10th July 1951

To

Panchanan Kundu, Esq.,
20 A. Pashupati Bose Lane,
Calcutta 3.

Dear Sir,

During the absence of Prof. K.S.Krishnan, I have opened your letter written on the 4th July and posted on the 7th, which was received in our office on the 10th July.

As I ~~am~~ was afraid, the papers may not reach you in time. We have sent you a telegram reading as under :

" Papers Unavoidably Delayed. Please

Collect from Laboratory"

We have suggested that you ~~could~~ can collect the papers when you come to Delhi for the interview, from the National Physical Laboratory. The papers will be lodged with the undersigned and you can collect it any time from 9 A.M to 5.P.M.

Yours truly,

(T.V.Ramamurti)
Technical Secretary to
the Director.

we may start with expression (2) for U, and pick

out of it all those terms that involve   or its powers

As a preliminary to it we may first eliminate such

of the terms as obviously do not involve   or its

powers. Doing so, we are left with

from which we may now proceed to eliminate further

the unwanted terms. The various terms represent

interaction energies between the ion n 1, and

the other ions taken in succession. We

air off these terms in the following manner.

ample the interactions between ions,

1 and 2' , where 2' has the same

R-C TUNED OSCILLATORS*

by

P. Kundu, M.Sc. (*Associate*)†

SUMMARY

Generally a multivibrator-type oscillator is noted for its distorted output. But sinusoidal oscillations of good waveform can also be obtained from this system. In the present paper, the conditions for sinusoidal oscillations have been studied for both a straight-coupled and a cathode-coupled system. By proper choice of components, frequencies from a few cycles to a few Megacycles can be generated.

The limitation of these oscillators for wide frequency variation has been overcome by evolving a new type of frequency selective network, which, when connected in the feedback path of an amplifier, can shift the frequency of oscillation of the system up and down the frequency spectrum without affecting the loop gain of the system. It can be used as an oscillator, selective amplifier or a rejection filter at the same frequency for a given setting of the C R value of the network.

The salient feature is that the frequency can be varied by varying a single element and the necessity of using any automatic amplitude limiting device can be dispensed with due to the fact that the transmission constant of the network is independent of frequency.

1. Introduction

During recent years, much attention has been drawn to the R-C tuned oscillators [1-8] for generating sinusoidal oscillations. It is preferred because of its numerous advantages and superiority of performances over other types, especially in the audio range. The conventional L-C type oscillator is inconvenient due to the large inductance required, and with which it is rather impracticable to produce good waveform, due to the loss and non-linearity introduced by the coil in the lower audio frequencies. The range of frequency variation in one sweep of the capacity is only 3 : 1. This requires a large number of inductances for band switching, to cover the whole audio range. On the other hand, while the beat frequency oscillator is quite suitable for achieving a wide frequency coverage on a single tuning control, it has numerous shortcomings. "Pulling-in" of oscillators at low frequencies and spurious beat notes resulting from cross-modulation are occasional source of trouble; and it also requires frequent checking of calibration against the supply voltage to ensure the correctness of the calibration dial. Elaborate arrangements are required to get rid of all these troubles which, consequently, add to the cost and size.

R-C oscillators, on the other hand, are quite simple, easy to set up and suffer from none of disadvantages. In the present paper, the conditions for generating sinusoidal oscillations have been studied for both a straight-coupled and a cathode-coupled multivibrator. The limitation of these oscillators for producing wide frequency variation has been overcome by means of a new type of R-C selective network. An oscillator arrangement with this network compares favourably with those of existing types. The use of this network, of course, can quite conveniently be extended for a continuously tunable audio amplifier required for either selection or rejection of a particular frequency in a frequency spectrum. A basic design principle has also been developed for all these cases, so that the various elements can be proportioned to obtain the most satisfactory results.

2. Principle of Operation

Figure 1 shows symmetrical multivibrator circuit, which is essentially a two-stage R-C coupled amplifier fed back in a regenerative way. Self-oscillation in this system is only possible when the loop phase shift is zero and the loop gain is not less than unity. An amplifier with a purely resistive load produces a phase shift of 180 deg. But the R-C coupling network here introduces a phase lead at all frequencies. For oscillations to be sustained, this phase lead must be counteracted by an equivalent amount of phase lag. However, Fig. 1 omits components which are always present as stray, interelectrode and decoupling capacitances, and which play an important part in counteracting the phase lead so that at a particular frequency the feedback is purely positive. Supposing the gain of the system is quite high; oscillations then build up with growing amplitude from a minute unbalance in any part of the circuit. This cannot continue indefinitely, due to the non-linearity of the characteristic, and also due to the grid current which drives one valve beyond cut-off and makes further amplification impossible. This state of affairs continues while the charge on the grid condenser leaks away to raise the grid potential above the cut-off point, when the operation starts once again and ultimately drives the other valve beyond cut-off. The frequency of operation depends on the time-constants of the circuit elements.

It is quite clear that high amplification in the circuit drives a valve instantaneously into the non-linear portion of its characteristic and prevents it from conducting over a certain period of a complete cycle, giving rise to the relaxation type of oscillations.

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† Formerly with E.M.I., Ltd., Hayes, England, and C. G. de T.S.F., Paris; now with

U.D.C. No. 621.392.5.018.1 : 621.396.615.

If the bias and plate voltages of V_1 and V_2 are so arranged that the quiescent operating points of the valves are far beyond the grid current region and on the linear portion of the characteristic, and the circuit elements are so proportioned that the overall amplification is unity, then the circuit loss can be just supplied by the gain of the system only at a frequency at which the total phase shift is zero. This is the condition when sinusoidal oscillations will be generated, because the amplification at the harmonic frequencies will be too low to maintain oscillations, whereas at all other frequencies the feedback is other than positive.

Actual adjustment of amplification is, of course, rather critical. If it is too low, the system will fail to oscillate; if it is too much, the amplitude of oscillations will start increasing and the operating point will begin to swing into the non-linear portion of the characteristic. As a result, the mutual conductance of the valve will be reduced below its initial value reducing the overall gain to unity and giving rise to distortion as in any other self-maintained oscillators.

The output of the oscillator will, however, be essentially sinusoidal if care is taken that the voltage feedback to the grid is only very slightly in excess of what is needed for the maintenance of the oscillations, so that the swing of the operating point is limited only to a very small portion into the non-linear portion of the characteristic.

3. Theoretical Considerations

In Fig. 2 is shown one of the symmetrical stages of a two-stage coupled back-to-front arrangement. Adjustments have been made for linear operation, and provision for counteracting the phase lead due to the coupling capacitance C is made by connecting shunting capacitances across the anode load and the grid leak. Pentodes have been used to minimize the effect of the interelectrode capacitances. Since the two stages are identical, then for the limiting conditions of oscillation each stage must introduce a phase shift of 180 deg with a gain of unity.

The equivalent circuit of Fig. 2 is shown in Fig. 3*a*, in which the valve is shown as a constant current generator with a current output of $i = -g_m e_g$, where g_m is the mutual conductance of the valve.

Let

r_a = anode impedance of the valve
R_1 = plate load
R_2 = grid leak resistance
C = coupling capacitance
C_1 = shunting capacitance across the load + output and stray capacitances to the left of C
C_2 = shunting capacitance across the grid leak + input and stray capacitances to the right of C.

Assuming $r_a \gg R_1$, the equivalent circuit in Fig. 3*a* is reduced to Fig. 3*b*. By Kirchoff's Law:

$$i = i_{R_1} + i_{C_1} + i_C, \quad \text{where } i_C = i_{R_2} + i_{C_2}$$

Or, $i = i_{R_1} + i_{C_1} + i_{R_2} + i_{C_2}$

But

$$i_{R_1} = \frac{e_1}{R_1};\ i_{C_1} = C_1\frac{de_1}{dt};\ i_{R_2} = \frac{e}{R_2};\ i_{C_2} = C_2\frac{de}{dt}$$

Hence

$$\frac{e_1}{R_1} + C_1\frac{de_1}{dt} + \frac{e}{R_2} + C_2\frac{de}{dt} = -g_m e_g$$

Now,

$$e_1 = e + \frac{1}{C}\int\left(\frac{e}{R_2} + C_2\frac{de}{dt}\right)dt$$

$$\therefore \frac{de_1}{dt} = \frac{de}{dt} + \frac{1}{C}\left(\frac{e}{R_2} + C_2\frac{de}{dt}\right)$$

Since the amplification of each stage is unity, e_g must be equal to e

$$\therefore \frac{1}{R_1}\left\{e + \frac{1}{C}\int\left(\frac{e}{R_2} + C_2\frac{de}{dt}\right)dt\right\} + C_1\left\{\frac{de}{dt} + \frac{1}{C}\left(\frac{e}{R_2} + C_2\frac{de}{dt}\right)\right\} + \frac{e}{R_2} + C_2\frac{de}{dt} = -g_m e$$

Differentiating the above equation:

$$\frac{1}{R_1}\frac{de}{dt} + \frac{e}{CR_1R_2} + \frac{C_2}{CR_1}\frac{de}{dt} + C_1\frac{d^2e}{dt^2} + \frac{C_1}{CR_2}\frac{de}{dt} + \frac{C_1C_2}{C}\frac{d^2e}{dt^2} + \frac{1}{R_2}\frac{de}{dt} + C_2\frac{d^2e}{dt^2} + g_m\frac{de}{dt} = 0$$

Writing it as a differential equation:

$$\frac{d^2e}{dt^2} + \left(\frac{1}{R_1} + \frac{C_2}{CR_1} + \frac{C_1}{CR_2} + \frac{1}{R_2} + g_m\right) \Big/ \frac{CC_1 + CC_2 + C_1C_2}{C}\,\frac{de}{dt} + \frac{e}{CR_1R_2}\,\frac{C}{CC_1 + CC_2 + C_1C_2} = 0$$

a/ This is linear differential equation of the first degree and of the second order, with constant coefficients, which can be written in the more convenient and standard form

b/ $$\frac{d^2y}{dt^2} + 2\delta\frac{dy}{dt} + a^2y = 0$$

—Galley Three follows—

For undamped oscillatory solution of this equation the coefficient of the middle term is required to be zero or negative, when the frequency of oscillation is given by

$$f = \frac{1}{T} = \frac{a}{2\pi}$$

$$\text{i.e. } f = \frac{1}{2\pi\sqrt{R_1R_2(CC_1 + CC_2 + C_1C_2)}} \quad (1)$$

And for the oscillations to be maintained, the mutual conductance must be such that

$$g_m \nless \frac{1}{R_1} + \frac{1}{R_2} + \frac{C_1}{C}\cdot\frac{1}{R_2} + \frac{C_2}{C}\cdot\frac{1}{R_1} \quad \ldots (2a)$$

For the limiting condition of oscillation

$$g_mR_1 = 1 + \frac{R_1}{R_2} + \frac{C_2}{C} + \frac{C_1R_1}{CR_2} \quad \ldots\ldots\ldots (2b)$$

So long as the condition given in equation (2*b*) is maintained, the output of the oscillator will be sinusoidal.

Since the interelectrode capacitances of a pentode are very small, and since the gain required from the amplifier is not high, the effect of these capacitances can be neglected over a certain frequency range.

When

$$C = C_2 \gg C_1 \quad \text{and} \quad R_1 \ll R_2$$

the condition of maintenance is $g_mR_1 = 2$ . . (3*a*)

$$\text{and the frequency is } f = \frac{1}{2\pi C\sqrt{R_1R_2}} \quad \ldots . (3b)$$

The relations between the currents and voltages in the different branches of the circuit are as shown in the vector diagram of Fig. 4. Starting with the current i_{R_2} as the reference vector, i_C will be the resultant of i_{R_2} and i_{C_2}. The voltage e_C will lag by 90 deg. behind i_C and e_1 must be the resultant of e and e_C. i_{C_1} will lead by 90 deg. and i_{R_1} will be in phase with e_1. Then the total current $i = -g_me_g$ must be the resultant of $(\vec{i_C} + \vec{i_{C_1}})$ and i_{R_1}. The vector diagram is drawn here, showing the condition when the output voltage e is 180 deg. out of phase with the input voltage e_g.

i_c

In this circuit each stage is associated with a frequency selective network, so that the elements of both stages are to be varied simultaneously for any variation of frequency. The circuit becomes much more simplified if a cathode-coupled amplifier system is used with its input being fed-back from the output through a coupling impedance, as shown in Fig. 5*a*.

This system can be regarded as a cathode follower V_1, driving a grounded grid amplifier V_2 through the cathode resistance R_k. The high input impedance resulting from the cathode follower action and the shielding and impedance stabilization due to the grounded grid stage contribute to the wide-band characteristic of the amplifier.[9] The linearity of the amplifier is also much improved by the high degeneration in the cathode circuit. Moreover, this has an additional advantage that under the normal operating condition, grid current cannot flow due to the limiting action of plate current in the negative grid region.[8]

There is a phase reversal of 360 deg. between the input and the output, so that a sinusoidal oscillatory condition can be maintained if no additional phase shift is introduced by the coupling impedance at a frequency at which the gain is unity.

The maintenance condition and the frequency of oscillations have been derived in Appendix I.

Equation (ii) suggests that the frequency can be varied over a range 1 : 10 by varying C and C_2, by means of a two-ganged condenser; but as in the previous case, variation of R_1 varies the gain of the system and upsets the optimum operating condition, so variation of R_1 must be accompanied by an inverse variation of the mutual conductance of the valve. Generally, a variable-mu valve incorporated with some amplitude limiting device serves the purpose quite well so long as the variation of gain is very small.

Where an oscillator capable of a small range of frequency variation is required, the system of Fig. 5a is very advantageous due to its simplicity, stability and purity of wave form. With proper adjustment it can be set up to any frequency from a few cycles to a few Megacycles. The percentage of distortion can be controlled by manual variation of R_K. Automatic amplitude stabilization can be achieved by replacing R_K by a lamp.

a/

A comparison between a Class-A L-C type and an R-C type oscillator might be interesting here. In both the cases, an amplifier is fed-back through a tuning network. In L-C tuning, oscillation can be kept easily to a single frequency, because both the amplitude and the phase are altered sharply on both sides of resonance of a tuned circuit, whereas this sharpness of selectivity is far less with R-C tuning. Moreover, an R-C circuit lacks the quality of storing and exchanging of energy of an L-C circuit, which shapes the sinusoidal nature of an oscillation even when the energy is fed in the circuit by pulses.

This is why a slight variation of gain in an R-C circuit upsets the optimum operating condition and gives rise to distortion. It is always advisable to use some sort of automatic amplitude-limiting device like a lamp or a thermistor in the positive or negative feedback path, so that the overall gain is adjusted to the required value by the r.m.s. value of the output voltage, unlike the self-maintained oscillators, which depend on the non-linearity of the characteristic to limit the amplitude for stable oscillation.

4. Considerations for the Second Type of 6 Oscillator (Fig. 6)

It is clear from the above discussion that stable and uniform performance can be achieved over a wide frequency range if use is made of a feedback network which, besides having no loading effect on the amplifier, has an attenuation coefficient which is constant with frequency and a phase coefficient which passes through zero or 180 deg. at the frequency of operation.

In general, the arrangement can be considered as an amplifier coupled back to its input through a frequency selective network as shown in Fig. 6.

Let A = gain of the amplifier

β = fraction of the output voltage fed-back to the input.

Both A and β are complex functions of frequency

where $A = |A| \underline{|\varphi}$

and $\beta = |\beta| \underline{|\psi}$

According to the Nyquist criterion of oscillation, self-oscillation in this system can only be possible if the Nyquist diagram of loop gain in the complex plane, with frequency as the varying parameter encloses the point (1, 0) of the complex plane.

i.e. if $G = |A|\,|\beta|\,\underline{|\varphi + \psi}$ ....(4a)

oscillations occur when

$|A|\,|\beta| = 1$ ....(4b)

and

$\varphi + \beta = 0$ ....(4c)

Now, if A_0 = gain of the amplifier when the feedback is taken into account,

Then

$$A_0 = \frac{A}{1 - A\beta} \text{ ....(5)}$$

Conditions:

(i) When β is positive

and $A\beta = 1$

$A_0 = \infty$

This is the condition for sinusoidal oscillations as the circuit loss is supplied by the gain of the system at a frequency at which the overall phase shift is zero

(ii) When β is positive

and $A\beta < 1$

$A_0 \geqslant A$

i.e. when the amount of regeneration is such that the gain round the loop is just below a point required for oscillation, the system can be used as a tunable frequency selective amplifier.

(iii) When β is negative, i.e. when the feedback is degenerative and when $A\beta \geqslant 1$,

Then $A_0 = \dfrac{1}{\beta} \leqslant A$

With proper arrangement, a frequency can be tuned out from the output of the system by adjusting the elements of the network and the system can be used as a rejection filter.

5. The Feedback Network

When a resistance R_1, R_2 and a capacitance C_1, C_2 are connected in series across the Cap/ secondary AC of a centre tapped transformer T, as shown in the Fig. 7a, the phase of the voltage across OP and OP′ can be varied between zero and 180 deg. (not including) by varying the CR values.[10, 11] The amplitude of the voltage remains constant at one-half of the input voltage for all degrees of phase shift. This is due to the fact that the locus of P or P′, must be on the circumference of the semi-circle CPA or CP′A, respectively, drawn on CA as diameter, because of the quadrature relation of the two voltages V_{R_1} and V_{C_1} (or V_{R_2} and V_{C_2}), the input voltage CA $(E' + E)$ being always the resultant of the two as shown in the vector diagram of Fig. 7b.

When $C_1R_1 = C_2R_2$, the voltage across P′P supplies a balanced output with respect to the centre tapped point O. The phase lead of $\overrightarrow{P'P}$ being θ_1 with respect to the reference vector $\overrightarrow{CA}$

then $\tan\alpha = \dfrac{X_{C_1}}{R_1}$ But $\theta_1 = 2\alpha$

$$\therefore \theta_1 = 2\tan^{-1}\frac{X_{C_1}}{R_1} \text{ ....(6a)}$$

If another resistance R_3 and capacitance C_3 are connected in series across P′P, then the output voltage $\overrightarrow{OB}$ remains constant at one-half of the voltage P′P (P′P = CA = 2OB) over all degrees of phase shift.

Case (1): When P′P is connected to point 1, 1, as in Fig. 7a, the output voltave OB leads P′P by an angle θ_2 given by $\theta_2 = 2\tan^{-1}\dfrac{X_C}{R_3}$ ......(6b)

Here the phase advance θ_2 due to C_3R_3 is additive to the phase advance θ_1 due to C_1R_1 and C_2R_2 with respect to the reference vector $\overrightarrow{CA}$.

——Galley Five follows—

For $\psi = \theta_1 + \theta_2$ to be 180 deg.,

$2 \tan^{-1} \frac{X_{C_1}}{R_1} + 2 \tan^{-1} \frac{X_{C_3}}{R_3}$ must be equal to 180 deg.

or i.e. $\tan^{-1} \frac{X_{C_1}}{R_1} + \tan^{-1} \frac{X_{C_3}}{R_3} = \frac{\pi}{2}$ $\dfrac{\frac{X_{C_1}}{R_1} + \frac{X_{C_1}}{R_3}}{1 - \frac{X_{C_1} X_{C_3}}{R_1 R_3}} = \infty$

or,

or $1 = \dfrac{1}{\omega^2 C_1 R_1 C_3 R_3}$

Then the angular frequency:

$$\omega = \frac{1}{\sqrt{C_1 R_1 C_3 R_3}} \quad \ldots\ldots\ldots\ldots\ldots (7a)$$

Case (2): When P′P is connected to point 2, 2, in Fig. 7*a*, the output $\overrightarrow{OB}$ undergoes a phase retard of θ_2 with respect to P′P where

$$\theta_2 = 2 \tan^{-1} \frac{R_3}{C_3}$$

For $\overrightarrow{OB}$ to be in phase with $\overrightarrow{CA}$, i.e., ψ to be zero, θ_1 must be equal to θ_2

Or

$$\tan^{-1} \frac{X_{C_1}}{R_1} = \tan^{-1} \frac{X_{C_3}}{R_3}$$

When the angular frequency

$$\omega = \frac{1}{\sqrt{C_1 R_1 C_3 R_3}} \quad \ldots\ldots\ldots\ldots\ldots (7b)$$

The transformer can be replaced by a triode phase-splitting valve having equal anode and cathode loads. Since the same current passes through the resistances, an input voltage is converted to two output voltages of equal magnitude and opposite phase across the anode and cathode of the valve. The tube characteristic has negligible effect on the output voltage due to the degeneration at the cathode circuit, while equality and phase opposition of the voltage are maintained for frequencies from zero up to a point where the effect of the distributed and stray capacitances can be neglected.

The gain of the system is $\dfrac{1}{\frac{1}{g_m r} + \frac{2}{\mu} + 1}$

i.e., when $\mu \gg 1$ and $g_m r \gg 1$, the gain is approximately unity per phase and the condition of balance is only the equality of anode and cathode resistances.

The network can be connected as shown in Figs. 8*a* and 8*b*.

In both cases, the magnitude of the output e_o is independent of frequency and approximately equal to the input e_g. But in Fig. 8*a*, the phase of e_o is zero with respect to the input voltage e_g, while in Fig. 8*b*, the output voltage e_o is 180 deg. out of phase with respect to e_g, at a frequency given by the equations (7*a*) and (7*b*) respectively.

It must be pointed out that the above vector diagram is drawn on the assumption that

(i) The loading effect of the network is negligible, i.e. there is negligible feed-back from the anode to the cathode of the valve.

(ii) Effects of stray and distributed capacitances are negligible.

(iii) The respective CR values are so proportioned that the impedance between PP′ is very large compared to the impedance between AC.

The use of Fig. 8*b* is preferred to 8*a*, because in the latter case the output is required to be taken through a blocking capacitor in order to block the d.c. anode potential present at the output. Since this extra component, together with its associated leak resistor, becomes part of the phase shift network, it is important to ensure a negligible phase shift in the combination. The d.c. potential present in the output of Fig. 8*b* can be arranged to have no effect by adjusting the cathode bias of the next stage.

6. The Complete Oscillator

If any one of these networks is inserted in the feedback chain of an amplifier system and the phase shift round the loop is adjusted to zero by means of an appropriate number of amplifiers, and the loop gain is adjusted to unity, then this system will be a source of sinusoidal oscillations.

The amplifier should be linear and should have a constant time delay wide band characteristic. Generally, the output of an amplifier suffers a phase lead at low frequencies, due to the reactance of the coupling capacity, and a phase lag at high frequencies due to the shunting effect of the stray and interelectrode capacities; care should be taken to make the phase characteristic ~~linear~~ over as wide a band as possible. t/

t

of the amplifier constant

( ) D

(Shift for amplifier ~~constant~~)

Since the transmission constant through the selective network is almost unity, the gain required from the amplifier for the limiting condition of oscillation is just over unity. The low gain demanded from the amplifier will allow it to operate more linearly with low distortion.

Under these conditions, the equality of the absolute values of the gain and phase shift of the amplifier ensures a constancy of amplitude and stability of the oscillator, since the R-C network determines only the frequency.

It should be noted that at the generated frequency this type of oscillator behaves ~~is~~ a selective tuned circuit, the decrement of which can be varied continuously by controlling the amount of reaction in the regenerative loop from slight positive through zero to negative damping. At slight positive damping it acts as a selective amplifier. With due care to phase shift and gain the same system can be used as a rejection filter also. a/

a

—Galley Six follows—

In Fig. 9, an oscillator arrangement is shown in which only two valves are required. The amplifier stage V_1 should have a constant phase shift of 180 deg. over the whole of frequencies to be considered. The network produces a phase shift of another 180 deg., and thus fulfils the condition for oscillation. The anode and the cathode resistances of V_2 should be relatively small in value so that the output impedance of the phase inverter is low. The amount of regeneration can be controlled by the potenionmeter P, which can be set for a stable oscillation with a very low distortion. Both the valves can be the sections of a twin triode or the valve V_1 can preferably be a variable-mu pentode and V_2 a high-g_m and high-mu triode.

7. Discussion

In the conventional "Wien bridge" or a parallel-T circuit, two or three ganged elements are required for the continuous frequency coverage. If the elements are not in perfect track, the amount of feedback varies, and as a result the optimum operating condition is upset; some form of automatic limiting device must be used to maintain the constancy of output and the purity of wave form. When a lamp or a thermistor is used, the performance deteriorates when the overall temperature varies the normal resistance of the element.

Whereas the Wien bridge and parallel-T circuits are null networks at the frequency of operation, the feedback circuits of Figs. 8*a* and 9 have a constant attenuation coefficient at all frequencies and a phase shift of 180 deg. or 0 deg. at the frequency of operation.

In the oscillator described in this paper, the frequency can be varied by varying a single element (either C_3 or R_3) at a time, and since the feedback voltage is independent of the operating frequency, no automatic amplitude limiting device is necessary. The whole audio range can be covered in decimal steps by changing C_1, C_2 and C_3 in steps of 10, and using a variable resistance R_3 (100 : 1) to cover each band continuously.

The frequency stability is dependent on the

(i) Mechanical rigidity.

(ii) Temperature coefficient of the frequency determining elements.

(iii) Valve parameters and the input and output capacitances.

The effect of valve parameters can be minimized by using an amplifier valve of high-r_a; preferably a pentode. Since the gain required is just over unity, the anode load is very small, and as a result small variation of r_a will have negligible effects.

The lower gain also reduces the effect due to Miller capacitance at the input, while the shunting effect due to the output capacitance is negligible when the anode resistance of the amplifier V_1 is low.

8. Conclusions

The network described herein can also be used as a constant output R-C filter of all-pass-type, its phase-coefficient being a function of frequency.

Regarding the types shown in Fig. 2 and Fig. 5, it can be pointed out that these systems can be a source of frequency modulated signal. A convenient way of achieving the frequency modulation is to shunt one anode load by a triode valve and to vary its static resistance by varying the control grid voltage.

9. Acknowledgments

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—Galley Seven follows—Appendix

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Appendix

Analysis of Oscillator of Fig. 5a

It is assumed that $r_p \gg R_1$ (V_2 being a pentode); ref. Fig. 5*a*, *b*, *c*.

$$i_1 = g_{m1} e_{gc1}$$
$$i_2 = g_{m2} e_{gc2}$$

Let

R_K $$e_{gK} = (i_1 - i_2)\,\cancel{r_1} = e_{gc2}$$

Then

$$e_{gc2} = R_K(g_{m1} e_{gc1} - g_{m2} e_{gc2})$$

$$\text{or } e_{gc2} = \frac{g_{m1} e_{gc1} R_K}{1 + g_{m2} R_K}$$

Voltage across

$$Z = i_2 Z$$
$$= g_{m2} e_{gc2} Z$$
$$= g_{m2} \frac{g_{m1} e_{gc1} R_K}{1 + g_{m2} R_K} \cdot \frac{Z_1(Z_2 + Z_3)}{Z_1 + Z_2 + Z_3}$$

therefore voltage

$$e_o = g_{m2} \frac{g_{m1} e_{gc1} R_K}{1 + g_{m2} R_K} \cdot \frac{Z_1 Z_2}{Z_1 + Z_2 + Z_3}$$

The conditions for oscillations will be satisfied when

$$e_o = e_{gc1} + i_1 R_K$$

i.e.
$$g_{m2} \frac{g_{m1} e_{gc1} R_K}{1 + g_{m2} R_K} \cdot \frac{Z_1 Z_2}{Z_1 + Z_2 + Z_3}$$
$$= e_{gc1}(1 + g_{m1} R_K)$$

Or.
$$g_{m2} \frac{g_{m1} R_K}{1 + g_{m2} R_K} \cdot \frac{1}{(1 + g_{m1} R_K)} \cdot \frac{Z_1 Z_2}{Z_1 + Z_2 + Z_3} = 1$$

Or
$$\frac{g_{m1} g_{m2} R_K}{(1 + g_{m2} R_K)(1 + g_{m1} R_K)} \cdot \frac{R_1 R_2 C}{\sum CR + \frac{1}{j\omega}\left[1 - \omega^2 R_1 R_2 \sum CC\right]} \quad \ldots \text{(i)}$$

where

$$\sum CR = CR_1 + CR_2 + C_1 R_1 + C_2 R_2$$
$$\sum CC = CC_1 + CC_2 + C_1 C_2$$

From which the frequency of oscillation

$$f = \frac{1}{2\pi \sqrt{R_1 R_2 (CC_1 + CC_2 + C_1 C_2)}} \quad \ldots \text{(ii)}$$

and the limiting condition for the maintenance of oscillation

$$\frac{g_{m1} g_{m2} R_K R_1}{(1 + g_{m1} R_K)(1 + g_{m2} R_K)} \cdot \frac{1}{1 + \frac{R_1}{R_2} + \frac{C_2}{C} + \frac{C_1}{C} \cdot \frac{R_1}{R_2}} = 1 \quad \ldots \text{(iii)}$$

when

$$C = C_2 \gg C_1 \text{ and } R_1 \ll R_2$$

$$f = \frac{1}{2\pi C \sqrt{R_1 R_2}} \quad \ldots\ldots\ldots\ldots \text{(iv)}$$

and
$$\frac{g_{m1} g_{m2} R_K R_1}{(1 + g_{m1} R_K)(1 + g_{m2} R_K)} = 2 \quad \ldots \text{(v)}$$

when

$$g_{m1} = g_{m2} = g_m \quad \text{and } g_m R_K = 1$$

8 $$g_m R_1 = \cancel{4} \quad \ldots\ldots\ldots\ldots \text{(vi)}$$

Fig. 1.—Basic symmetrical multivibrator circuit.

Fig. 2.—One stage of a symmetrical back-to-back multivibrator circuit.

Fig. 3.—Equivalent circuits of Fig. 2.

Fig. 4.—Vector diagram for circuit of Fig. 3b.

Fig. 5.—(a) Cathode-coupled back-to-back mulvibrator circuit; (b) and (c) Equivalent circuits of (a).

Fig. 6.—Representation of oscillator as an amplifier coupled back to input through frequency selective network.

Fig' 7.—Basic feedback network and vector diagrams.

Fig. 8.—Connection of network to phase-splitting triode.

Fig. 9.—Circuit arrangement of complete oscillator using only two valves.

Prof. Krishnan,
National Physical Laboratory,
New Delhi.

Dear Sir,

I am much obliged to you for the interview you so kindly granted to me on the 7th ultimo in Delhi, after I was introduced to you by Dr. Sarwate. As desired by you, I was anxiously waiting to see you in Calcutta when you come here, but unfortunately for me, I learnt on enquiry that you had left for U.K. I would be glad if my services be of use to the Electronics Section of the National Physical Laboratory and I would be grateful, if considering my qualification and experience, a suitable opportunity be given to me.

I will have to appear for an interview before the Union Public Service Commission by the 13th July. I am therefore, very much in need of the Papers I left with you. I would be much obliged if you will kindly advise the office to send those Papers to me as soon as possible.

I hope, you will excuse me for the trouble.

Trusting this finds you in best of health,

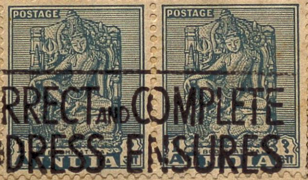
I remain,

Sir,

Yours sincerely,

20A, Pashupati Bose Lane,
Calcutta-3.
4th July, 1951.

~~PK~~ Panchanan Kundu
(Panchanan Kundu)



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CORRECT AND COMPLETE
ADDRESS ENSURES
QUICK DELIVERY

Prof. Krishnan,

Director,

National Physical Laboratory,

New Delhi.

From:

P. Kundu.
20A, Pashupati Bose Lane,
<u>Calcutta-3.</u>

Name: Panchanan Kundu. 20A, PASHUPATI BOSE LANE. CALCUTTA. 3.

Age: 27years.

Educational Qualifications:

i. Matriculation	Ist Div	Cal Univ	1939.
ii. I.Sc.	,,	,,	1941.
iii. B.Sc. with Honours in Physics		,,	1943.
iv. M.Sc. in Applied Physics, specialising in "Line and Radio Communication Engineering".		,,	1945.

Research Experience:

Research Scholar in Cal Univ, worked on "Frequency-modulated R-C coupled Oscillator", under Dr. H. Rakshit, D.Sc., F.N.I., in the Univ College of Science, Calcutta.

1946-1947.

Education & Training in U.K. & France:

i. Post-graduate Diploma in "Radio Communication Engineering" from the Marconi College of Wireless Communication, Chelmsford, Eng.

ii. Apprenticeship in the Works and the Test Division of the Marconi W/T Company, Chelmsford, England.

Session: 1947-1948.

Experience & Training:

i. Joined the Research and Design Dept. of the E.M.I.Ltd., Middlx., England, as a Design Engineer and worked thete for on year.
Oct. 1948 to Sept. 1949.

ii. Completed a course of training on "Radar and Radio Aids" for communication, navigation and landing from the Signals Training Establishments of the British Ministry of Civil Aviation, with practical Experience in the Stansted and the London Airport.
Oct. 1949 to Feb. 1950.

iii. Worked as an Apprentice Engineer for a course of practical training in the British Broadcasting Corporation.
March 1950 to June 1950.

iv. Worked as an Apprentice Research Engineer in the Centre de Recherches of the Compagnie Generale de Telegraphie Sans Fil in Paris,
July 1950 to Feb. 1951.

Patent:

"A Linear Time-base Gemerator"
Patent right applied by the Cie. Gele. T.S.F. on the 7th Feb, '51
No. 604409.

Publications:

i. "New Types of R-C Tuned Oscillators"
To be published in Journal of the Brit. Inst. R.E.

ii. "Measurement of Dielectric Constant and Loss of some Solid Materials at Centrimetric Waves".

Membership:

Assoc. Brit. I.R.E.

COMPAGNIE GENERALE DE TELEGRAPHIE SANS FIL

Centre de Recherches Techniques.   Paris, The I4 Feb '5I.

Report on the Apprenticeship of Mr. P. KUNDU.

Mr. P. KUNDU has undergone an apprenticeship at the Centre de Recherches Techniques de la COMPAGNIE GENERALE TELEGRAPHIE SANS FIL from July I950 to Feb I95I. During this period, he has worked in different laboratories. From July I950 to Sept I950, he devoted himself in the study of radar circuit technique, hyperfrequency aerials applications of magnetrons, under the guidance of Mr GUTTON, Director of General Research Department.

Then he worked in the Department Electronique, under the guidance of Mr. WARNECKE. He took an active part in the qorks concerning the electron optical part of certain types of progressive wave tubes and the study of the circuits associated with them. He completed a certain number of measurements concerning the dielectric constant and H.F. resistance of some materials of high loss. Finally, he has invented A Linear Time-base Generator which has been an object of application for a patent.

His work has been very satisfactory, and he has shown a good knowledge of theory as well as of practical works. This, mingled with a pleasant character, that permitted him to keep good relations with the personnel of the laboratories, will surely make of him an excellent collaborator.

Sd/- R.WARNECKE   P. GUENARD

Director of the Electronic Dept.   Chief of the Laboratories.

CSF 750

W/Gx

Brevet déposé le -7 FEV 1951
sous le n° 604409

B R E V E T D'I N V E N T I O N

BASE DE TEMPS LINEAIRE A TUBES A VIDE

COMPAGNIE GENERALE DE TELEGRAPHIE SANS FIL

La présente invention , système P. KUNDU, se rapporte aux générateurs de tension en dents de scie employés comme bases de temps dans les oscillographes électrostatiques ou électromagnétiques utilisés dans les appareils de mesure ou de télévision, ou dans toutes les applications nécessitant une tension à variation linéaire et à retour rapide. Elle vise à réaliser un générateur dont le montage ne comporte que deux tubes à vide et qui réunit la condition de linéarité très satisfaisante avec la simplicité, la stabilité, la faible valeur de tension d'alimentation et la possibilité de réglage dans une large gamme de fréquences de dents de scie.

Parmi les bases de temps connues, celles qui emploient les tubes à gaz, bien que simples et satisfaisantes dans les conditions normales, présentent les inconvénients de gamme de fréquences limitée, de fonctionnement instable, et de défaut de synchronisation, en raison des phénomènes d'ionisation. Il est donc préférable d'employer des tubes à vide, mais les montages correspondants sont généralement plus complexes et nécessitent un plus grand nombre de tubes. Certains montages spéciaux élaborés en vue d'éliminer cet inconvénient tels que l'oscillateur à blocage périodique, ne fonctionnent que dans une gamme de fréquences limitée; d'autres, tels que le montage à couplage cathodique ou le circuit connu sous le nom de transitron, nécessitent un tube supplémentaire ou une tension d'alimentation élevée par rapport à la pointe de dents de scie pour assurer la linéarité du balayage; d'autres encore, tels que l'intégrateur Miller, sont presque parfaits de ces différents points de vue mais donnent une variation de tension dans le sens négatif.

La présente invention, qui élimine ces inconvénients, consiste dans un montage générateur de dents de scie, dans lequel le condensateur aux bornes duquel la tension en dents de scie est prélevée, est chargé à travers une triode à vide, tandis que la décharge s'effectue à travers un autre tube à vide, branché en shunt sur ce condensateur et dont le circuit de sortie est combiné avec le circuit d'entrée de la triode précédente de façon à exercer une contre-réaction sur celui-ci par le courant anodique du tube en shunt; celui de ces deux tubes dont la conduction de courant correspond à la pente de dents de scie, est en outre muni d'un circuit de contre-réaction.

L'invention est applicable aux générateurs à auto-relaxation aussi bien qu'à ceux synchronisés par une impulsion extérieure; elle permet de réaliser des bases de temps à charge lente correspondant à la variation de tension dans le sens positif et à décharge rapide correspondant au retour à zéro, ou inversement à charge rapide et à décharge lente. Elle admet de multiples variantes dont les possibilités seront comprises en examinant les figures 1 à 5 qui représentent quelques exemples non limitatifs de réalisation de montages conformes à l'invention.

Le schéma de la figure 1 montre un condensateur C, aux bornes duquel on prélève la tension de sortie, en série avec un tube à vide V_1 contenant dans son circuit d'anode une résistance R_c et dans son circuit de cathode une résistance R_K (l'ensemble étant branché sur une source de tension continue E_b). Le circuit de grille de cette lampe contient une résistance R_2 qui est commune au circuit anodique d'un second tube à vide V_2 branché en shunt sur le condensateur C et ayant son anode connectée à la grille du tube V_1. La grille du tube V_2 est couplée à la plaque du tube V_1 par un condensateur C_1, tandis que son espace grille-cathode est shunté par une résistance de fuite R_1. Les résistances R_K et R_2 peuvent être réglables.

L'existence, entre le condensateur C et la source, du tube V_1 avec une résistance anodique R_c et une résistance de contre-réaction R_K confère à la charge un caractère à courant sensiblement constant, puisque les caractéristiques dynamiques d'une triode à contre-réaction de courant tendent de se rapprocher de celles d'une pentode .Par conséquent, la tension aux bornes du condensateur C croit de façon sensiblement linéaire jusqu'à une valeur de pointe qui n'est pas beaucoup plus basse que la tension d'alimentation, ce qui correspond à l'avantage du dispositif conforme à l'invention de ne pas nécessiter une tension d'alimentation élevée relativement à une tension donnée de pointe de dent de scie.

Le dispositif fonctionne comme suit : après l'application de la tension d'alimentation, le condensateur C commence à se charger à courant constant comme il a été indiqué, tandis que le condensateur C_1 se charge à travers les résistances R_C et R_1 shuntée par la résistance interne grille-cathode de V_2, jusqu'à ce que son armature inférieure soit amenée au potentiel de la grille de V_2 qui est maintenu au voisinage du zéro par auto-polarisation due au passage du courant grille dans la résistance R_1. Ce potentiel étant voisin du zéro, le tube V_2 ne laisse pas passer le courant tant que son potentiel anodique déterminé par la tension aux bornes du condensateur C n'a pas atteint une valeur suffisante. Aussitôt que cette valeur est atteinte, le courant anodique passe par V_2,par conséquent le condensateur C commence à se décharger à travers ce tube. Le passage du courant anodique produit une chute de tension dans la résistance R_2, c'est-à-dire une contre-réaction supplémentaire sur le tube V_1 dont le potentiel de grille subit une variation à tendance négative qui bloque V_1 et arrête la charge de C. Cette variation, grâce à la propriété d'inversion de phase entre les tensions grille et plaque de tubes électroniques, donne lieu à une variation à tendance positive du potentiel anodique de V_1, variation qui est transmise par le condensateur C_1 à la grille de V_2, ce qui augmente le courant de plaque de ce tube et la chute de tension dans la résistance R_2. Le phénomène est cumulatif, de sorte que le condensateur C est soumis à la décharge à courant continuellement augmentant et limité seulement par la saturation du tube V_2; il se décharge donc très rapidement, et la tension à ses bornes tombe à zéro , ce qui arrête le passage de courant par V_2 et débloque V_1 qui recommence à conduire le courant. La chute de tension dans la résistance R_C apporte au potentiel de plaque de V_1 une variation à tendance négative qui est transmise par C_1 à la grille de V_2 et bloque ce tube, une variation à tendance positive étant engendrée sur la plaque de V_2, c'est-à-dire sur la grille de V_1, cette variation activant le passage de ce tube de l'état bloqué à l'état de conduction, tandis que le blocage du tube V_2 est accentué par cumulation du phénomène. On a ainsi un arrêt très net et très énergique de la période de décharge, après quoi la charge du condensateur C recommence et le cycle de fonctionnement se répète.

On a ainsi réalisé un générateur à auto-relaxation avec montée de tension linéaire et avec retour rapide. La rapidité du retour sera encore améliorée en prenant pour V_2 un tube à forte amplification et à faible résistance interne. Toutefois, un meilleur moyen pour améliorer à la fois la linéarité de montée et la rapidité du retour consiste à employer comme V_2 une pentode, ainsi que le montre la figure 2 dont le montage ne diffère de la figure 1 que par l'adjonction d'un potentiomètre R_E avec condensateur de découplage C_2 pour alimenter l'écran de la pentode V_2.

La résistance R_K sera choisie par compromis entre une valeur élevée favorable à la linéarité et une valeur faible favorable à l'amplification de l'étage V_1 qui influence la rapidité du retour.

La variante de la figure 3 dérive de la figure 1 et se rapporte à un générateur à contrôle extérieur de relaxation. Dans ce but, il suffit de couper la liaison entre le condensateur C_1 et l'anode de V_1 et d'amener sur ce condensateur les impulsions négatives de contrôle. La résistance R_C devient dans ce cas superflue et la résistance R_K peut recevoir une valeur élevée, puisque la plaque de V_1 fonctionne à un potentiel plus élevé que dans la figure 1 et l'étage assure une meilleure amplification.

Le montage suivant la figure 4, analogue à la figure 2, fonctionne de façon que la charge du condensateur C soit rapide et que la décharge s'effectue lentement suivant la loi linéaire. Dans ce cas, les rôles des tubes V_1 et V_2 étant permutés, la contre-réaction par la résistance R_K devra être reportée sur le tube V_2. Celui-ci sera constitué par une lampe à forte amplification et à forte résistance interne, par exemple par une pentode, dont le potentiel d'écran est réglé de façon que la lampe travaille au-delà du coude de la caractéristique. De cette façon, on réalise une décharge à courant constant et à allure très sensiblement linéaire.

Le montage de la figure 5 dérive à la fois de ceux des figures 3 et 4, en constituant un générateur à contrôle extérieur de relaxation comme celui de la figure 3 , et en fournissant une dent de scie à montée rapide et à retour linéaire lent, avec utilisation comme V_2 d'une pentode réglée au-dessus du coude de la caractéristique comme dans la figure 4. Pour obtenir le montage de la figure 5, il suffit de couper dans la figure 4 la liaison entre le condensateur C_1 et la plaque de V_1 et d'appliquer à ce condensateur les impulsions de contrôle qui, à l'encontre de la figure 3, devront être positives, de façon que le tube V_2 soit bloqué normalement et qu'il reçoive une tension positive pendant la période de commande en vue de le rendre conducteur. Dans chacun des cas des figures 3 et 5 la durée de balayage est déterminée par la largeur de l'impulsion de contrôle.

L'amplitude de dent de scie peut être réglée en faisant varier la résistance de fuite R_1 ou la tension d'écran prise sur le potentiomètre R_E lorsqu'on utilise une pentode comme tube en shunt. La gamme de fréquences de dents de scie peut être changée comme d'habitude en substituant des valeurs différentes de la capacité C, tandis que la fréquence à l'intérieur d'une gamme donnée peut être ajustée en agissant sur la résistance de contre-réaction R_K, ce qui fait varier le courant et par conséquent la vitesse de charge . La synchronisation pourrait aussi être réalisée en injectant une impulsion positive en série dans le circuit de la résistance de fuite. On remarquera également qu'on pourrait recueillir accessoirement une onde rectangulaire aux bornes d'une quelconque des résistances traversées par le courant constant dans les montages décrits.

- RESUME -

1) - Montage générateur de dents de scie, comportant un tube à vide en série avec le condensateur chargé et un autre tube à vide en shunt sur ce condensateur pour sa décharge, le circuit de sortie du tube shunt étant combiné avec le circuit d'entrée du tube série de façon à exercer sur celui-ci une contre-réaction par son courant de plaque; celui de ces deux tubes dont la conduction correspond à la pente de dent de scie, étant en outre muni d'un circuit de contre-réaction de courant.

2)- Montage ci-dessus dans lequel le tube shunt est une pentode.

3) - Montages suivant 1) et 2)fonctionnant en auto-relaxation.

4)-Montages suivant 1) et 2) fonctionnant avec synchronisation par impulsions extérieures.

5) - Montages suivant 3) ou 4) dans lesquels la charge correspond à la pente de dent de scie et la décharge au retour rapide.

6)- Montages suivant 3) ou 4) dans lesquels la décharge correspond à la pente de dent de scie et la charge à la montée rapide.

7)- Moyens de réglage d'amplitude et de fréquence dans différents montages ci-dessus.

Par procuration de :

HUIT PAGES COMPAGNIE GENERALE DE TELEGRAPHIE SANS FIL

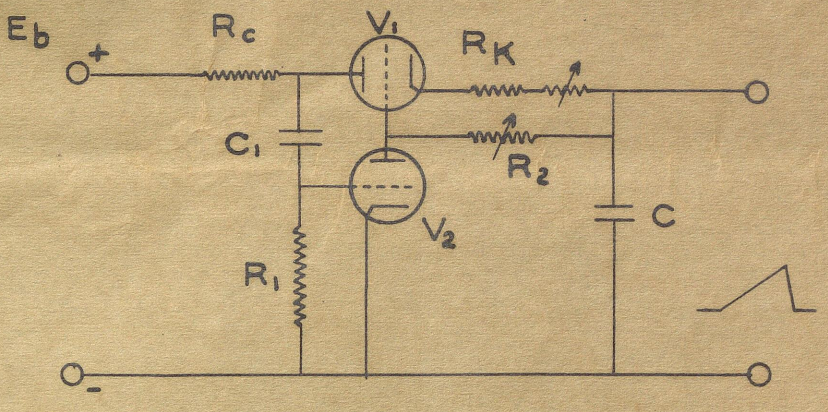


Fig. 1

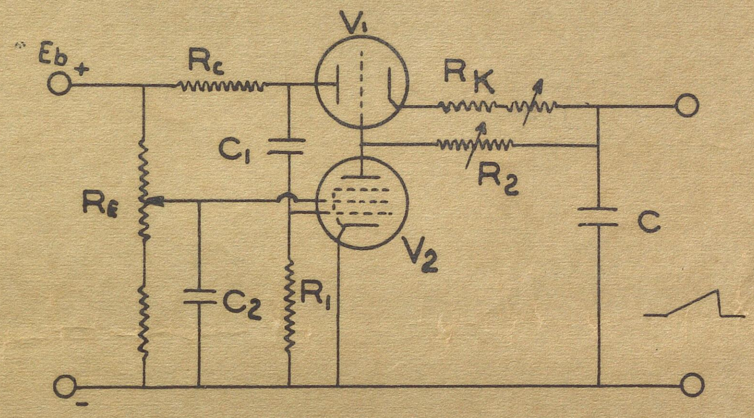


Fig. 2

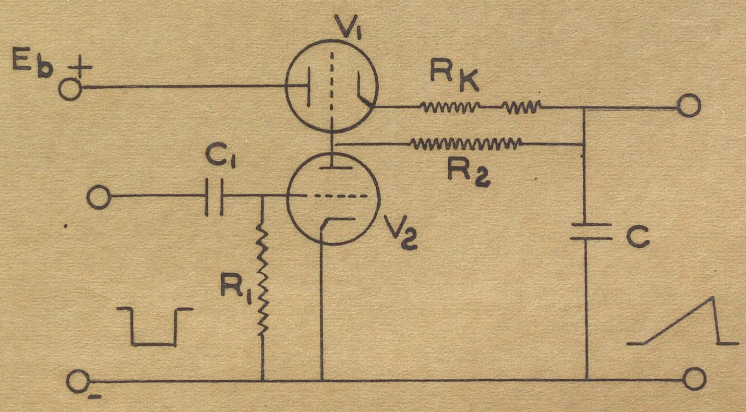


Fig. 3

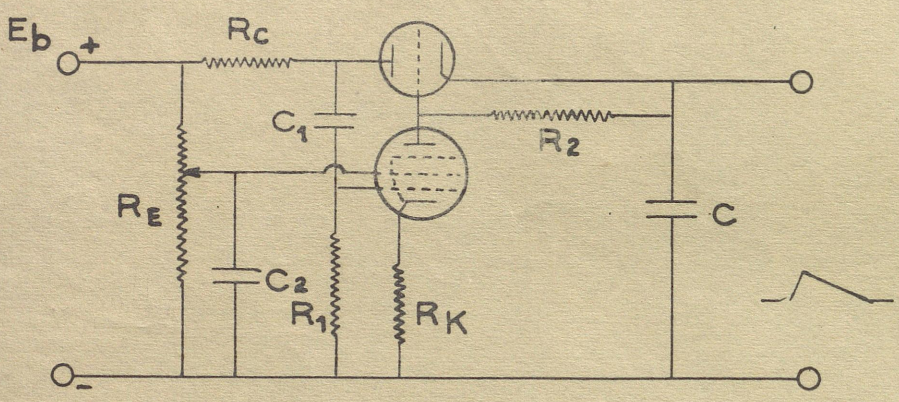


Fig. 4

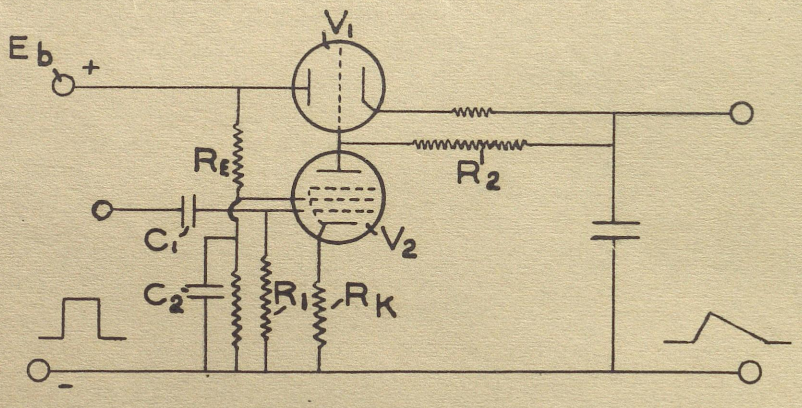


Fig. 5

Cie GÉNÉRALE DE TÉLÉGRAPHIE SANS FIL
CENTRE DE RECHERCHES TECHNIQUES
23, rue du Maroc, PARIS XIXe — TÉL. BOT. 17-06
DÉPARTEMENT ÉLECTRONIQUE

DEPARTEMENT RECHERCHES ELECTRONIQUES. Enreg. Labo : 7.526

Mr. KUNDU -

Mr. REVERDIN.-

MESURE DE L'ANGLE DE PERTES ET DE LA CONSTANTE DIÉLECTRIQUE POUR LONGUEUR D'ONDE CENTIMETRIQUE

-:-:-:-

INTRODUCTION -

Il a été nécessaire de mesurer la constante diélectrique et l'angle de pertes de plusieurs substances utilisables comme éléments atténuateurs dans la bande des ondes centimétriques. Plusieurs méthodes ont été suggérées dans la littérature [1.2]. La méthode employant la mesure d'onde stationnaire dans la ligne de mesure devant l'échantillon diélectrique est la plus pratique pour des substances solides.

La méthode employée ici consiste dans la mesure de la constante diélectrique complexe d'un échantillon par la méthode conventionnelle de la mesure de l'impédence complexe determinaison de la ligne coaxiale, lorsque la substance même est employée comme terminaison de la ligne. La substance est introduite sous forme de disque mince dans la dernière partie du conducteur central de la ligne coaxiale. La constante diélectrique et l'angle de pertes peuvent être déduits du taux d'onde stationnaire dans la ligne et du déplacement du minimum de tension entre la condition de réflection totale et la condition de terminaison avec l'échantillon. Le taux d'onde stationnaire devient très grand pour les substances à faible conductibilité et par conséquent la précision de la mesure n'est pas très grande.

....

THEORIE -

La constante diélectrique complexe d'une substance peut être exprimée de la façon suivante :

$$\varepsilon_c = (\varepsilon' - j\varepsilon'') = \varepsilon'\left(1 - j\frac{\varepsilon''}{\varepsilon'}\right) = \varepsilon'\left(1 - j\frac{\sigma}{\omega\varepsilon'}\right) = \varepsilon_0 k(1 - j\,\mathrm{tg}\,\delta), \quad (1)$$

où ε' = composante réelle de la constante diélectrique = $k\varepsilon_0$

k = capacité inductibe spécifique = constante diélectrique relative.

ε_0 = constante diélectrique dans le vide = 8,854. 10^{-14} farad/cm.

σ = conductivité de la substance.

ε'' = facteur de perte

$\mathrm{tg}\,\delta = \frac{\varepsilon''}{\varepsilon'}$ = quotient de conduction par courant de déplacement dans la substance.

Lorsque $\frac{\varepsilon''}{\varepsilon'}$ est petit, $\mathrm{tg}\,\delta = \sin\delta$ = facteur de perte du diélectrique.

Si y_s est l'admittance de charge complexe de la ligne, on a

$$y_s = y_0(g + ib), \quad (2)$$

où $y_0 = \frac{1}{Z_0}$ = admittance caractéristique de la ligne.

Z_0 = impédance caractéristique de la ligne.

g = conductibilité = $\frac{Z_0}{R}$

b = susceptibilité = $\omega C Z_0$

C = capacité de la terminaison avec échantillon

ω = fréquence

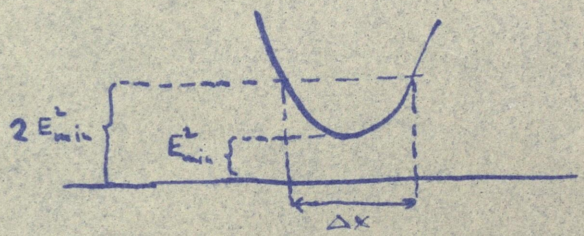
R = résistance shunt équivalente à travers la capacité. /....

Les valeurs de g et de b peuvent être obtenues directement du diagramme de Smith pour les valeurs correspondantes du déplacement du minimum et du coefficient de réflexion.

Lorsque le taux d'onde stationnaire est très grand, il est préférable d'employer une autre méthode[2]. Il est vrai que

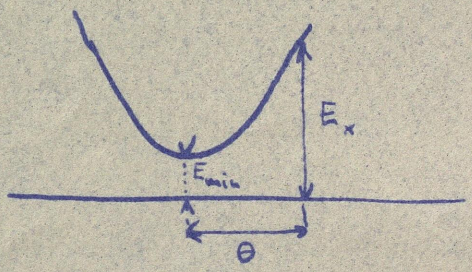
$$\rho = \frac{E_{max}}{E_{min}} = \frac{\lambda}{\pi \Delta x} , \qquad (4)$$

où Δx = largeur de minimum de l'onde stationnaire (voir fig.).



on peut ainsi exprimer le taux d'onde stationnaire par :

$$\rho = \left[\frac{\left(\frac{E_x}{E_{min}}\right)^2 - \cos^2\theta}{\sin^2\theta} \right]^{1/2} \qquad (5)$$



....

E_x = potentiel au point

E_{min} = minimum de potentiel

$\theta = \frac{2\pi}{\lambda} \ell$ = distance le long de la ligne ou degrés entre E_x et E_{min}.

METHODE EXPERIMENTALE.

L'appareil utilisé est montré sur la figure 1. L'impédance du porte échantillon est adapté avec l'impédance caractéristique de la ligne de mesure. On détermine d'abord la position du minimum pour la condition de réflexion totale, lorsque les bouts dela ligne sont court-circuitées en introduisant un disque de leiton de même dimension que l'échantillon. Ensuite ce disque de laiton est remplacé par l'échantillon et on mesure le déplacement de minimum ℓ_r ainsi que le taux d'onde stationnaire $\rho = E_{max}/E_{min}$ qui permet ensuite de calculer le coefficient de réflexion. Le déplacement du minimum est exprimé par ℓ_r/λ. Dans la figure 2 la position de l'onde stationnaire pour une ligne à perte est montrée pour différentes impédances de terminaison.

Le taux d'onde stationnaire peut être calculé directement à partir des lectures du minimum et du maximum de courant de sortie sur le galvanomètre indicateur. On a

$$\rho = \frac{E_{max}}{E_{min}} = \sqrt{\frac{J_{max}}{J_{min}}} \qquad (3)$$

Les grandeurs nécessaires pour le calcul sont :

A = surface du disque (échantillon).

ℓ = épaisseur du disque

C_a = capacité entre les faces de terminaison en l'absence de l'échantillon *

$= 8,854 \frac{A}{\ell} \cdot 10^{-14}$ farad/cm².

lorsque A est mesuré en cm et ℓ en cm.

....

Lorsque C = capacité équivalente de la terminaison avec échantillon

$$k = \frac{C}{C_a}$$

et puisque $b = \omega C Z_o$ on trouve finalement pour la valeur recherchée k :

$$\boxed{k = \frac{b \cdot l}{Z_o \cdot \omega \cdot 8{,}844 \cdot 10^{-14} \cdot A}} \qquad (4)$$

On a aussi $\operatorname{tg} \delta = \frac{\varepsilon''}{\varepsilon'} = \frac{\sigma}{\omega \varepsilon'}$.

Mais $\sigma = \frac{l}{A} \frac{1}{R}$, où R = la résistance shunt équivalente à travers la capacité.

On a $R = \frac{Z_o}{g}$, ce qui donne pour facteur de perte finalement :

$$\boxed{\operatorname{tg} \delta = \frac{l \cdot g}{Z_o \cdot \omega \cdot 8{,}844 \cdot 10^{-14} \cdot A \cdot k}} \qquad (5)$$

....

Valeurs numériques utilisées dans l'expérience :

$Z_0 = 72\ \Omega$

$\lambda = 23,2$ cm

$\omega = \frac{2\pi c}{23,2}$

r = rayon de l'échantillon en (cm).

$$k = 5.85\ \frac{b \cdot l(cm)}{r^2\,(cm^2)} \quad (6)$$

$$\mathrm{tg}\,\delta = 5.85\ \frac{g \cdot l(cm)}{r^2\,(cm^2) \cdot k} = \frac{g}{b}$$

....

RESULTATS EXPERIMENTAUX

8 échantillons de provenances différentes ont été mesurés. Les résultats de mesures sont donnés dans le tableau ci-dessous :

Matériel	ε	$tg\,\delta$	y_s	$R(\Omega)/mm$
Trolitul	2,7	0,009	0.0I + j I.05	∞
Disque A	24,3	0,I3	0.I + j 4.I	∞
Disque B	27,2	0.I3	0.6 + j 5	$300.IO^3$
Disque C	6,5	0.I2	0.3 + j 2.5	∞
Disques D	22,0	0.9I	6 - j 5.5	0
Disques E	2,6	0.0I5	0.8 + j 4	∞
Disques F	30,0	I,2	6 - j 7	5
Disques G	2I,0	0,98	6 + j 5	50

Les mesures ont été effectuées avec une longueur d'onde de 23,2 cm. Dans la littérature on indique pour le trolitul $\varepsilon = 2,5$ et $tg\,\delta = 0,0002$. Il semble donc que les valeurs de ε sont justes et que la méthode donne des valeurs trop fortes pour l'angle de perte.

....

La dernière colonne donne les valeurs de résistance continue des différents matériaux. On reconnait que l'angle de perte est plus élevé lorsque la résistance par cm est plus faible. Disque D est un conducteur tandisque les disques de trolitul, disque A et B sont des isolants. La provenance des disques est connue mais pas leur constitution exacte.

Les disques D et F sont magnétiques. La valeur de ε indiquée n'est plus la constante diélectrique seul mais probablement une fonction de la constante diélectrique et de la perméabilité.

-:-:-:-:-:-

LE 4 JANVIER 1951

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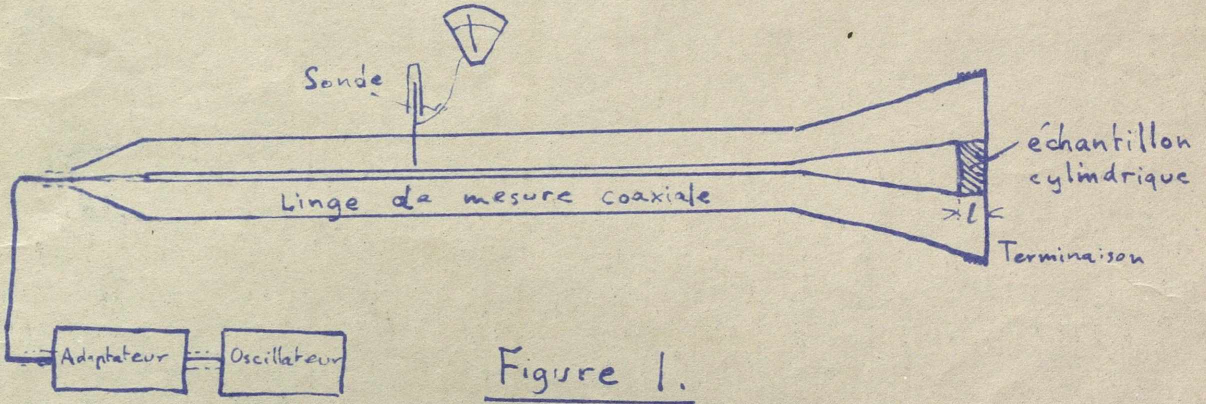

Figure 1.

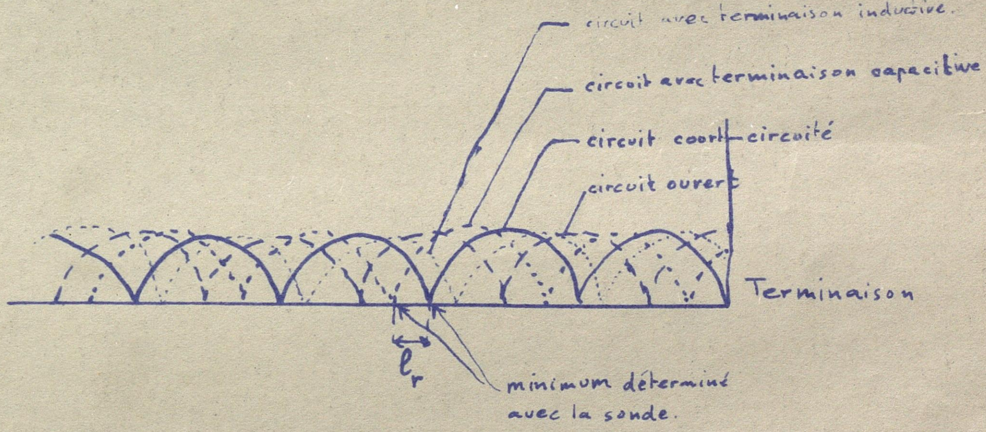

Figure 2.