pi}{3}$. Writing ω_2 for ω we get from equation (1c)

$$\frac{1-\omega_2^2 r_1 r_2 \Sigma CC}{\omega_2 \Sigma Cr}=\tan\theta=\sqrt{3} \quad \ldots \quad (6a)$$

Or,

$$\omega_2^2 r_1 r_2 \Sigma CC + \sqrt{3}\,\omega_2 \Sigma Cr - 1 = 0 \qquad \ldots \quad (6b)$$

i.e.,

$$\omega_2 = \frac{-\sqrt{3}\Sigma Cr \pm \sqrt{3}\Sigma Cr\left[1 + \dfrac{4r_1r_2\Sigma CC}{3(\Sigma Cr)^2}\right]^{\frac{1}{2}}}{2r_1 r_2 \Sigma CC}$$

Neglecting, as before, the negative value of ω_2 and taking $r_2 \gg r_1$ so that $\dfrac{4r_1r_2\Sigma CC}{3(\Sigma Cr)^2} \ll 1$, we have by expanding the square root by the Binomial Theorem and retaining the first two terms

$$\omega_2 = \frac{-\sqrt{3}\Sigma Cr + \sqrt{3}\Sigma Cr\left[1 + \dfrac{2r_1r_2\Sigma CC}{3(\Sigma Cr)^2}\right]}{2r_1 r_2 \Sigma CC}$$

$$= \frac{1}{\sqrt{3}\Sigma Cr} \doteq \frac{1}{\sqrt{3}r_2(C + C_2)} \qquad \ldots \quad (6c)$$

Or with further approximation, when $c \gg c_2$,

$$\omega_2 = \frac{1}{\sqrt{3}Cr_2}, \qquad i.e., \qquad f_2 = \frac{1}{2\pi\sqrt{3}Cr_2} \qquad \ldots \quad (7a)$$

Here also condition 3(*ii*) for the maintenance of oscillation becomes

$$gr_1 \nless 2\left[1 + \frac{r_1}{r_2} + \frac{C_1}{C} \cdot \frac{r_1}{r_2} + \frac{C_2}{C}\right] \qquad \ldots \quad (7b)$$

Vector Relations

The relations between the currents and voltages in the different branches of the circuit are depicted vectorially in Figs. 3 and 4 where the suffixes in i and v refer to the currents and voltages for the respective branches. Figs.

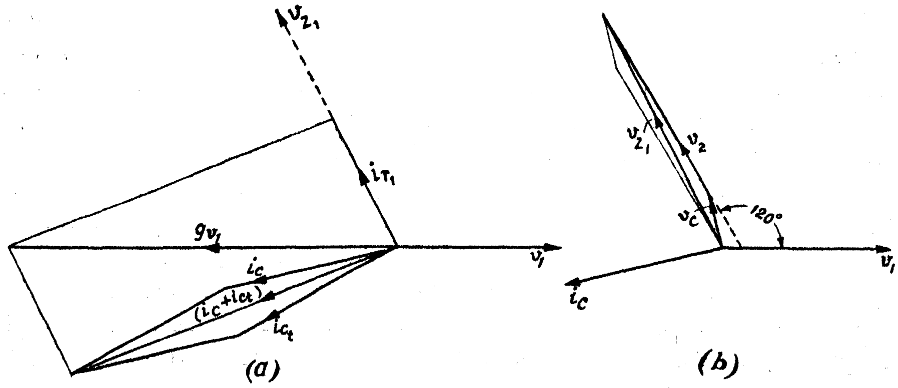


FIG. 3

3(*a*) and (*b*) represent the relations for Case I and Figs. 4(*a*) and (*b*) for Case II. From Figs. 3(*b*) and 4(*b*) it can easily be seen that the phase advance

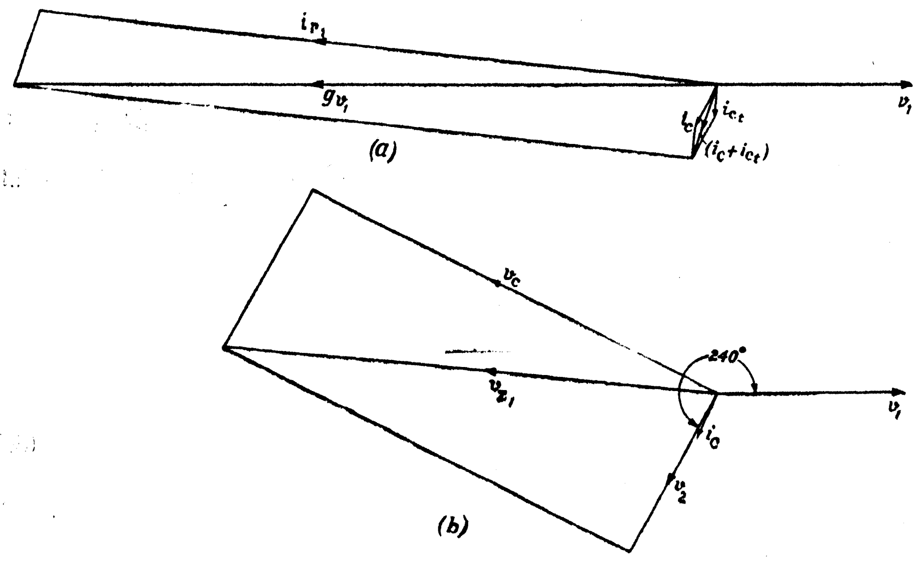


FIG. 4

per stage at f_1 and f_2, corresponding to Cases I and II, are 120° and 240° respectively, so that the feedback in both the cases is positive.

Simultaneous R.F. and L.F. Oscillations

From the above analyses we see that as the gain is the same for the two cases there is probability of the two frequencies f_1 and f_2 being maintained simultaneously. Now eqns. (5*a*) and (7*a*) show that under average working conditions f_1 can be easily made to lie in the radio frequency range and f_2 generally limited to the audio frequency range. It may, however, be mentioned in this connection that this probability of more than one oscillation being simultaneously maintained is a general feature of multiphase oscillators. For example, in a 7-phase oscillator two sets of oscillations are possible, each having two frequencies—one in the radio and the other in the audio frequency range. Thus for one set the phase advances per stage are $154\frac{2}{7}°$ and $205\frac{5}{7}°$, while the corresponding values for the other set are $102\frac{6}{7}°$ and $257\frac{1}{7}°$. It can be shown that this latter set of oscillations requires greater gain and can thus be easily suppressed. In the present three-phase oscillator, although the two modes of oscillation are equally probable, since the required gain is the same in either case, experiments have shown that under the above arrangement the radio frequency f_1 only is maintained to the exclusion of the audio frequency f_2. This is due to the fact that in a system where oscillation grows up from an initial impulse, having frequency components from zero to infinity, the highest probable frequency will build up quickly and so in the present case, as the building up of oscillation at f_1 is much quicker than that at f_2, f_1 grows up first and once f_1 is built up it is not generally possible for f_2 to grow thereafter. Under certain favourable conditions, however, it is possible to have both radio and audio frequency oscillations simultaneously maintained. This is discussed later.

Production of audio frequency oscillations

From the above considerations it is obvious that to obtain f_2 we must somehow suppress f_1 without of course adversely affecting f_2. This can be done by connecting a capacitance from the anode to cathode of only one stage. The magnitude of this capacity should be such that it will effectively reduce the overall gain of the system below unity at the radio frequency but its reactance at the audio frequency should be large compared with the load resistance r_1.

It will be seen from the discussion of harmonic distortion that oscillations are purest when the three stages are identically the same and the overall gain is just greater than unity. Any asymmetry in the stages requires greater load resistances for the production of oscillations. In actual practice it is more difficult to maintain symmetry for f_2 than for f_1. An attempt to suppress f_1 by connecting a suitable capacitance as explained above with a view to build up the audio frequency introduces further asymmetry, specially for the highest audio frequencies. Hence if initially the radio frequency oscillation be going on at critical maintenance condition, the use of such a capacitance will no doubt suppress f_1, but f_2 may not actually be maintained in that condition. In such a case it will be necessary to increase the load resistance and then only f_2 will build up.

The double function of suppressing f_1 and increasing the gain without at the same time introducing any asymmetry, can be effected simultaneously by incorporating in each stage a small resistance r in series with r_1 and forming part of the total load, but left unshunted by C_1, as shown in Fig. 5(*a*). Fig. 5(*b*) shows the simplified equivalent circuit at radio frequency. Now if C_1

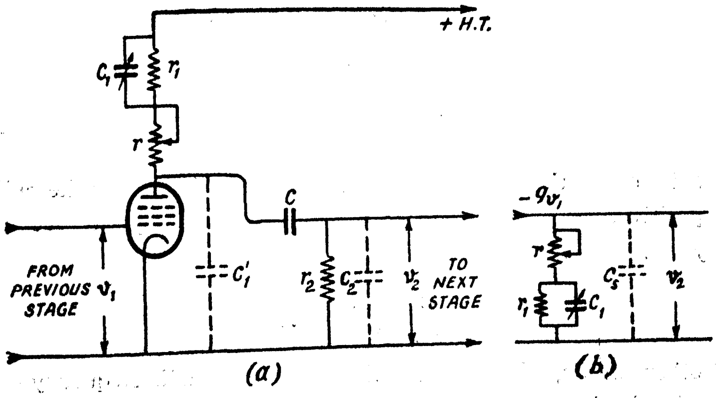


FIG. 5

be large compared with C_s but not large enough to have any appreciable shunting influence at the highest audio frequency we can, as a first approximation, neglect C_s in our equivalent circuit. This will not vitiate our inference of the effect of r at radio frequency but will on the other hand simplify the analysis.

The output voltage is then given by

$$v_2 = -gv_1\left[r + \frac{r_1}{1+j\omega C_1 r_1}\right] = \frac{gv_1}{1+\omega^2 C_1^2 r_1^2}[(R+\omega^2 C_1^2 r_1^2 r)^2 + (\omega C_1 r_1^2)^2]^{\frac{1}{2}} \cdot e^{\pi+\theta} \quad \ldots \quad (8)$$

where $\tan\theta = -\dfrac{\omega C_1 r_1^2}{R+\omega^2 C_1^2 r_1^2 r}$ and $R = r_1 + r$.

Or, $$v_2 = |Av_1| \cdot \underline{/\pi+\theta} \quad \ldots \quad (9a)$$

where the amplification A is given by

$$A = \frac{g}{1+\omega^2 C_1^2 r_1^2}\left[(R+\omega^2 C_1^2 r_1^2 r)^2 + (\omega C_1 r_1^2)^2\right]^{\frac{1}{2}} \quad \ldots \quad (9b)$$

As before the conditions for maintenance of oscillation are (i) $A \not< 1$ and (ii) $3(\pi+\theta) = 2n\pi$. For the radio frequency mode under consideration, $\theta = -\dfrac{\pi}{3}$ and hence condition (ii) is satisfied if

$$-\frac{\omega C_1 r_1^2}{R+\omega^2 C_1^2 r_1^2 r} = \tan\theta = -\sqrt{3} \quad \ldots \quad (9c)$$

This gives
$$\omega = \frac{C_1 r_1^2 \pm \sqrt{C_1^2 r_1^4 - 12 C_1^2 r_1^2 r R}}{2\sqrt{3} C_1^2 r_1^2 r}$$

Or, neglecting the positive sign before the radical which is obviously inadmissible, if we consider the limit when $r = 0$,

$$\omega = \frac{C_1 r_1^2 - C_1 r_1 \sqrt{r_1^2 - 12 r R}}{2\sqrt{3} C_1^2 r_1^2 r}$$

$$= \frac{r_1 - \sqrt{r_1^2 - 12 r R}}{2\sqrt{3} C_1 r_1 r} \quad \ldots \quad (9d)$$

Now in order that ω may be real, the expression under the radical sign must be positive. Remembering that $R = r_1 + r$, it is easy to see that the necessary condition is

$$r \not> \frac{r_1}{6+4\sqrt{3}} \not> 0.077 r_1 \quad \ldots \quad (10)$$

In other words, if we make $r > 0.077 r_1$, no radio frequency oscillation can be maintained. In practice the effect of C_s is to make the limiting value of r slightly greater than that given by equation (10).

Comparing equations (9b) and (9c) we have

$$A = \frac{2g(R+\omega^2 C_1^2 r_1^2 r)}{1+\omega^2 C_1^2 r_1^2}$$

$$= 2g\left(r + \frac{r_1}{1+\omega^2 C_1^2 r_1^2} \right) \quad \ldots \quad (11a)$$

On substitution from (9d), this further reduces to

$$A = \frac{2gr}{1 - \dfrac{6r}{r_1 - \sqrt{r_1^2 - 12rR}}} \quad \ldots \quad (11b)$$

Since the value of r required to just suppress the r.f. oscillations is a small fraction of r_1, as given by equation (10).

$$\sqrt{r_1^2 - 12rR} = r_1 - 6r\left[1 + 4\frac{r}{r_1} + 24\left(\frac{r}{r_1}\right)^2 + \text{higher powers of } \frac{r}{r_1}\right]$$

Hence, as a first approximation, for small values of r,

$$A = \frac{2gr}{4\dfrac{r}{r_1} + 8\left(\dfrac{r}{r_1}\right)^2} = \frac{gr_1}{2+4\dfrac{r}{r_1}} \quad \ldots \quad (11c)$$

Similarly the expression (9d) for frequency becomes

$$\omega = \frac{\sqrt{3}}{C_1 r_1}\left(1 + 4\frac{r}{r_1}\right) \quad \ldots \quad (11d)$$

Equations (11c) and (11d) show that as r is gradually increased from zero the frequency of the generated r. f. oscillations continuously increases whereas the gain of the system continuously decreases. The variations of $f = \frac{\omega}{2\pi}$ and A with r, for two different values of r_1, are shown in Table I.

These variations are shown plotted in Fig. 6. Fig. 6(*a*) refers to $r_1 = 3000\Omega$ and Fig. 6(*b*) to $r_1 = 1850\Omega$. The two sets of curves at once show that if r_1 is near the critical value required for maintenance then gain is reduced below unity for values of r less than that given by equation (10). If, therefore, the value of r_1 be such that the r.f. oscillations are just maintained and r is gradually increased, the limitation of gain is primarily responsible for the stoppage of r.f. oscillations. Fig. 6 further shows that when r_1 is sufficiently greater than the critical value, the limitation of phase shift is attained first and the r.f. oscillations stop even though the gain is greater than unity. The value of r required for suppression of the r.f. oscillations for such values of r_1 is given by equation (10).

TABLE I

$C_1 = 150\mu\mu f$; $g = 1.2 \times 10^{-3}$ mho

$r_1 = 3{,}000\,\Omega$; $gr_1 = 3.6$; $0.077r_1 = 231\Omega$			$r_1 = 1{,}850\Omega$; $gr_1 = 2.22$; $0.077r_1 = 142.5\Omega$		
r	$f \times 10^{-5}$	A	r	$f \times 10^{-5}$	A
0	6.13	1.80	0	9.93	1.11
10	6.21	1.79	18.5	10.33	1.09
40	6.45	1.75	37	10.72	1.07
60	6.62	1.73	55.5	11.12	1.05
100	6.94	1.69	74	11.52	1.03
150	7.35	1.64	92.5	11.92	1.01
200	7.76	1.59	111	12.31	0.99
230	8.00	1.56	140	12.93	0.96

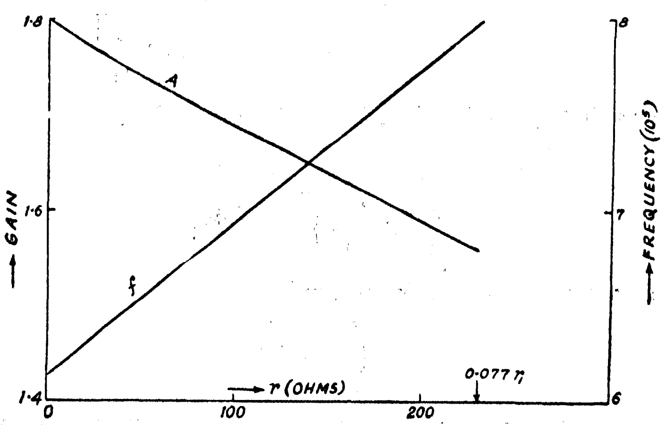


FIG. 6(a)

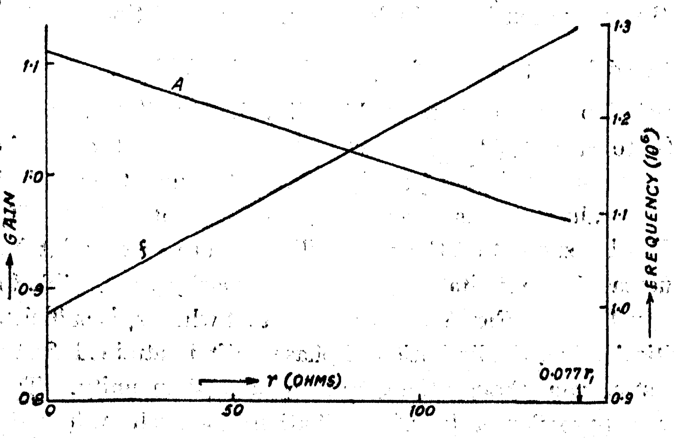


FIG. 6(b)

Thus with r_1 near the critical value, if r is gradually increased, the amplitude of the r.f. oscillations would gradually decrease to zero and then die out. On the other hand if the value of r_1 is much greater than the critical value the amplitude will no doubt decrease with increase of r, but instead of dying out on reaching zero amplitude the oscillations suddenly stop when the amplitude is still sufficiently large. These features are clearly observed when the experiments are performed with a cathode ray oscillograph as a visual indicator of the oscillations.

EXPERIMENTAL RESULTS

A series of experimental tests were performed to verify the validity of the theoretical relations discussed above. The complete oscillator was fitted up according to Fig. 1, modified by the introduction of r as shown in Fig. 5. The three resistances r_1 were wound with manganin wire on thin mica cards.

A. *Radio Frequency Oscillations—r Short-circuited*

In this condition it is immaterial whether C_1 is directly connected across r_1 as shown in Fig. 1. or connected between the anode and the H. T. negative line. In practice, for studying the effect of the variations of C_1, a three-gang condenser was used for the C_1's, the common shaft being connected to the H. T. negative line.

According to eqn. (5a) we have

$$f_1 = \frac{\sqrt{3}}{2\pi r_1 \; (C_1 + C_s)}$$

Or

$$C_1 = \frac{\sqrt{3}}{2\pi r_1} \cdot \frac{1}{f_1} - C_s \qquad \ldots \; (12)$$

TABLE II

Case	$C_1(\mu\mu f)$	f_1(kc/s)	$1/f_1 \times 10^6$	Cs ($\mu\mu$f) from graph
I	65 106 185 270 350	1490 1085 695 516 440	0.67 0.92 1.44 1.93 2.27	44
II	80 150 200 250 300 350	1538 922 735 610 512 440	0.65 1.08 1.36 1.64 1.95 2.27	28

This shows that if f_1 be changed by varying C_1 and $\frac{1}{f_1}$ is plotted along y-axis and C_1 along x-axis, the graph will be a straight line cutting the x-axis on the negative side, the intercept giving the value of C_s. This gives us a ready method of estimating experimentally the total stray and inter-electrode capacity on the two sides of the coupling condenser. Table II gives the record of observations on two typical oscillators. The values for Case I refer to the oscillator using 6K7G type valves and those for Case II to the second oscillator using 6SK7 type valves in which the connections were short and made with special care to minimise stray wiring capacities. The load resistance r_1 was in each case 2,000 ohms.

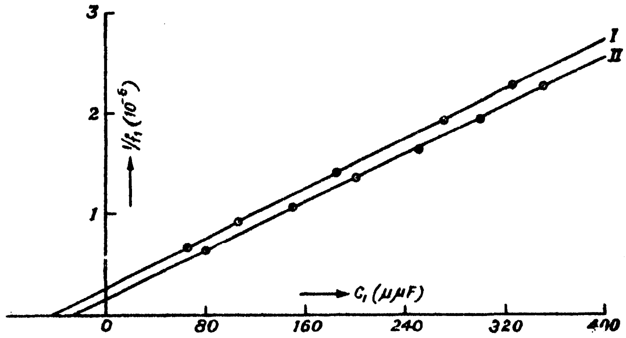


FIG. 7

The results are shown graphically in Fig. 7 from which we see that the stray capacity in Case II has been considerably diminished. It will be noted that there is a slight difference in the inclinations of the two straight lines to the x-axis, although according to eqn. (12) such a discrepancy should not have existed since r_1 was the same in each case. If, however, we remember that eqn. (12) is derived from (5a) which in itself is an approximation from (4c) it will be clear that the slope of the straight line is more correctly,

$$\tan^{-1} \frac{2\pi r_1}{\sqrt{3}\left(1+\frac{C_2}{C}\right)}$$

Now the values of C in the two cases concerned were different and C_2 automatically changed due to altered wiring. This explains the difference in the two slopes.

Fig. 8(a), Plate IX, shows a typical oscillographic record of the r.f. oscillations generated by the oscillator, frequency 500 Kc/s.

It is evident that to make f_1 very high, valves with high mutual conductance and low input and output capacitance must be used. Using 6SK7 valves with 1500 ohm load and having no external C_1 a frequency of 9 Mc/s has been obtained.

Quartz Control —The r.f. oscillations can be established by using a single quartz crystal in place of the coupling condenser C in any one stage. The crystal is selected to have resonant frequency within the range of the oscillation. The three-gang condenser for the C_1's in the three stages is adjusted till the crystal frequency is obtained. In this condition the coupling condenser C of any one stage is replaced by the mounted quartz crystal. It is found that the frequency of the maintained oscillations remains constant even though the ganged condenser C_1 is varied within wide limits.

Three-phase symmetry.—In the r.f. mode the phase shift produced by any one stage is primarily dependent upon $(C_1 + C_s)$ and r_1 when the grid leak and coupling condenser are, as usual, comparatively very large. By using three similar valves, identical wire-wound resistances for r_1, ganged condenser for C_1 and making symmetrical connections it is easy to maintain the phase shift produced by each stage at $120°$.

The equality of output voltage at each plate is checked by thermionic voltmeters. To test the 3-phase symmetry, the grids of the three oscillator valves are connected to the grids of three other triodes of which the anodes are tied together. The H.T. of the triodes is applied through a common load resistance. If the triodes are identical, it is obvious that the combined output will be zero when 3-phase symmetry exists. This has been verified experimentally.

B. *Audio Frequency Oscillations*

The short-circuits across the r's are removed for this mode of oscillation. The 3-gang condenser used for the r.f. mode is disconnected and three equal mica condensers are connected across the three resistances r_1. For generating oscillations of varying audio frequency the three coupling condensers C are this time replaced by a 3-gang condenser having each section completely insulated from the other two.

For a given setting of C the value of r is gradually increased till a stage is reached when the *r.f.* oscillation (f_1) stops and the audio frequency (f_2) builds up immediately. Fig. 8(*b*), Plate IX, shows a typical record of the a.f. oscillations, frequency 600 *c/s*. In a typical case using 6K7G valves with $r_1 = 3{,}000$ ohms and $C_1 = 150\mu\mu f$, the minimnm value of r required to suppress f_1 was found to be 230 ohms. This agrees with eqn. (10). As explained in connection with Fig. 6, it has been found that when r_1 is very near the critical value, f_1 stops for r much less than that given by eqn. (10). In a typical case with $r_1 = 1850$ ohms., f_1 stopped at $r = 100$ ohms. In such critical cases, however, it has been observed that f_2 does not build up as soon as f_1 dies down but that it is necessary to increase r still further for the start and maintenance of f_2. This would seem rather puzzling as with increase of r the audio gain increases, although the *r.f.* gain decreases.

It can be shown, however, that if the three stages are not identically the same and the grid leaks of two stages are equal while that of the third stage is either greater or smaller than the other two, the overall gain of the system for audio frequencies, is less than $gr_1/2$. The r.f. gain is not affected by variations in r_2. Under these conditions, therefore, when r_1 is very near the critical value, the r.f. oscillations are normally maintained when $r=0$. As r is gradually increased the r.f. gain decreases and finally f_1 dies down. The a.f. gain no doubt increases with increase of r but due to this asymmetry may still be less than unity for the value of r which just suppresses the r.f. oscillations. This explains why in such cases r has to be further increased in order to maintain the a.f. oscillations.

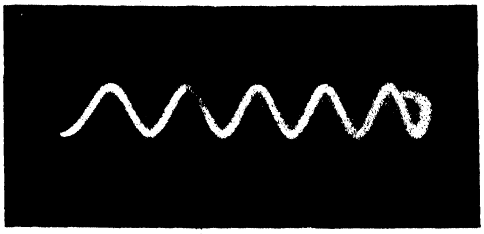
C. *Simultaneous R.F. and A.F. Oscillations*

When r_1 is neither very nearly equal to nor much greater than the critical value, it is found that if r is slowly increased from a small value with a view to build up the a.f. oscillations, a transition stage is reached when both the a.f. and r.f. oscillations may be simultaneously present. The oscillographic record, Fig. 8(*c*), Plate IX, depicts this condition with the time base synchronised to the generated a.f. oscillations, as may be noted from the trace of the fly-back path.

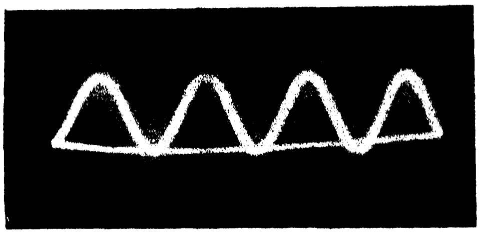
It may be noted in this connection that in a system capable of maintaining simultaneously two oscillations at frequencies widely different from each other, the higher one is generally believed to be modulated by the lower one. But no such modulation is present in the oscillographic record as is evident from the constancy of the amplitude of the r.f. oscillations. A radio receiver used to receive these oscillations also failed to indicate the presence of modulation under these conditions. When, however, r_1 is further increased and the valves no more operate within the linear regions of their characteristics, the audio oscillations cease to be sinusoidal and the trace on the cathode ray oscillograph indicates variations of the amplitude of the r.f. oscillations, indicating modulation. The radio receiver in that case gives the audio frequency output.

HARMONIC DISTORTION

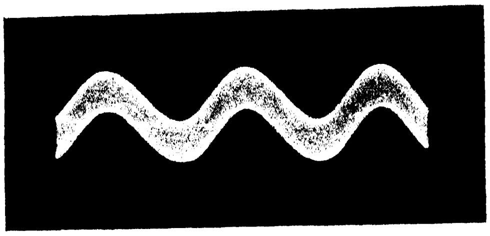
The phase-shift oscillator system behaves as a selective tuned circuit in so far as the overall phase shift is a function of the frequency, *i.e.*, at frequencies other than those given by equations (5a) and (7a) the feed back is not positive. Once oscillation has started from an initial minute impulse it will build up to such an amplitude that due to overloading the transconductance of a valve is reduced below the initial value and the overall gain of the amplifier is reduced just to unity. This will necessarily produce harmonics as in other conventional self-maintained oscillators. To reduce harmonic distortion it is therefore essential to make the load resistance r_1 just enough



(a)



(b)



(c)

to satisfy equation (5b). With further increase in r_1, the distortion will go on increasing.

It should be noted in this connection that, on a comparative basis, the radio frequency oscillation is purer than the audio frequency one. This will be appreciated if we remember that the limitation of the amplitude of oscillations is due to the variation of g with variations of grid voltage during an oscillation cycle. The allowable change in g being the same for either mode of oscillation, since the expression for effective gain is the same for either case, it is obvious that the particular mode for which the excursion of anode voltage is smaller will necessarily be purer. Now the effective anode load for the r.f. mode is half that for the a.f. mode with a given value of r_1 because of the shunting influence of C_1. The harmonic distortion is therefore comparatively less for the r.f. oscillations.

Further, as already mentioned, when the three stages are not identical, the grid leak of one stage being different from those of the other two, the overall a.f. gain of the system is less, for any given value of the load r_1, than if three-fold symmetry existed. In other words for a given overall gain a larger value of r_1 is required if three-fold symmetry is not maintained. In a system consisting of three identical stages oscillations are therefore maintained at a relatively lower value of r_1 and hence it is essential to maintain symmetry of the stages in order to keep the harmonics at a low level. For the r.f. mode the gain and phase shift depend primarily upon the resistances r_1, a ganged condenser being used for C_1. In practice, by using wire-wound resistances for r_1, it is fairly easy to maintain symmetry for this mode. For the a.f. oscillations when a ganged condenser is used for C, it is necessary to match the three grid leaks r_2 to equality if harmonic distortion is to be reduced. If this precaution is not adopted the oscillator will give greater distortion for the a.f. mode than for the r.f. mode. Again, when the grid excursion of any stage is sufficiently high to drive it positive and causes grid current to flow, the effective value of r_2 and hence the a.f. oscillations are thereby affected. From this point of view also the r.f. oscillations are purer than the a.f. ones.

CONCLUSION

In the present paper some of the essential features of the symmetrical type 3-phase R-C tuned oscillator have been discussed. The properties of the asymmetrical type will form the subject matter of a subsequent communication. It may however, be mentioned in this connection that the symmetrical r.f. oscillator can easily be made to generate frequency-modulated oscillations. A very simple method would be to shunt any of the three valves by, say, a triode and apply the modulating audio voltage to the grid of this triode. The effective anode load resistance of the corresponding stage will then vary and consequently produce wide-band frequency modulation.

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