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A NOTE ON SOME TUNABLE OSCILLATORS

N. B. CHAKRABORTY* and K. D. DIXIT

J. K. Institute of Applied Physics, University of Allahabad, Allahabad

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ABSTRACT. Some arrangements of voltage tunable two-path oscillators capable of large frequency deviation are discussed. Possible combinations of the transfer functions of the individual paths together with the tuning equation and the constraints for stable amplitude of oscillation are given. In some circuit arrangements the variation of frequency with modulating voltage is found to be linear over a wide range.

The present note discusses some circuit arrangements of voltage tunable oscillators which are theoretically capable of very large frequency deviation. In practical arrangements a frequency tuning ratio of two to one and in some cases five to one have been achieved. Tuning is accomplished by means of variation of the gains of the individual paths of a two-path oscillator.

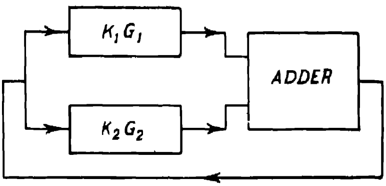

Fig. 1. Schematic diagram of a two-feedback loop.

If $G_1(p)$ and $G_2(p)$ are the transfer functions of the two paths of the feedback loop in Fig. 1 the characteristic equation of the loop is given by

$$K_1G_1+K_2G_2 = 1 \qquad \dots \quad (1)$$

Here K_1 and K_2 are the gain multiplying factors and are in general variable in accordance with the tuning voltage. If the adder is non-ideal and has a transfer function $F(p)$ then Eq. (1) is modified to

$$K_1G_1+K_2G_2 = \frac{1}{F} \qquad (1.a)$$

The condition for proper working is that the imaginary axis be the root locus with unity feedback.

In Table I some possible combinations of transfer functions together with the tuning equation and the constraints for stable amplitude of oscillation are

* Now at the Institute of Radio Physics and Electronics, University of Calcutta.

presented. All the circuits have the property that in theory it should be possible to tune their frequency of operation electronically from very low to very high frequencies. Furthermore the magnitude of their positive feedback voltage should be independent of the frequency of resonance. In some arrangements the curve of frequency versus modulating voltage is linear over a wide range. In a few arrangements where the stray capacitances can be taken into account in forming the transfer functions the tuning range can obviously be extended to figure of merit of the tubes employed or a fraction of it depending on the gain required

It will be observed that types 1-6 are practical at audio frequencies, types 1, 5 and 6 at low and very low frequencies, while types 1 6, 7 and 8 are useful at radio frequencies. In Fig. 2(a) and Fig. 2(b) are presented circuit arrangements

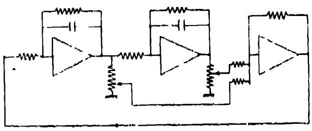

Fig. 2(a). Schematic diagram of the circuit arrangement of type 1 for low and very low frequencies

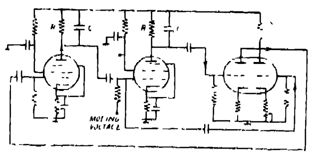

Fig. 2(b) Schematic diagram of the circuit arrangement of type 1 for A. F. and R. F.

of oscillator of type 1 for use at low and at radio frequencies For *RF* an interesting form derived from type 6 making use of delay lines can be made as shown in Fig. 3(a) and (b). Its tuning equation is

$$w\delta = \cos^{-1}(-K_1/2) \tag{2}$$

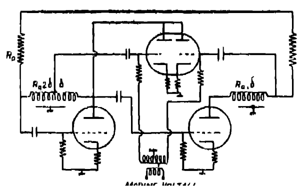

Fig. 3(a). Schematic diagram of a circuit arrangement of type 6.

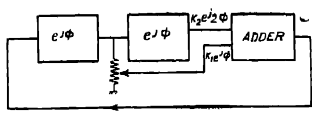

Fig. 3(b). Functional diagram of type 6.

The frequency of excursion in this case is limited to $\frac{1}{2\delta}$. The linearity of modulation around $\frac{1}{4\delta}$ has been observed to be quite good. It is to be noted that

TABLE I

Type	$G_1(j\omega)$	$G_2(j\omega)$	Constraint	Tuning equation	Differential equation	Working frequency
1	$\frac{1}{1+j\omega}$	$-\frac{1}{(1+j\omega)^2}$	$K_1=2$	$\omega = \sqrt{K_2-1}$	$\frac{d^2x}{dt^2} - 2\frac{dx}{dt} + x - K_2x - \frac{d}{dt}F(x) - F(x) = 0$	L.F., A.F. & R.F.
2	$\frac{\omega}{1+j\omega}$	$-\left(\frac{j\omega}{1+j\omega}\right)^2$	$K_1=2$	$\omega = \frac{1}{\sqrt{K_2-1}}$	$\frac{d^2x}{dt^2}(1+K_2) - 2\frac{dx}{dt} + x - \frac{d^2}{dt^2}F(x) - \frac{d}{dt}F(x) = 0$	A.F.
3	$\frac{1}{1+j\omega}$	$\frac{j\omega}{(1+j\omega)^2}$	$K_1-K_2=2$	$\omega = \sqrt{K_2-1}$	$\frac{d^2x}{dt^2} + 2\frac{dx}{dt} - x - F(x) - \frac{d}{dt}F(x) - K_2\frac{dx}{dt} = 0$	A.F.
4	$\frac{1}{1+j\omega}$	$\frac{1-j\omega}{(1+j\omega^2)}$	$K_1-K_2=2$	$\omega = \sqrt{2K_2-1}$	$\frac{d^2x}{dt^2} - 2\frac{dx}{dt} + x - K_2x - K_2\frac{dx}{dt} - F(x) - \frac{d}{dt} = 0$	L.F., A.F. & low R.F.
5	$\frac{1-j\omega}{1+j\omega}$	$\left(\frac{1-j\omega}{1-j\omega}\right)^2$	$K_2=-1$	$\omega = \sqrt{\frac{2-K_1}{2-K_1}}$	$\frac{d^2x}{dt^2} - 2\frac{dx}{dt} - x - K_1x - K_1\frac{d^2x}{dt^2} - \frac{d^2}{dt^2}F(x) + 2\frac{d}{dt}F(x) - F(x) = 0$	L.F., A.F. & low R.F.
6	$ejF(\omega)$	$ej2F(\omega)$	$K_2=-1$	$F(\omega) = \cos^{-1}\frac{K_1}{2}$	$x(t) - K_1x(t-\delta) + F(x(t-2\delta)) = 0$ if $F(\omega) = -\omega\delta$.	R.F.
7	$\frac{1-e^{-j\omega\delta}}{1+e^{-j\omega\delta}}$	$-\left(\frac{1-e^{-j\omega\delta}}{1+e^{-j\omega\delta}}\right)^2$	$K_1=0$	$\omega = \frac{2}{\delta}\tan^{-1}\sqrt{\frac{1}{K_2}}$	$x(t) - 2x(t-\delta) - x(t-2\delta)$ $-F(x(t)) - 2F(x(t-\delta)) + F(x(t-2\delta)) = 0$	R.F.
8	$\frac{1+e^{-j\omega\delta}}{1-e^{-j\omega\delta}}$	$-\left(\frac{1+e^{-j\omega\delta}}{1-e^{-j\omega\delta}}\right)^2$	$K_1=0$	$\omega = \frac{2}{\delta}\cos^{-1}\sqrt{\frac{1}{K_2}}$	$x(t) - 2x(t-\delta) - x(t-2\delta)$ $-F(x(t)) - 2F(x(t-\delta)) + F(x(t-2\delta)) = 0$	R.F.

in arrangements of types 6, 7 and 8, means will have to be adapted to supress harmonics.

The forms 1—5 have identical characteristic equations. However the way nonlinearities enter into the equation is different and consequently the amplitude stability and other related properties will be different. The nonlinear defining equations in the different cases are also presented in Table I. $F(x)$ represents the nonlinear parameter basically governing the amplitude of oscillation.

It has been found that the variation of frequency is limited in practice to a ratio of five to one in types 1—5, while in the rest it is about two to one. It is thought that this may be due to the very considerable variation of the slope of the loop phase shift. Another cause is the non-ideal behaviour of the adder.

The disturbing effect of the adder can be removed in the following manner. Supposing that the transfer function of the adder $\frac{b}{p+b}$, is the modified loop, equation is

$$(K_1G_1+K_2G_2)\cdot\frac{b}{p+b}=1 \qquad \ldots \quad (1.c)$$

The characteristic equation of the loop in the oscillating condition will now be

$$(p^2+K)(p+C)=0 \qquad \ldots \quad (3)$$

It will be found in this case that both K_1 and K_2 are to be varied. For type 1 for example K_1 and K_2 are related by

$$(2b+1)(2+b)=K_1b(2+b)+b(1-K_1-K_2) \qquad \ldots \quad (4)$$

TUNABLE AMPLIFIERS

It is evident that at a gain setting smaller than that required for self-oscillation all the circuits can be used as tunable amplifiers and hence as spectrum analysers. The nature of the selectivity curve and the variation of the maximum output of the response curve with the tuning frequency will obviously depend on the point of observation of the output as well as the nature of the transfer functions. In the diagram of Fig. 1 for example for inputs E_1, E_2 and E_A applied respectively at the input to G_1 circuit, G_2 circuit and the adder the voltage outputs at the corresponding output points will be given by

$$\frac{e_1}{G_1}=f\ (K_1G_1E_1+K_2G_2E_2+E_A)+E_1, \qquad \ldots \quad (5.a)$$

$$\frac{e_2}{G_2}=f\cdot(K_1G_1E_1+K_2G_2E_2+E_A)+E_2, \qquad \ldots \quad (5.b)$$

$$e_A=f.(K_1G_1E_1+K_2G_2E_2+E_A) \qquad \ldots \quad (5c)$$

where

$$f = \frac{1}{1-K_1G_1-K_2G_2}.$$

In the arrangement of Fig. 1, for example, if E_A be the input applied to the adder the output e_2 will be given by

$$e_2 = \frac{G_2}{1-K_1G_1-K_2G_2}E_A = \frac{1}{(2-K_1)jw_0}E_A, \qquad \text{at } p = jw_0.$$

The peak of response will therefore depend on the tuning frequency. If on the other hand, the voltage obtained at the output of G_1 circuit is transmitted through a circuit having a transfer function $\frac{p}{p+1}$ one obtains a transfer characteristic which does not depend on the tuning frequency. It should be noted that a much simpler solution is possible for the arrangement of type 3.

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