Nuclear resonance absorption circuit

1959 J. Sci. Instrum. 36 481

(http://iopscience.iop.org/0950-7671/36/12/301)

ORIGINAL CONTRIBUTIONS

Nuclear resonance absorption circuit

By F. N. H. Robinson,* M.A., D.Phil., Clarendon Laboratory, Oxford

[*Paper first received* 13 *April, and in final form* 14 *May*, 1959]

The new circuit described in Part 1 has the sensitivity and flexibility of conventional circuits using a separate oscillator, combined with the convenience and freedom from microphonics of marginal oscillator circuits. In particular, it permits the use of very low levels of oscillation.

Because the factors governing the sensitivity of conventional circuits are not widely known, this first paper is largely devoted to their delineation. Only the principles and theory of the new circuit are given in Part 1. Practical design details are relegated to the second paper.

Part 1. Sensitivity considerations

Nuclear resonance absorption in a specimen surrounded by a coil forming part of a tuned circuit changes the quality factor Q by

$$\delta(1/Q) = 4\pi\eta\chi'' \tag{1}$$

where η is a filling factor and χ'' is the imaginary part of the nuclear susceptibility. The excellence of any circuit for detecting nuclear resonance (n.r.) absorption can therefore be specified by giving that change $\delta(1/Q)$ which yields a signal equal to noise.

The simplest n.r. circuit (Rollin[1]) is shown in Fig. 1, in which a constant current generator I drives the resonant circuit (L, C with loss G) at its natural frequency. N.R. absorption changes the shunt impedance of the circuit and thus the voltage across it. This change is detected by a receiver of bandwidth B and noise figure F.

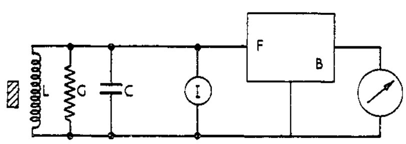

Fig. 1. Simple nuclear resonance circuit using a separate oscillator

A change $\delta(1/Q)$ results in a fractional change $Q\delta(1/Q)$ in the shunt impedance and so, if the r.m.s. voltage in the absence of signal is V_1, it is changed by:

$$\delta V_1 = -V_1 Q\delta(1/Q) \tag{2}$$

The effective r.m.s. noise voltage across the coil is

$$V_n = (4kTFB/G)^{1/2} = (4kTFBQ/\omega c)^{1/2} \tag{3}$$

Thus unity signal-to-noise ratio is achieved when

$$\delta(1/Q) = \frac{1}{V_1}\left(\frac{4kTFB}{Q\omega c}\right)^{1/2} \tag{4}$$

Identical results are obtained for bridge circuits, the only purpose of which is to give the experimenter freedom to minimize the receiver noise figure by working with the optimum input.

The voltage V_1 cannot be indefinitely increased because of nuclear saturation. If the maximum permissible r.f. magnetic field is H_1 then energy considerations yield the following relation

$$V_1^2 C \leqslant H_1^2 U/4\pi\eta \tag{5}$$

where U is the specimen volume.

At the optimum level a value of χ'' satisfying

$$H_1\chi'' = \left(\frac{kTFB}{\pi\omega QU\eta}\right)^{1/2} \tag{6}$$

gives a signal equal to noise. Both H_1 and χ'' are beyond the experimenter's control. To optimize the signal-to-noise ratio one can only increase Q, U, η and decrease F and B. Of these only Q, F, B pertain to the circuit. Factor Q is limited partly by the experimental arrangement and partly by the physical properties of materials. Bandwidth B is limited by the stability of the equipment and by the time available for measurement. Thus, all that the circuit designer can do is to attempt to ensure that F does not appreciably exceed unity.

It is interesting to note that C does not enter into equation (6) directly. The L/C ratio of the circuit is immaterial, except in so far as it affects Q and determines V_1 through equation (3).

This simple arrangement is, in principle, capable of the highest possible sensitivity; however, it suffers from certain well-known disadvantages. In particular, it is inconvenient in that it requires the accurate tracking of the oscillator I and the n.r. circuit; and it is prone to microphonic noise, due to small charges in the capacity of the coil and its leads.

A small change ΔC in capacity produces a spurious signal equivalent to a change $\Delta(1/Q)$ given by

$$\Delta(1/Q) = Q(\Delta C/C)^2 \tag{7a}$$

if the circuit is perfectly tuned; and a spurious signal

$$\Delta(1/Q) = Q(\Delta C_0/C)(\Delta C/C) \tag{7b}$$

if it is mistuned by an amount ΔC_0. In practice, it is difficult to tune a circuit to an accuracy greater than α/Q where α is of the order of $0{\cdot}2$ (this implies maximizing the response of the circuit to within 4%), so that

$$\Delta(1/Q) \simeq 0{\cdot}2(\Delta C/C) \tag{7c}$$

In bridge circuits microphonics are even more serious.

An alternative arrangement without these disadvantages is the marginal oscillator circuit described by Pound and Knight.[2] It is illustrated in Fig. 2. Oscillations are maintained by a non-linear negative conductance $-G'$ provided

* English Electric Research Fellow.

by a valve feedback circuit. The sources of noise have been depicted as a current generator I_n, in which is included the Johnson noise of the circuit, so that I_n has the r.m.s. value

$$I_n = (4kTF_0GB)^{1/2} \tag{8}$$

where F_0 is the noise figure of the feedback circuit.

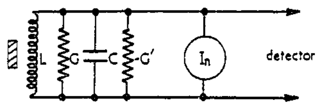

Fig. 2. Basic marginal oscillator circuit

Let the dependence of G' on level be of the form

$$G' = G_0 - G_1V_1 - G_2V_1^2 \ldots \tag{9}$$

then the level of oscillation V_1 is determined by equating the average power delivered by G' to the power dissipated in G.

$$\frac{1}{2\pi}\int_0^{2\pi}(G_0V_1^2\sin^2 x - G_1V_1^3\sin^3 x - G_2V_1^4\sin^4 x)dx = \frac{1}{2\pi}\int_0^{2\pi}GV_1^2\sin^2 x dx$$

Therefore

$$V_1 = \left(4/3\,\frac{G_0 - G}{G_2}\right)^{1/2} \tag{10}$$

Since $G = \omega c/Q$, one has for the signal voltage

$$\delta V_1 = -\tfrac{1}{2}V_1\frac{\omega c}{G_0 - G}\,\delta(1/Q) \tag{11}$$

Provided that the bandwidth B is small compared with ω/Q, which is usually the case, it can be shown[(3)] that, except for a vanishingly small locking region near ω, the circuit behaves towards noise as an admittance $|G_0 - G|$, so that the noise voltage is

$$V_n = I_n/(G_0 - G) \tag{12}$$

and unity signal-to-noise ratio requires a change

$$\delta(1/Q) = \frac{2}{V_1}\frac{I_n}{\omega c} = \frac{2}{V_1}\left(\frac{4kTF_0B}{Q\omega c}\right)^{1/2} \tag{13}$$

This is just twice the detectable change for the circuit of Fig. 1 if $F = F_0$. The factor 2 obviously occurs because in this case the voltage depends on the root of the circuit impedance.

APPLICATION TO CONVENTIONAL CIRCUITS

Now consider the noise figure of two types of marginal oscillator circuit, true negative resistance oscillators and feedback circuits. A typical negative resistance oscillator is the transitron circuit of Knoebel and Hahn[(4)] (Fig. 3). The sources of noise in this case are Johnson noise in G and shot noise in I_2. Since this is a small fraction of the total cathode current, partition ensures that it displays full shot noise.

$$I_n^2 = 4kTGB + 2eI_2B$$

Inserting numerical values for e and k and putting $T = 293°$ K one obtains

$$I_n^2 = 4kTGB(1 + 20I_2/G)$$

or

$$F_0 = 1 + 20I_2/G \tag{14}$$

Since in a typical case $I_2 = 2 \times 10^{-3}$ A and $G = 2 \times 10^{-4}$ mho, one has a value of $F_0 = 201$ and this circuit is very insensitive.

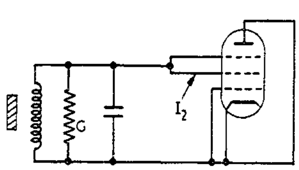

Fig. 3. Skeleton transitron oscillator

In Fig. 4, the structure of a feedback oscillator using a valve of mutual conductance g_m and noise resistance R_n is shown. The oscillation condition is

$$\alpha\beta g_m = G$$

and the noise current generator, in addition to Johnson noise, contains shot noise from the anode current given by

$$I_s^2 = \beta^2 4kTR_nBg_m^2$$

This leads to a noise figure

$$F_0 = 1 + R_nG/\alpha^2 \tag{15}$$

where α is the fraction of the circuit voltage actually applied to the grid. Only in the inductively-coupled tuned grid oscillator is this unity. For the Colpitts and Hartley circuits the value of α is from $\frac{1}{3}$ to $\frac{1}{2}$, in the Pound, Knight, Watkins circuit[(5)] $\alpha \sim \frac{2}{3}$, while in the Clapp circuit $\alpha \sim 0{\cdot}2$.

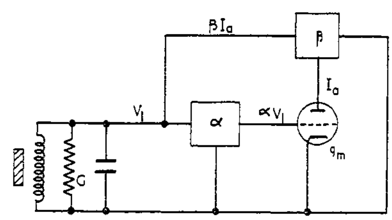

Fig. 4. Basic feedback oscillator

Equation (13) shows that even if F_0 is unity, marginal oscillators are only half as sensitive as separate oscillator systems using the same method of detection. In fact, F_0 is unlikely to be unity except in the most favourable cases. At high values of V_1 either one must use a long grid base valve with a low g_m and high R_n, or reduce α, or the valve is run into grid current, thus introducing additional noise. At low levels of V_1 the valve must be used near cut-off, thus decreasing g_m and increasing R_n. Only when V_1 lies between 100 and 500 mV is it possible to make $F_0 \simeq 1$.

LOW LEVEL OPERATIONS

Suppose that the feedback, or equivalently G, can be adjusted to a fractional accuracy ϵ so that

$$G_0 = G(1 + \epsilon)$$

Then the level is given by

$$V_1^2 = (4\epsilon/3)(G_0/G_2)$$

The negative conductance is proportional to the valve mutual conductance and so:

$$\epsilon = -(3/8g_m)V_1^2(d^2g_m/dV^2)$$

If the triode is interpreted in terms of the usual equivalent diode, the anode voltage of which is $(V_g + V_a/\mu)$ one has

$$I_a = K(V_g + V_a/\mu)^{3/2}$$
$$g_m = 3/2K(V_g + V_a/\mu)^{1/2}$$
$$d^2g_m/dV^2 = -3/8K(V_g + V_a/\mu)^{-3/2}$$

and so

$$\epsilon = (\tfrac{1}{24})(V_1 g_m/I_a)^2 \quad (16)$$

For tubes such as the types 6AK5 and EF91, $g_m/I_a \sim \frac{1}{2}$ at the normal working point increasing to about 1 at a tenth the normal current, while for high gain tubes such as the type E180F the corresponding figures are 1·2 and 2·4. If one takes the latter figure, a level of 10 mV requires an accuracy ϵ of better than twenty-four parts per million. In most practical cases it is the difficulty of maintaining the stability of the circuit components to this accuracy which sets a lower limit to the level, although it should be noted that a more fundamental limit at still lower levels is set by the locking phenomenon mentioned earlier. The region over which the oscillator locks to noise increases as a high negative power of V_1. These considerations apply only to circuits without automatic level control (a.l.c.). Using a.l.c. it is possible to compensate the effect of component instability. Blume[6] has achieved levels of 40 μV using high g_m triodes in a modified Pound–Watkins circuit with a refined a.l.c. system.

In typical circuits the tube providing regeneration is run at a current approaching one-tenth of its normal rating. As a result g_m is reduced and R_n increased five-fold. If one takes as an example a circuit with $G = 1/5000$ mho using a triode-connected type 6AK5 tube, $R_n = 5 \times 500\,\Omega$ and, with $\alpha = \frac{1}{2}$, one has $F_0 = 1 + 2$. Assuming a perfect detector the circuit is $2\sqrt{3} \sim 3{\cdot}5$ times less sensitive than ideal.

COMPARISON OF CIRCUITS

Marginal oscillators are, nevertheless, widely used because of the convenience of having only one frequency to control and their freedom from microphonics. Since the frequency adjusts itself to changes in C, a change ΔC produces a spurious signal given by

$$\delta(1/Q) = (1/Q)(\Delta C/C) \quad (17)$$

Comparing this with equation (7*c*) it can be seen that marginal oscillators are $Q/5$ times (i.e. about ten times) less sensitive to microphonics.

The disadvantages of marginal oscillators are threefold. Firstly, they are less sensitive; secondly, they are difficult to adjust at levels below about 20 mV; and finally they cannot be used with circuits of a low L/C ratio. If the shunt impedance of the tuned circuit is low, not only is the noise figure impaired but also the circuit may fail to oscillate at all.

THE NEW CIRCUIT

All the advantages stem from the use of the same circuit as the n.r. circuit and as the tank circuit of the oscillator. All the disadvantages stem from the use of one tube to provide both regeneration and amplitude limiting. In the circuit illustrated in Fig. 5, these functions are separated. The result is an oscillator with the convenience and freedom from microphonics of a marginal oscillator, yet possessing the sensitivity, range of level and freedom of choice of impedance of separate oscillator circuits.

Ideally the amplifier is linear, broad-band, low noise and with a high value of gain A with a constant phase shift, while the limiter has the characteristic shown in Fig. 6. Given these characteristics, oscillations will occur provided that

$$AV_0/V_i > GR_f \quad (18a)$$

with an amplitude

$$V_1 = V_0/GR_f \quad (18b)$$

Since V_1 depends linearly on $1/G$ and thus on Q, the signal voltage is

$$\delta V_1 = V_1 Q\delta(1/Q) \quad (2)$$

and since no noise is passed by the limiter, the noise voltage at the input as seen by the detector is

$$V_n = (4kTFB/G)^{1/2}$$

Thus unity signal-to-noise ratio occurs when

$$\delta(1/Q) = \frac{1}{V_1}\left(\frac{4kTFB}{Q\omega c}\right)^{1/2} \quad (19)$$

exactly as in Rollin's simple circuit (Fig. 1).

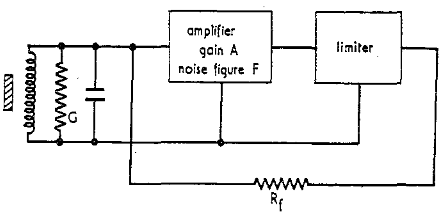

Fig. 5. Limited self-oscillator

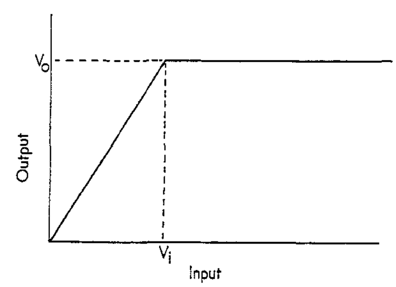

Fig. 6. Ideal limiter characteristic

Practical limiters are not ideal. They also generate noise. Let the limiter output be V_3 for an input V_2, then the oscillation condition becomes

$$AV_3/GRV_2 = 1 \quad (20a)$$

A noise voltage V_n at the amplifier input is fed back as a noise voltage

$$V_n' = V_n(A/GR)(\partial V_3/\partial V_2) \quad (20b)$$

Thus

$$V_n'/V_n = (V_2/V_3)(\partial V_3/\partial V_2) \quad (20c)$$

With even very imperfect limiters, this quantity is small, less than $\frac{1}{3}$. In practice, it can be made very much less than this with little difficulty.

The most useful practical limiter is the saturated pentode. In Part 2 of this paper it is shown that the output at the input frequency is approximately equal to the mean anode current. Let this be I_0 passing through a load resistance R_L and displaying full shot noise. The noise fed into the input circuit is

$$I_L^2 = 2eI_0BR_L^2/R^2$$

The available r.f. voltage output is

$$V_L = I_0 R_L$$

and the input level

$$V_1 = V_L / GR$$

Combining these equations and comparing the noise current with Johnson noise in G, one obtains an oscillator noise figure

$$F_0 - 1 = 20(V_1^2 G / I_0) \quad (21)$$

With $I_0 = 5 \times 10^{-3}$ A

$$F_0 - 1 \simeq 4000\, V_1^2 G \quad (22)$$

Unless one wishes to combine high level operation with a low circuit impedance, this is negligible. With $G = 1/5000$ and $V_1 = 0 \cdot 5, F_0 - 1 = 0 \cdot 2$. At lower levels it will certainly be negligible.

Thus, apart from a combination of extreme circumstances, the imperfections of practical limiters have little effect on the sensitivity. A direct experimental comparison of a circuit of this type with a separate oscillator circuit using the same coil, specimen, etc., and a level of 10 mV revealed no difference in signal-to-noise ratio, apart from the complete absence of microphonics with the limited oscillator.

Because limiting is done at high level there is, in principle, no lower limit to the level at the n.r. coil. In practice, stable oscillations with only 100 μV across the coil can be achieved without using automatic level control and without critical adjustments. Whereas in a conventional marginal oscillator circuit the level is set by a delicate balance between circuit loss and regeneration, here the level depends on these quantities in a linear way, so that small changes in the regeneration do not have a disproportionate effect on the level. Since low levels are easily achieved and since ample regeneration is available, circuits having a very low L/C can be used without either saturating the resonance or completely killing the oscillations. Thus, for example, the ^{7}Li resonance at 3·5 Mc/s has been observed in a specimen of lithium fluoride having a relaxation time of 30 s at 4·2° K, using a coil tuned with 5000 pF having a shunt impedance of approximately 400 Ω. This latitude in the L/C ratio of the circuit is a great convenience in all but the simplest experimental situations.

The above theoretical considerations indicate that this type of circuit is capable of the highest sensitivity. The practical results obtained with the circuits described in the succeeding paper show that this sensitivity is easily achieved and not vitiated by microphonics or component instability.

REFERENCES

(1) ROLLIN, B. V. *Nature* (*London*), **158**, p. 669 (1946).
(2) POUND, R. V., and KNIGHT, W. D. *Rev. Sci. Instrum.*, **21**, p. 219 (1950).
(3) VAN DER ZIEL, A. *Noise* (London: Chapman and Hall Ltd., 1955).
(4) KNOEBEL, H. W., and HAHN, E. L. *Rev. Sci. Instrum.*, **22**, p. 904 (1951).
(5) POUND, R. V. *Progr. Nuclear Phys.*, **2**, p. 21 (1952).
(6) BLUME, R. J. *Rev. Sci. Instrum.*, **29**, p. 574 (1958).

Part 2. Practical design

THE LIMITER

The design hinges on the performance of the limiter. Conventional limiters used in f.m. circuits use either biased diodes or saturated pentodes. Both types of limiter are adequate for the present purpose, but pentode limiters economize in components. Therefore discussion of diode limiters has been omitted.

The basic pentode circuit is shown in Fig. 7. A radio-frequency voltage V (peak) from a source of impedance R_s

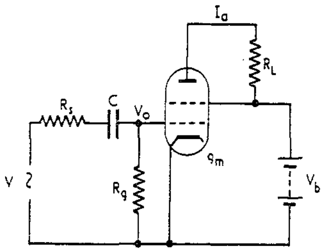

Fig. 7. Pentode limiter circuit

is rectified at the signal grid. The steady negative voltage V_0 developed across R_g may be shown to be

$$V_0 = V \cos \theta \quad (1)$$

where

$$\tan \theta - \theta = \pi R_s / R_g \quad (2)$$

and 2 θ is the phase angle during which grid current flows. The table gives numerical results.

Numerical results of basic pentode circuit

2 θ (deg.)	60	90	120	150	160
R_s/R_g	0·016	0·07	0·22	0·8	1
V_0/V	0·866	0·707	0·5	0·26	0·17

In Fig. 8 the behaviour of the grid voltage and anode current during a complete cycle is illustrated. In this diagram V_c

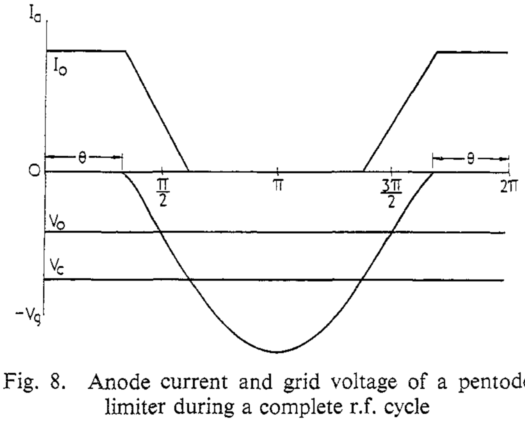

Fig. 8. Anode current and grid voltage of a pentode limiter during a complete r.f. cycle

denotes the cut-off grid voltage at which anode current ceases. For a fixed value of R_s/R_g the form of the diagram depends only on the ratio V/V_c. If θ is near to $\pi/2$ and $V/V_c \gg 1$, the anode current consists of square pulses each lasting approximately half a cycle and having a peak value I_0 equal to the anode current at zero bias. The mean current $\bar{I}$ and the fundamental component I_1 are then given by

$$I_1 = (4/\pi)\bar{I} = (2/\pi)I_0 \qquad (3)$$

Under these conditions, i.e. $\theta \simeq \pi/2$, $R_s/R_g \gg 1$, $V_0 \ll V$ and $V \gg V_c$, the output at the input frequency is independent of the input V. If θ is reduced by increasing R_g limiting occurs at lower values of V, e.g. when $V \sim V_c$, but at higher values of V the limiting action is over-effective and the output decreases with increasing input. In practice, measurements show that satisfactory limiting over a range of inputs from

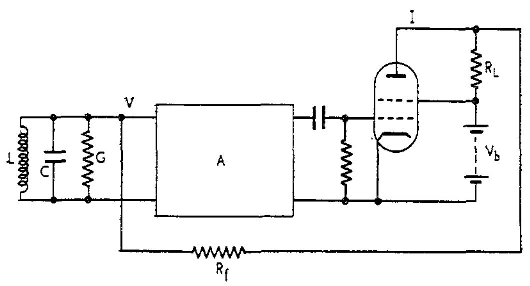

Fig. 9. Self-oscillator with pentode limiter

$\frac{1}{2}V_c < V < 4V_c$ is obtained with $R_s/R_g \simeq 0{\cdot}2$ and $V_0/V \simeq 0{\cdot}5, 2\,\theta \simeq 120$ deg. A detailed analysis of the circuit is possible but tedious, and for this purpose unnecessary since only a very moderate limiting action is required.

The cut-off voltage of short grid base pentodes, such as the type 6AK5, is of the order of 4 V with the normal screen voltage (150 V). By reducing this to 25 V or less, V_c can be reduced to about 0·5 V.

GENERAL DESIGN

Suppose, then, that one wishes to build an oscillator to operate at a level V volts. Satisfactory limiting requires an input of the order of V_c volts at the limiter. The gain A of the preceding amplifier must therefore be:

$$A = V_c/V \qquad (4)$$

Now consider the circuit shown in Fig. 9. Let A be chosen so that at the lowest desired level of oscillation V', $AV' = V_c'$ where V_c' is the cut-off voltage at a supply voltage V_b', and let R_f be adjusted to maintain oscillations at this level. Then if V_b is increased to $V_b'' = aV_b'$, V_c is increased in the same ratio. The peak anode current and the output of the limiter are also increased. If this, too, increases in the same ratio a the whole circuit will oscillate at a level $a\,V'$. Conditions at the limiter will remain unchanged. In practice, the limiter output increases at a greater rate than a and so the level rises more than a times and the voltage at the limiter grid exceeds V_c''. Fortunately, adequate limiting occurs over an eight-to-one ratio of input to cut-off voltage and the circuit will accommodate this change.

If R_f is chosen so that, at the lowest desired level and the lowest possible supply voltage V_b, the level at the limiter is of the order of $\frac{1}{2}V_c'$, then approximately a 20 : 1 change in level can be effected simply by increasing V_b. The adjustment of R_f is most easily carried out by making R_f equal to one-third of the value which just sustains oscillations. An error of 2 : 1 in R_f makes little difference to the performance.

Since V_c' is of the order of 0·5 V, the gain of the amplifier must be approximately 25 if levels between 20 and 400 mV are required, and 1000 if levels down to $\frac{1}{2}$ mV are required. The oscillator must contain an even number of valves (including the limiter) to preserve the correct phase of the fed-back voltage. With modern pentodes a gain bandwidth product of approximately 200 Mc/s is possible. Thus a three-stage amplifier will give a gain of 1000 over a band of 20 Mc/s, while a single-stage will give a gain of 25 over a band of 8 Mc/s. If it is feasible to provide some sort of coarse tuning of the interstage circuits these bandwidth figures can, of course, be increased.

TWO PRACTICAL CIRCUITS

The precise form of the circuit depends on the particular application. The two circuits described here were designed: (1) to give levels between 10 and 200 mV at frequencies between 20 and 60 Mc/s with some coarse tuning of the inter-stage circuit; and (2) to give levels between 1 and 250 mV at frequencies from 200 kc/s to 20 Mc/s with no tuning adjustments.

The first circuit is shown in Fig. 10. The first stage V_1 uses a type E180F pentode having a mutual conductance 16·5 mA/V with input and output capacities of 11·5 and 3·5 pF. The limiter V_2 uses a type 6AK5 (chosen primarily

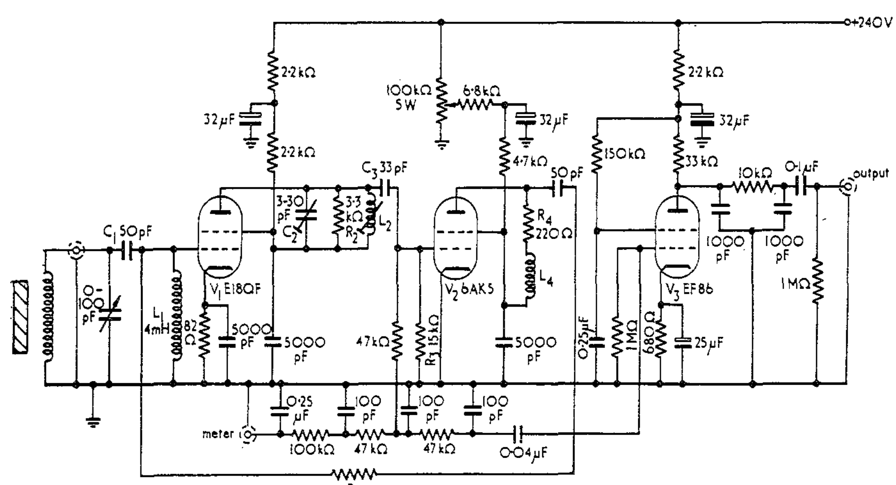

Fig. 10. Tunable medium level limited oscillator

for its low input capacity of 4 pF). The anode load of V_1 is tuned by L_2 (which is permeability tuned) and C_2 (3 – 30 pF). With C_2 at its minimum setting, a gain of 50 over a 4 Mc/s band is achieved without retuning; L_2 is wound to resonate with $C_2 = 3$ pF, together with the valve capacities (8 pF) at 63 Mc/s at its minimum setting. With C_2 and L_2 at maximum, the resonant frequency is 17 Mc/s. A coverage of 15 to 65 Mc/s is thus achieved.

The capacitor C_1 (50 pF) and the r.f. choke L_1 (4 mH) remove direct field modulation pick-up. The choke is wound astatically, i.e. it consists of two well separated pies (at the opposite ends of a $\frac{1}{2}$ W resistor) connected in opposition. This minimizes direct pick-up in the choke itself.

The grid leak time constant C_3 (33 pF), R_3 (15 kΩ) is chosen low enough to prevent squegging (discontinuous bursts of oscillation) at the highest frequency. If this is f and the coil has a quality factor Q then, roughly speaking,

$$fCR < Q \qquad (5)$$

Since the signal is rectified at the limiter grid it seems only sensible to use the rectified signal thus produced both as the output and to monitor the level. The network shown filters out the r.f. without increasing the grid leak time constant. A level of 20 mV provides a rectified voltage of $50 \times 0{\cdot}02 \times \frac{1}{2} = \frac{1}{2}$ V and thus gives a meter current of 2·5 μA. The final stage V_3 is a conventional low-frequency amplifier included on the same chassis to minimize pick-up in the leads.

The limiter grid current flows through R_3. If one assumes that this displays full shot noise, the resulting noise voltage across R_3 is

$$V_n^2 = 2eI_gR_3^2B = 2eV_0R_3B$$

In terms of Johnson noise this is

$$V_n^2 = 20V_0 \,.\, 4kTR_3B$$

The ratio of this noise voltage to amplified Johnson noise from the tuned circuit is (allowing for 50% detection efficiency)

$$F^1 = 80V_0R_3/A^2Z \qquad (6)$$

where A is the amplifier gain and Z the impedance of the tuned circuit. With $V_0 = 3$, $R_g = 15 \times 10^3$, $A = 50$, $Z = 5 \times 10^3$, this is unity. This value of V_0 corresponds to a level of 120 mV, so that noise from this source is important at the higher levels. At 200 mV the overall noise figure is 3, while at 10 mV it approaches unity. This deterioration in sensitivity at high levels could, of course, be avoided by using a separate diode detector.

The anode load of the limiter R_4, L_4 calls for some comment. It is unnecessary for the limiter to provide gain. The load R_4 is therefore made low (220 Ω) to give a wide frequency range. The inductance L_4, which tunes with the valve capacity at 110 Mc/s, is included solely to provide phase compensation at the higher frequencies. Direct current is used for the heater supply.

The second circuit (Fig. 11) uses a three-stage amplifier. The audio-stage is omitted from the diagram. The first three stages again use type E180F valves chosen for their high gain bandwidth product. The limiter V_4 is a type 6AK5. The choke L_1 is 10 mH, again astatically wound, and the limiter grid leak time constant is chosen to avoid squegging at 20 Mc/s.

The arrangements for monitoring the level and taking off the detected signal remain as before, but now, because of the increased gain, grid current noise is unimportant. The loads of the first three stages are tuned with the valve capacities to resonate at 28 Mc/s with a Q of $1/\sqrt{2}$. This requires $R_2 = R_3 = 560$ Ω, $L_2 = L_3 = 2$ μH since the circuit capacity is 15 pF. The input capacity of the limiter is less (4) than that of the amplifier tubes (11·5) and so $L_4 = 4$ μH and $R_4 = 1$ kΩ ($C_4 = 8$ pF).

A circuit resonant at f_0 with $Q = 1/\sqrt{2}$ has the property that its impedance is sensibly constant up to $f_0/\sqrt{2}$. The gain of the amplifier is therefore constant at 1400 up to 20 Mc/s. Measurements show a variation of ±20%. The phase of the r.f. voltage, however, is shifted as the frequency increases. The shift reaches −110 deg. at 20 Mc/s. This is partially compensated by tuning the limiter load at 28 Mc/s with a Q of 3. The remaining phase shift is further reduced by the use of the R_fC_f feedback network, chosen so that $R_fC_f \simeq 1$ at 20 Mc/s. With $R_f = 1$ MΩ this requires $C_f = 0{\cdot}01$ pF, which is obtained by placing two wires in proximity to each other.

The lower frequency limit is determined by the interstage coupling capacities; 200 kc/s was chosen rather arbitrarily as the lowest frequency likely to be of interest.

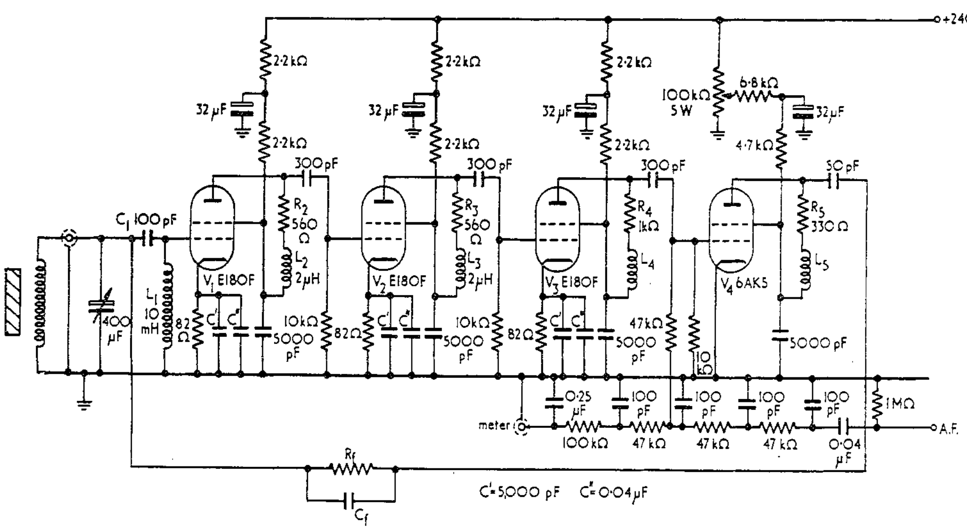

Fig. 11. Broad band low level limited oscillator

An oscillation level of 1 mV produces 1 V rectified at the limiter grid and 5 μA meter current. This can be checked by connecting a low-impedance signal generator in place of the resonance coil. Measurements of the limiter output show that with varying supply voltages, satisfactory limiting is achieved when the meter current lies between 2·5 and 50 μA. At a fixed supply voltage the range of tolerable limiting encompasses a range of 5 : 1 in input level. The procedure for adjusting R_f is therefore to choose a value giving 2·5 μA at the lowest limiter supply voltage; C_f is then trimmed if the frequency is above 15 Mc/s to ensure that oscillation can be maintained at 20 Mc/s. Too small a value of C_f is indicated if the oscillations become excessively noisy, or if changes in the level setting pull the frequency excessively.

The noise resistance of a pentode-connected type E180F tube is 500 Ω. The noise figure therefore increases if the circuit impedance is below 1000 Ω. However, this problem is common to all n.r. circuits. By reducing R_f, circuits having impedances of only 100 Ω can be made to oscillate, and it is possible to cover 18 – 3 Mc/s using the same coil, merely by adding tuning capacity.

RESULTS

No attempt has been made to measure the sensitivity of circuit 1, but a direct comparison of circuit 2 has been made with a separate oscillator circuit. The test was made at 5 Mc/s using the ^{7}Li signal from lithium fluoride at 4·2° K. The level was set at 10 μA (2 mV) and the signal-to-noise ratio estimated on an oscilloscope using a 50 c/s, 30 G sweep. The feedback resistor R_f was then removed and a standard signal generator connected through a high resistance to the input. The output was adjusted to give the same meter reading and the signal-to-noise ratio again measured. The signal and noise were the same (ratio approximately 5 : 1) within the accuracy of the measurement (50%), although even with the most careful tuning of the generator, the second observation was rendered extremely difficult because of microphonics. These were completely absent using the self-oscillator. This measurement shows that the oscillator circuit in itself does not introduce noise. In both cases any noise in excess of Johnson noise was solely due to the amplifier noise figure.

The circuit appears to be remarkably free from microphonics. In one experiment the whole apparatus was visibly shaking due to vibration from a pump, but the signal remained perfectly steady.

A photograph of the complete circuit is shown in Fig. 12 with the outer case and base plate removed. All the components are mounted on a single circular chassis and grouped around the tuning capacitor which is enclosed in the central screening can. Standard miniature i.f. transformer formers are used for the inductances L_2, L_3, L_4, L_5 and the r.f. leads are kept short by taking all d.c. leads, such as heater and screen supplies, directly through the chassis and out of the way at the valve bases. Insulated ½ W resistors make excellent insulated lead-throughs. Apart from careful location and orientation of the valve sockets, no attempt is made at screening or heater filtering; nevertheless, the circuit is stable with R_f removed. The tuning dial and sockets are located on the top plate (also circular) and the whole assembly fits inside a cylindrical can and is completed by screwing on a base plate. This carries a clamp for securing the lead to the n.r. coil.

The circuit as it stands covers levels from less than 1 to 15 mV. At the latter level the third amplifier tube overloads. The range can be extended upwards by five times by removing the cathode bypass capacitors on V_3. The gain is reduced by 2·3. A further increase in level can be made by performing the same office for V_2 and then V_1. With all bypass capacitors removed the overall gain is approximately 100 and the maximum level approximately 300 mV. Of course, adjustment of R_f must be made at the same time.

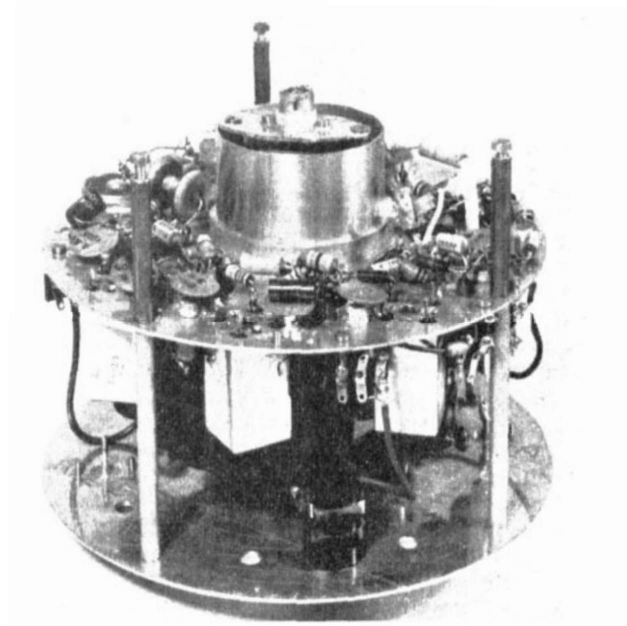

Fig. 12. The broad band, low level circuit of Fig. 11. The first r.f. stage is on the left below the non-inductive r.f. choke. The limiter is on the right below the length of stripped coaxial line forming C_f

It is perhaps worth remarking that the circuit contains no more components than are necessary to build a receiver to detect n.r. signals at the same level. None of the component values are critical, with the possible exception of the phase-correcting circuits L_5 and C_f at frequencies above 15 Mc/s. If it oscillates and if the limiter voltage controls the level, the sensitivity is little improved by adjusting R_f.

NOTES ON THE COMPONENTS

The tuning capacitor is Jackson Bros 200 + 176 pF connected in parallel. The astatic chokes are made by removing two pies from standard Bulgin short wave chokes and mounting them on 1 MΩ, ¼ W resistors. All bypass capacitors, with the exception of the 32 μF electrolytics and the 0·04 μF metallized paper cathode bypasses, are ceramic. All resistors are ½ W insulated carbon, except those in the network leading from the limiter grid, the anode loads and R_f, which are ¼ W uninsulated. All the valves, except the self-screened audio-stage, have screening cans. The i.f. coil formers are 0·3 in. diameter, 1 in. long in ⅞ in. square cans, and have permeability tuning slugs. A standard Belling–Lee television coaxial plug and socket are used for the connexion to the n.r. coil.