# Wireless World

**RADIO AND ELECTRONICS**

APRIL 1949

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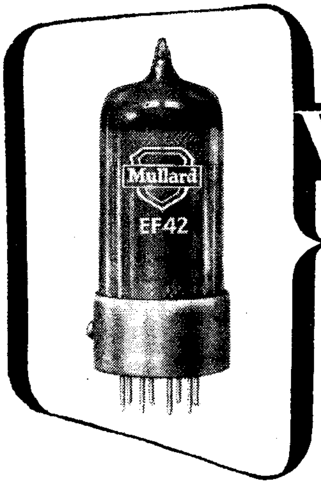


# *Valves and their applications*

## THE EF42 IN THE INPUT STAGE OF A WIDE-BAND OSCILLOGRAPH AMPLIFIER

The input stage of an oscillograph is generally less difficult to design than are the output stages as the operating conditions are rather less stringent. The inter-electrode capacitances appearing on the anode of the stage are rather lower in the former case so that a satisfactory frequency response is more easily obtained; in addition the fact that only relatively small voltages are required to drive the output stages means that the valve can be run under much lower current conditions.

For these reasons it is convenient to arrange for the input stage of the amplifier to provide as much gain as possible so that the overall sensitivity of the oscillograph shall be high.

A single valve circuit suitable for feeding the output stages*, is illustrated in Fig. 1. As can be seen, cathode compensation has again been used, and the condenser $C_k$ can conveniently be 200-400pF. when the anode load $R_L$ is 4.7 K$\Omega$.

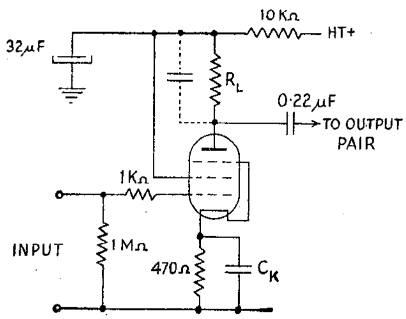


FIG. 1

On the whole, this method of operation is rather wasteful since so much of the amplification is thrown away at low frequencies in order to boost the H.F. response, and a rather better solution can be obtained if the effects of the stray capacitances to earth at the anode of the stage can be reduced. In the case of a complete oscillograph amplifier this can be done by feeding a signal from the anode of the appropriate output valve of the push-pull pair via a very small capacitance (in this case about 2pF) to the anode of the first valve, when, by positive feedback, the effective capacitance to earth appearing across the anode circuit is reduced, but only up to the frequencies amplified by the output stage. When this is carried out, $C_k$ can be increased to 500$\mu$F, thus giving full gain at low frequencies, and $R_L$ can be as high as 10K$\Omega$. The complete response curve of such a practical amplifier - C.R.T. combination is illustrated in Fig. 2; curves 1 and 2 are obtained respectively without and with the positive feedback network. The sensitivity is such that a 15mV signal (peak to peak) gives a trace of 1 cm. on the tube, while the transient response gives only a 5 per cent. "overshoot" to a square wave of 0.2 micro-secs rise time.

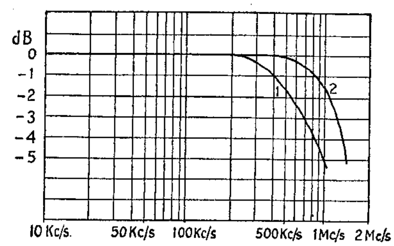


FIG. 2

In practice, some compensation is necessary for the phase delay introduced in the amplifier as this makes the positive feedback network operate at low efficiency at some frequencies but the simple correcting network illustrated in Fig. 3 is found to be quite effective.

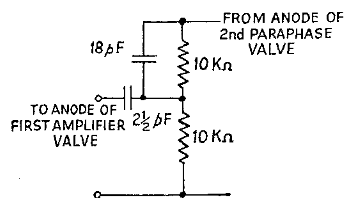


FIG. 3

* See "The EF42 in the Output Stage of Wide-Band Oscillograph Amplifier." — *Wireless World*, January, 1949.

**Mullard**

*Reprints of this report from the Mullard Laboratories, together with circuit diagram of the input stage and feedback network, can be obtained free of charge from the address below.*

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# Wireless World

RADIO AND ELECTRONICS

APRIL 1949

## *Comments of the Month*

### PROGRAMME—COMPLETE WITH INTERFERENCE

WHEN so much attention is being paid to interference in television reception, we think that more notice should be taken of a form of interference which is apparently being radiated from Alexandra Palace itself along with the programme! Few viewers can have failed to notice a pattern of vertical or near-vertical bars which appear as a background to the picture. Sometimes the bars are steady, but more often they are continually varying in position and angle.

Were it not for one fact, we should hardly have the temerity to suggest that the B.B.C. could be to blame, but there seems to be conclusive evidence that this is so. The interference appears only when a particular camera is in use. When several cameras are employed to give different angles of view, as in the transmission of a play, it is most evident that the interference occurs only with one of them. The trouble disappears instantly whenever a change is made to one of the others.

The interference varies in intensity from time to time, but was extremely bad during the play "And so to Bed" on 6th March, and greatly detracted from the entertainment value of the production. It has been noticeable now for some eighteen months and while we do not doubt that the B.B.C. is aware of it and has attempted to cure it, we do feel that it is high time that some more active measures were taken.

The pattern appears to be produced by c.w. interference of a frequency in the neighbourhood of 1 Mc/s and this agrees with the suggestion which we have heard that it is actually due to pick-up of the Brookmans Park Home Service signal on the camera circuits. Certainly the appearance of the interference coincided, as far as we can remember, with the bringing into service of the new aerial system at Brookmans Park.

If this suggestion be correct, we think that the B.B.C. should take viewers into their confidence. Uncertainty and confusion as to the nature of this particular form of interference should be dispelled as quickly as possible. Until the trouble is overcome, use of the affected camera channel should cease. At present it must be doing considerable harm to television as in many areas it is far more noticeable than ignition interference.

### WHAT IS A COMPONENT ?

COMING so soon after the impressive and highly successful annual exhibition recently staged by the Radio Component Manufacturers' Federation, the raising of this question is perhaps rather ungracious. But the task of those whose duty it is to record happenings in the world of wireless is not being made any easier by the inclusion of more and more complex devices under the general classification of components.

To say that components should be defined as "devices represented by a single symbol on a circuit diagram" is obviously an over-simplification. According to that, a built-in loudspeaker could properly be described as a component but, by ordinary usage, an extension speaker certainly could not. Such things—and valves as well, for that matter—are commonly described as "accessories." That word, in its turn, we have heard defined rather fancifully as "devices capable of useful separate existence." As we understand this definition, a normal plug-in valve would be an accessory, while a wire-end miniature, of the kind used in hearing aids, would be a component.

All this is becoming very difficult and confusing. Clearly enough, there is so much room for differences of opinion that any classification—if indeed we need one—must come from some body endowed with dictatorial powers.

# SINGLE-VALVE FREQUENCY-

## New Principle Giving Wide Coverage

**By K. C. JOHNSON, B.A.**

IN order to vary the frequency of an oscillator it is generally necessary to change the resonant frequency of a tuned circuit, and there are many applications, such as "wobbulators," a.f.c. systems, or f.m. signal generators, where this has to be done electronically. Usually, for these devices, the well-known "reactance-valve" arrangement has been used, in which a resistance-capacitance phase-shift network is connected between the anode and the grid of a variable-mu pentode. A phase-shift between the anode current and the anode voltage is thus introduced, so that the anode impedance is effectively reactive and, moreover, the value of the reactance depends on the slope and hence the bias of the valve. If this reactance is used as part of the tuned circuit of a conventional oscillator, it is possible to modulate the oscillation frequency by means of the bias voltage applied to the reactance valve.

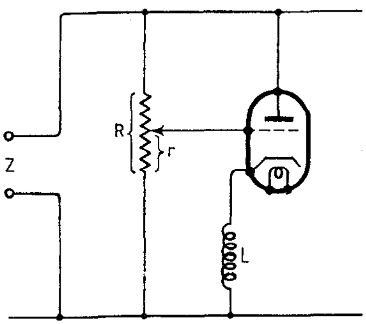



Fig. 1. Indicates how an ideal cathode-follower may be used to obtain a large effective inductance $Z = L\left(\frac{R}{r}\right)$.

This arrangement, however, cannot be made to give a very wide frequency coverage, for the phase-shift network must be designed to give a shift of as nearly as possible 90° throughout the range, so as to avoid variations in the oscillation amplitude as well as the frequency. This means that there must be heavy attenuation in the phase-shifting network, so that the coverage can only be small even at low frequencies, whilst at high frequencies the valve input capacitance becomes serious, making the network design difficult and the coverage extremely small. Lastly, there is the practical consideration that although a reactance valve must necessarily be separate from an oscillator, the arrangement to be described allows the two functions to be combined and a single valve to be used.

**Modulation Principle.**—The alternative method of "direct reactance modulation," which can give constant amplitude over a wide frequency range, does not use phase-shift networks at all, but depends on the principle* that the effective value of any impedance can be altered simply by arranging that although the current flows unchanged, the voltage actually applied to the impedance is only a definite fraction of the whole. Fig. 1 indicates how this principle can be used to obtain a large effective inductance, for if the valve is considered as an ideal cathode-follower, then the voltage swing actually applied to L is only $\frac{r}{R}$ of the swing across the terminals, but the whole of the current swing in L flows in at the terminals, so that they appear as an inductance taking less current, that is as a larger inductance, than L.

But valves are more suitable for dividing currents than voltages, since the suppressor grid of a pentode divides the cathode current between the anode and the screen in a ratio depending directly on the suppressor voltage and which can therefore be varied electronically; fortunately the principle holds just as well if it is the current which is divided instead of the voltage. That is to say that the effective value of any impedance can be altered by arranging that although the voltage is applied unchanged, the current actually passing through the impedance is only a definite fraction of the whole. It is not possible to give a simple circuit to illustrate this, but it can be used to obtain electronic modulation of impedances over wide ranges.

A particularly useful variation of this current division principle is possible with inductances, however, using the property of mutual inductance, for if L is the inductance of a coil carrying an alternating current $i_o \sin pt$, and a fraction $\frac{1}{x}$ of this same current flows in an ancilliary winding with a mutual inductance M between the two, then the voltage across the first coil is $(i_o p \text{ L} \cos pt + i_o p \frac{\text{M}}{x} \cos pt)$ Thus the effective inductance of the first coil is $\text{L} + \frac{\text{M}}{x}$, and it is possible to arrange that M is negative by winding so that the current flows round in opposite directions in the two coils. If the second coil is wound with more turns than the first and the flux leakage kept small, it is even possible to have M greater than L so that the effective inductance can be reduced from L right down to zero as the division fraction $\frac{1}{x}$ is increased.

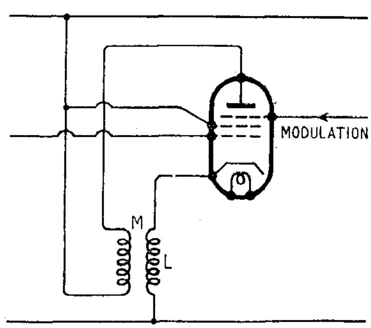



Fig. 2. The effective cathode load of this valve is $\left(\text{L} + \text{M}\frac{\text{I}_a}{\text{I}_k}\right)$ and it can be modulated by the suppressor grid voltage.

Fig 2 shows how this principle

* Prov. Pat. applied for.

# MODULATED OSCILLATORS

can be used to obtain a cathode load in a valve circuit whose effective value is always inductive but can be varied in magnitude electronically. The first coil, with self-inductance L, is placed in the cathode lead and carries the whole alternating cathode current $I_k$, while the second coil, with a

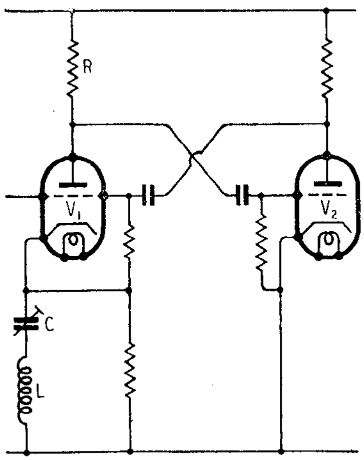



**Fig. 3. An oscillator in which the tuned circuit elements are in series with a valve.**

mutual inductance of M to the first, carries only the anode current $I_a$. But the action of the suppressor grid in a pentode is to divide the cathode current, between the anode and the screen in a ratio which depends on the suppressor voltage, but not on the cathode current itself, so that the effective cathode lead inductance is $\left(L + M\dfrac{I_a}{I_k}\right)$, where $\dfrac{I_a}{I_k}$ is the fraction of the total current which flows to the anode and so through the second coil.

**Oscillator Circuits.**—In order to use this principle to modulate the frequency of a practical oscillator, it is necessary to devise a circuit in which the actual oscillatory current flows through a valve in series with the tuned circuit elements. Fortunately this is easier than it might appear, since to make a "series tuned circuit" oscillate a voltage must be forced across it proportional to the current flowing (see E. W. Herold "Negative Resistance," *Proc. I.R.E.*, Oct., 1935) and this is most easily arranged by the circuit shown in Fig. 3. Here the resistance R develops a voltage proportional to the current flowing in L and C, which is phase-inverted by the second valve and applied to the tuned circuit by cathode-follower action in the first valve, so as to increase the current and excite oscillations.

But the second valve in this circuit is merely serving to invert the voltage, like the second valve in the familiar Franklin oscillator, and it can be quite satisfactorily replaced by a phase-inverting auto-transformer to give the circuit of Fig. 4 which is otherwise similar to Fig. 3. This transformer consists of a coil tuned to the central oscillation frequency and tapped so as to give a small gain, but heavily damped by a shunt resistance so that it cannot introduce undesirable phase-shifts, so tending to control the oscillation frequency, and reduce the range of frequency modulation obtainable.

**The Completed Oscillator.**—The completed frequency-modulated oscillator combines the mutual inductance modulation principle shown in Fig. 2 with either of the oscillator circuits of Figs. 3 and 4. The two valve circuit is slightly easier to make up, in practice, but the more economical single-valve circuit is shown in Fig. 5 and it will be seen that a mutually inductive coil in the anode of a pentode is used to modulate the effective value of the main tuning inductance without disturbing the oscillatory circuit seriously.

This means that the amplitude of oscillation can be very nearly constant over frequency ranges at least as great as $\pm 15\%$, and at central frequencies up to at least 10 Mc/s, since all the effects of circuit capacities can be tuned out and the valve input capacitance is unimportant. It is hoped to

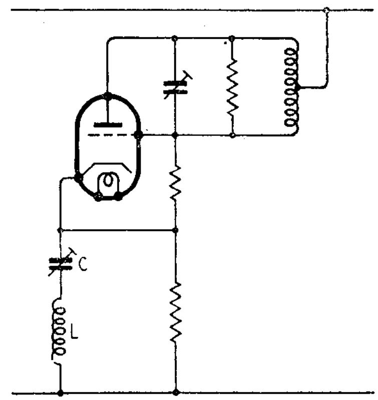
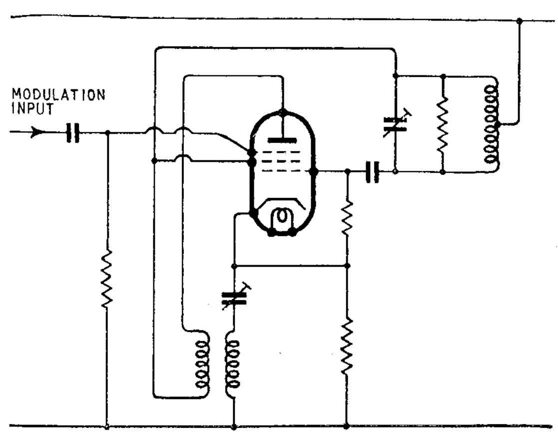



**Fig. 4. The second valve in Fig. 3 can be replaced by a phase-inverting transformer.**

deal with many of the more practical features of this circuit and in particular to give details of single-valve scanning oscillators for broadcast and television receiver alignment, in a second article to be published shortly; but it will already be clear that it has many advantages over the more usual reactance-valve arrangement, particularly as regards the frequency range covered and the higher central frequencies obtainable.

**Fig. 5. Completed circuit of a single-valve frequency-modulated oscillator capable of $\pm 15\%$ deviation.**



# RANGE OF V.H.F.

## *Part 2.—Ground Communication*

By M. V. CALLENDAR, M.A., (E. K. Cole, Ltd.)

THE actual range of communication between two stations on the ground is more difficult to estimate than the range between a ground station and an aircraft, owing to the greater number of obstacles usually intervening in the line of sight in the former case. However, even for ground-to-ground transmission much useful information can be gained from propagation theory, especially with regard to comparative ranges on different wavelengths, with different powers, or using aerials at different heights.

The minimum field required for intelligible reception at a fully efficient receiver is about 3 $\mu$V/m at 160 Mc/s and 2 $\mu$V/m at 80 Mc/s and 30 Mc/s, cosmic noise being the limiting factor at the latter frequency.

The distance at which the field falls to the minimum required has been calculated for a variety of conditions from the available formulæ (see Appendix) and the following tables give the range in miles for communication between two stations using vertical dipole aerials at the heights stated above ground. They are not intended to cover cases where a miniature battery portable, a super-regenerative set, or some other relatively inefficient receiver is used; reasonably small deficiencies in the receiver and losses in transmission lines, etc., can, however, be allowed for by choosing the appropriate column in the table.

**TABLE I**

**RANGE IN MILES OVER LAND**

| Height of Aerials | 30 Mc/s | | | | 80 Mc/s | | | | 160 Mc/s | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | A | B | C | D | A | B | C | D | A | B | C | D |
| 6ft/6ft ... | 2.0 | 3.5 | 6 | 10. | 1.8 | 3 | 5 | 8 | 1.8 | 3 | 5 | 8 |
| 12ft/12ft ... | 2.5 | 4.5 | 7.5 | 12 | 3.0 | 5.0 | 8.5 | 14 | 3.5 | 6 | 9 | 14 |
| 6ft/30ft ... | 4.5 | 7.5 | 12 | 19 | 4 | 6.5 | 10 | 16 | 4 | 6.5 | 10 | 15 |
| 6ft/100ft ... | 8 | 13 | 20 | 30 | 6 | 10 | 16 | 25 | 7 | 11 | 16 | 24 |
| 6ft/600ft ... | 15 | 22 | 32 | 45 | 13 | 20 | 30 | 44 | 13 | 20 | 30 | 42 |
| 30ft/30ft ... | 5.5 | 9 | 15 | 25 | 6.5 | 11 | 18 | 26 | 8 | 13 | 20 | 28 |
| 100ft/100ft or 30ft/330ft ... | 15 | 22 | 32 | 45 | 16 | 26 | 38 | 52 | 19 | 28 | 38 | 49 |
| 100ft/1000ft ... | 35 | 50 | 70 | 90 | 45 | 60 | 75 | 90 | 50 | 60 | 70 | 80 |

*Explanation of Tables.*

Use columns A, B, C, and D in Tables as follows:—

*Loss* is intended to cover the effects of:—

(a) Loss in signal due to obstructions, hills, houses, rigging on ships, etc.

(b) External interfering noise, other than cosmic noise.

(c) Any deficit in noise factor, or signal/noise ratio, of the actual receiver relative to a standard efficient receiver of noise factor = 10dB (or signal/noise ratio of 10dB for 30 per cent mod. at 1,000 c/s at 3$\mu$V from 70 ohms).

(d) Deficiencies in aerials and feeders relative to a standard dipole.

*Loss of* 10 *dB* corresponds to an average efficient equipment using a high slope r.f. valve, etc., in average country or open suburban areas, and represents the maximum likely loss over sea.

*Loss of* 20 *dB* should be allowed for in range estimates for cases

**TABLE II**

**RANGE IN MILES OVER SEA**

| Frequency | 30 Mc/s | | | 80 Mc/s | | | 160 Mc/s | | |
|---|---|---|---|---|---|---|---|---|---|
| Worst Height $H_w$ | 180 ft | | | 40 ft | | | 14ft | | |
| Aerial Heights | B | C | D | B | C | D | B | C | D |
| Aerials Low (<.5 $H_w$) ... ... ... | 40 | 55 | 75 | 16 | 25 | 37 | 7 | 12 | 20 |
| 6ft/30ft ... ... ... | 40 | 55 | 75 | 15 | 23 | 35 | 7 | 12 | 20 |
| 6ft/100ft ... ... ... | 40 | 55 | 75 | 17 | 27 | 39 | 15 | 23 | 32 |
| 30ft/30ft ... ... ... | 40 | 55 | 75 | 14 | 22 | 33 | 8 | 14 | 22 |
| 30ft/100ft ... ... ... | 40 | 55 | 75 | 16 | 25 | 37 | 16 | 25 | 34 |
| 100ft/100ft ... ... ... | 37 | 50 | 70 | 19 | 30 | 44 | 28 | 38 | 49 |
| 100ft/1000ft ... ... ... | 55 | 75 | 95 | 55 | 70 | 85 | 60 | 70 | 80 |

*Worst Height $H_w$ for transmission over sea :* when the height of either aerial above the sea is within about $\pm$ 50% of $H_w$, range is actually less than for very low aerials; the maximum loss in range is between 20% and 30% and occurs when *both* aerials are at a height = $H_w$. Range figures are greater than for land when one or both aerial heights are less than $H_w$.

where low aerials are used in fully built-up areas, though larger losses may be encountered in blind spots close to houses or behind hills, especially on the highest frequencies.

## Aerials

(a) Plain vertical dipoles are assumed for the tables above. If horizontal aerials are used, the range will become progressively less if the height of either aerial is reduced below 0.5λ over land or below $H_w$ over sea.

(b) If aerials must not exceed 5ft total length, range is reduced 10 to 40 per cent (depending upon earth) on 30 Mc/s only.

(c) If optimum directors are used at both dipoles, range is increased to that for ten times higher power. Simpler arrangements, e.g., reflectors at one or both aerials, will give less increase and 3 element arrays slightly more.

**TABLE III**

| Loss | Power Radiated | | | | |
|---|---|---|---|---|---|
| | 0.1w | 1.0w | 10w | 100w | 1Kw |
| 0dB | B | C | D | | |
| 10dB | A | B | C | D | |
| 20dB | | A | B | C | D |

(d) *Aerial Height* is that to centre of aerial: in the case of an aerial on a hill, in a valley, or on a tall building, it may be measured relative to the average height of the ground between the stations.

The Tables should give fairly reliable figures for average range obtained and for the variation of this average range with height of aerials and with transmitter power, etc. For example, we find that a given increase in power has a much greater effect than in the case of communication with aircraft; under most conditions we have here *an increase of ten times in transmitter power increasing range by about 70 per cent.*

However, when the range is great (much over 20 miles at 30 Mc/s or 10 miles at 160 Mc/s) the rate of increase in range with power becomes progressively less than that given by the above rule. Again, it is seen that doubling of the height of both aerials is equivalent to an increase of 16 times in power in most cases over land, but this rule does not hold for most practical heights over sea, nor for very low or very high aerials over land.

It is evident that horizontal aerials should not be used for transmission over sea, or for mobile operation with low aerials over land (except on 160 Mc/s), and that the lower frequencies should have preference when working over sea.

The two main factors which are not under control—viz., obstructions in the line of sight, and electrical interference—will often have a large effect in reducing the range obtained in any specific practical case. However, they should not greatly affect the *relative* ranges except in the following respects:—

(a) Electrical interference is mainly that due to car ignition systems, and this is most serious around 30 to 60 Mc/s. If such interference is serious, a "loss" at the receiver is, of course, no longer of importance within limits (e.g., 10 db loss at receiver is no longer equivalent to a reduction of ten times in transmitter power). This interference is less serious when horizontal aerials are used.

(b) Screening by houses, hills, etc., becomes progressively more serious as the frequency increases. This will tend to cancel, or even override, the effects of the reduced interference encountered at these frequencies.

The Tables may be taken to apply equally to amplitude or to frequency modulation systems, but the signal/noise ratio may be better with the latter type of modulation once the receiver is within the service area.

Range for the lightweight portable type set is normally less than that tabulated above for two reasons and apart from the low power available:—

(a) Noise factor is worse for for sets using battery valves.

(b) The aerial is restricted in length and suffers increased electrical losses if the set is carried or worn when in use.

The following table assumes a set using 1.4-volt valves, with a 2-watt d.c. input and 0.2-watt transmitter output. Figures for 4ft-aerial refer to the case where the set is worn when in use, while those for the λ/2 aerial assume the station to be temporarily on the ground. Theoretical range (0 db loss) is about 1.5 times that shown, and range under poor conditions is about half that shown. Ranges given for 160 Mc/s might not be attainable above about 140 Mc/s owing to difficulties with valves.

**TABLE IV**

| Average Range for Portable | | 30 Mc/s | | 80 Mc/s | | 160 Mc/s | |
|---|---|---|---|---|---|---|---|
| Aerial Length | Aerial Height | | | | | | |
| | | Land | Sea | Land | Sea | Land | Sea |
| 4ft | 6ft (worn) | 1.5 | 18 | 1.4 | 9 | 0.9 | 3 |
| λ/2 | 6ft | 3.0 | 30 | 1.7 | 10 | 1.0 | 3 |
| λ/2 | 30ft | 8 | 30 | 7 | 8 | 4 | 3 |

## APPENDIX

As in Part 1 on air to ground communication, the figures here are based upon a paper on "Range of Low Power Radio Communication" by the present writer in *Jour I.E.E.* for November, 1948. This paper should be consulted if more complete formulæ and references are desired.

The horizon distance bears no simple relation to range when the aerials are at low heights. The following well-known simple formula is applicable to a good proportion of cases: $E = \frac{88\ h_1\ h_2}{\lambda d^2} \sqrt{P}$ where E is field in V/m. P is power radiated in watts and $h_1$ $h_2$ are aerial heights above ground; λ is wavelength and d is distance, all in metres. To cover the case of low aerials with vertical polarization, we must make the following simple correction:

Substitute 0.5λ (or $2.5\lambda^{3/2}$ if over sea) for $h_1$ or $h_2$ in all cases where $h_1$ or $h_2$ is less than the critical height of 0.5λ (or $2.5\lambda^{3/2}$ over sea).

For horizontal polarization the uncorrected formula holds down to the lowest practical heights.

However, at long distances

**Range of V.H.F.—**
additional attenuation occurs due to the earth's curvature, this extra attenuation reaching 3db at a distance of 13 $\lambda^{1/3}$ miles, and 8-10db at double this distance. This correction, and also smaller corrections required for very high aerials (over about 300ft) and for aerials near the critical height (or the "worst height" 1.8 $\lambda^{3/2}$ over sea) have been allowed for in the tables. Average constants (K = 10) are assumed for land.

The simple law given in the text for variation of range with power corresponds, of course, to the simple square law formula.

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# ELECTRONIC CIRCUITRY

## *Selections from a Designer's Notebook*

**By J. McG. SOWERBY** (*Cinema Television Ltd.*)

### Voltage Stabilizer for C.R. Tubes

As readers are well aware, the sensitivity of an electrostatic cathode-ray tube is inversely proportional to the final accelerating potential. Consequently, if a cathode-ray tube is to be more than a rather approximate device, and if any serious measurements are to be taken with it, this point must be allowed for. When recording photographically (in particular) some form of stabilized supply potential for the cathode ray tube should be provided. The type of circuit to be used will depend very much on the supply potential required, and at low voltage (up to 750 volts, say) the conventional series[1] or shunt[2] stabilizer is quite satisfactory. At higher potentials (750V to 5kV) the use of these conventional circuits is hampered by the lack of suitable valves among those currently available.

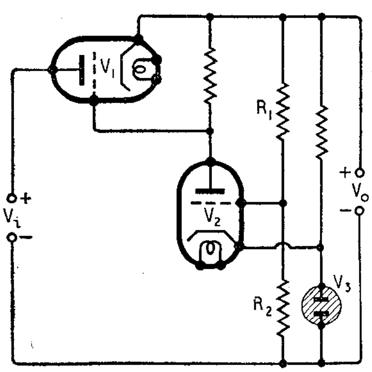


Fig. 1. Conventional series stabilizer.

The difficulty with conventional circuits is well brought out by consideration of one of them—as shown in Fig. 1. This represents a simplified series degenerative stabilizer. The reference potential is supplied by the drop across $V_3$ (100volts, say), and this is compared with the fraction of the output $V_0$ (2500 volts, say), obtained across the potential divider $R_1$, $R_2$. Any difference between these potentials is amplified by $V_2$ and applied to $V_1$ in such a way as to tend to keep $V_0$ constant. As the cathode-grid potential of $V_1$ will be small (10 volts, say), the anode-cathode potential of $V_2$ will approach $V_0$—taking the given figures it will be about 2400 volts. Now this is greatly in excess of the rated anode-cathode voltage of almost—if not quite—all small amplifier triodes and pentodes. Consequently, the circuit is rather unpractical and an alternative must be found.

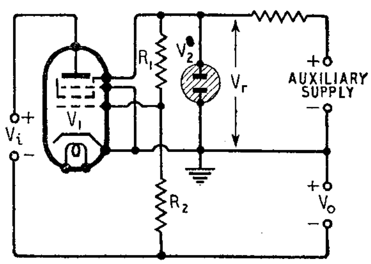


Fig. 2. Simple E.H.T. stabilizer.

A circuit which might be used to achieve the required result is shown in Fig. 2. Here $V_1$ is a so-called "time-base" pentode or tetrode rated at 10/30 watts dissipation with a top-cap anode rated to withstand several kilovolts peak. The EL38 is a typical example of this class of valve. The reference potential $V_r$ is supplied by the drop across $V_2$ which is supplied through a ballast resistor from an auxiliary source—such as a time base or amplifier power pack. When $V_r$ is large compared with the grid base of $V_1$, the stabilisation ratio, $S_0$, is given by:

$$S_0 = 1 + r_{a1}\frac{I_{a1}}{V_0} + \frac{\mu_1}{V_0/V_r + 1}$$
(approx.)

Where $r_{a1}$ = anode resistance of $V_1$, $I_{a1}$ = anode current of $V_1$ and $\mu_1$ = amplification factor of $V_1$.

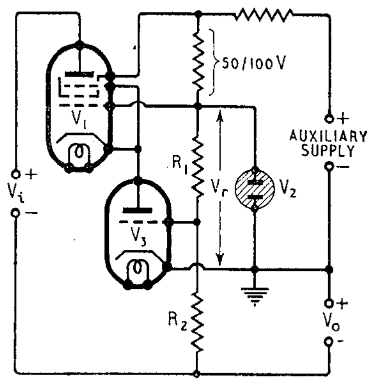


Fig. 3. Improved E.H.T. stabilizer.

If we now put in reasonable values for the parameters in this relation — assuming an EL38, ($I_{a1}$ = 2 mA, $r_{a1}$ = 100 k$\Omega$, $V_0$ = 2,500 volts, $V_r$ = 100 volts, and $\mu_1$ = 100), we find $S_0$ = 5. This means that mains fluctuations will be reduced by a factor of 5 before reaching the cathode-ray tube. Although this is a step in the right direction, the advantages of the stabilizer are not very marked.

An improved circuit[3] used by the writer with some success is shown in Fig. 3, and it will be seen that another valve has been added in series with $V_1$ of Fig. 2.

When this circuit is analysed we find that the stabilization ratio is:

$$S_0 = 1 + \frac{\mu_1 r_{a3} I_{a3}}{V_0} + \frac{\mu_1\mu_3}{V_0 V/_r + 1}$$
(approx.)

If we assume that $\mu_3$ = 50 and $r_{a3}$ = 30 k$\Omega$, and take the other parameters as before ($V_r$ = 100, $V_0$ = 2500 volts, etc.), we find $S_0$ = 2000 approximately. This, of course, is a very useful figure.

It will be noticed that $V_3$ may

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[1] Scroggie, M. G. *Wireless World*, October, November, December, 1948.

[2] Sowerby, J. McG, *Wireless World*, June, 1948.

[3] Patent applied for.

be almost any sort of triode provided it has a reasonably high amplification factor, and is capable of passing the load current for the cathode-ray tube and bleeder network (about 2 or 3 mA altogether). The maximum anode-cathode potential of $V_3$ is roughly $V_r$ plus the working grid bias of $V_1$, and need never exceed 150 volts at the maximum. Thus, the potential by which $V_i$ exceeds $V_0$ appears almost entirely across $V_1$, and this may be a valve of the EL38 class capable of withstanding about 3 kV.

The resistance looking back into the stabilizer terminals is approximately $R_0 = \frac{1}{g_{m3}} \cdot \frac{V_0}{V_r}$, but this is not usually of much interest; if it is, in some particular application, then to maintain a low $R_0$, $V_3$ should be a valve having a large mutual conductance.

A further point should be noted; provided $V_i$ is switched on only when $V_1$ and $V_3$ are "hot," $V_1$ need only withstand the fluctuations of the supply potential plus a margin for safety. Thus, with ordinary receiving valves stabilized output voltages up to about 10 kV may be obtained by the use of this circuit.

For the best results the heater of $V_3$ should be supplied via a "constant-voltage" transformer, and the network $R_1$, $R_2$—and any part of it which may be variable for adjusting $V_0$—should be screened.

## A Pulse Stretcher

IT sometimes happens in electronic control applications that one is required to close a relay momentarily (e.g. for 1/10 to ½ sec.) each time a relatively short pulse occurs somewhere in the circuit. A typical example is the need to operate an electro-mechanical counter whenever a light beam falling on a photocell is interrupted. The natural choice of circuit for this sort of service would be a time-delay trigger circuit of one kind or another as previously discussed in these columns. It is not always remembered that an otherwise undesirable quality of a cathode follower can often be utilized for the purpose, and this sometimes leads to a considerable simplification with resultant economy.

The circuit of the cathode-follower pulse stretcher is shown in Fig 4. A simple analysis of the circuit shows that, provided the amplitude of the input pulse is considerably greater than the grid base of the valve, $C_k$ charges (input increasing positively), with a time constant $C_k/g_m$ approximately, where $g_m$ is the mutual conductance of the valve. When the input pulse collapses from its peak to zero, $V_1$ is cut off and $C_k$ discharges through $R_k$, until $V_1$ again conducts, with a time constant of $C_kR_k$—exactly as would be expected. Consequently, for an input waveform as shown in Fig 5(a), an output waveform as at (b) will be obtained provided $R_k \gg \frac{1}{g_m}$.

It is convenient to note that the ratio of the rising and falling time

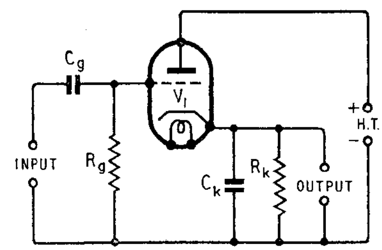



Fig. 4. Cathode-follower pulse stretcher.

constants is simply $g_mR_k$, so that for large inputs one may loosely say that the input pulse duration can be multiplied by the factor $g_mR_k$. It will sometimes happen that a relatively long time constant $R_kC_k$ will be needed, and it is worth remembering that it is generally preferable to increase $R_k$ rather than $C_k$, to obtain the desired result. The rate of rise of cathode potential is limited by $C_k/g_m$, and if the rate of rise of grid potential is in excess of that possible at the cathode, the valve will run into grid current in an attempt to charge up $C_k$ rapidly. A valve with high mutual conductance will help to reduce this trouble, and will also operate correctly with a lower input amplitude than is the case with a low $g_m$ valve.

If we use an EF50 for $V_1$, make $R_k = 100\ \text{k}\Omega$, and choose $C_k$ with due regard for the input waveform, an input pulse duration may certainly be stretched 100 times, or more, and the output waveform used to operate a valve controlling a relay or any other device. If pulse stretching by a greater

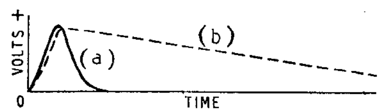



Fig. 5. Response (b) of the stretcher to a typical pulse input (a).

factor than this is needed it will usually be desirable to revert to the more complicated but more flexible time-delay trigger circuit.

## MANUFACTURERS' LITERATURE

Leaflet describing the "Bafflette Bonnie" extension loudspeaker, from Richard Allen, Caledonia Road, Batley, Yorkshire.

The 1949 catalogue of components and accessories made by Belling and Lee, Cambridge Arterial Road, Enfield, Middlesex, has now been issued and is available to manufacturers, government and professional organizations. Copies will be sent automatically to those who already have previous editions.

Data sheet of valves and c.r. tubes, from Ferranti (Electronics Dept.), Moston, Manchester, 10.

Technical publication No. 21 ("Sorbsil" Silica Gel for water and organic vapour adsorption), from Joseph Crosfield and Sons (Chemical Dept.), Warrington, Lancashire.

Data sheets for the following transmitting valves: DET18, ACT19, BR124, BR125 and TT12, from Marconi's Wireless Telegraph Co., Ltd., Chelmsford.

Illustrated leaflet describing the new Type "A" potentiometer, from Morganite Resistors, Paulsway, Bede Trading Estate, Jarrow, Co. Durham.

Leaflet giving technical details of the Redifon Type G41 short-wave transmitter (5-7½ kW), from Rediffusion, Ltd., Broomhill Road, Wandsworth, London, S.W.18.

Information sheets showing the application of electronic control methods in industry, from Sargrove Electronics, Sir Richard's Bridge, Walton-on-Thames. Examples are given of the many uses to which the Sargrove rectifier-photocell and "Phasitron" circuits can be put for counting, inspection and machine protection.

Folder giving dimensions and electrical data of "Gecalloy" radio dust cores, from Salford Electrical Instruments, Silk Street, Salford, 3, Lancashire.

# NOTES ON THE *Wireless World* TELEVISION RECEIVER

THE performance of the line time base is greatly affected by the core material of the line-scan transformer. Core losses are relied upon very largely for damping and the linearity control, although it does provide additional damping, does not give sufficient for some grades of iron.

During flyback, the equivalent circuit has the form shown in Fig. 1 (a), where L is the total effective inductance, C the shunt capacitance, R the resistance simulating the core losses and $R_1$ and $C_1$ the components of the linearity control. The circuit must have a natural frequency of about 34kc/s, so that one-half cycle is about equal to the fly-back time, and be so damped that the overshoot on the half cycle is only a few per cent.

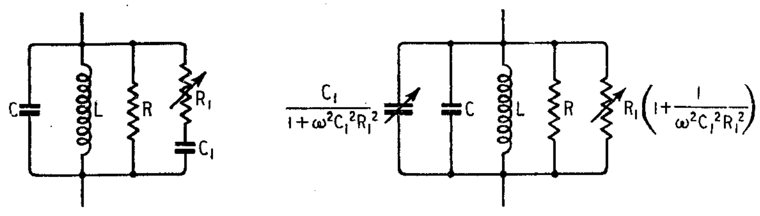



**Fig. 1. Equivalent circuit of output stage (a) and an alternative equivalent (b) which is approximately valid during the fly-back.**

At any single frequency $R_1$ and $C_1$ can be replaced by a capacitance and resistance as shown in Fig. 1 (b). A variation of $R_1$ thus varies simultaneously both the capacitance and resistance of the equivalent circuit. Under normal conditions the resistance variation predominates. However, if the transformer core losses are abnormal, the range of control given by the circuit is inadequate, for it is then necessary to change $C_1$ as well as $R_1$. This does not form a satisfactory arrangement, however, for the optimum value of $C_1$ is quite critical under some damping conditions.

It has been found better to modify the circuit to the form shown in Fig. 2, so that the damping resistance is directly in shunt with the transformer secondary. The capacitance effect of $C_s$ of the original diagram is now provided by a shunt capacitance on the secondary. This is represented by $C_A$ and $C_B$ in parallel.

A variable capacitor of 0.0015$\mu$F maximum should meet all requirements and $C_B$ would then be unnecessary. Such a capacitance is readily obtained from a 3-gang capacitor with its sections in parallel, but while this is convenient experimentally, it is rather clumsy as a permanent feature of the set. It is practically convenient, therefore, to make $C_A$ of 500pF only and to add an appropriate fixed capacitance $C_B$ in shunt. This must be found by trial, but will usually be 500pF. The capacitor should be rated for 750V peak. An ordinary bakelite-dielectric reaction capacitor has been used successfully for $C_A$ and is both cheap and compact. While the voltage is considerably above that normally used on such a component, one has been working satisfactorily for some time. No damage to other components is likely to result from a breakdown. The capacitor must be mounted with its shaft earthy.

As an alternative to $C_A$ and $C_B$, a 100-pF variable capacitor can be connected across the transformer primary. As this component must withstand up to 3-kV peak, it is usually easier to use the larger capacitance on the secondary.

The adjustment of $C_A$ is critical. Adjust the width for about a 6-in picture and examine it carefully. If the left-hand side is expanded and there is a whitish vertical line or band on the right of this expanded portion, reduce the resistance by $R_{19}$—$R_{23}$. If the left-hand side is linear but folded over, increase the resistance. Adjust the resistance until there is a small amount of expansion on the left.

Next increase picture width. The expansion will be reduced and may disappear and be replaced by a foldover. If so, increase the resistance. A vertical white line an inch or so from the left may now appear. Adjust $C_A$ to minimize it. Continue the process until the proper width is obtained. Check the fly-back time on the test pattern. If it is too great, $C_A$ must be reduced and this will necessitate reducing the width slightly and readjusting the resistance.

It is normally possible slightly to overscan the tube so that there should be no difficulty in obtaining the full width. When the values are approximately correct, the resistance adjustment is not very critical, but $C_A$ needs precise setting.

The circuit has been used successfully with transformers having cores of Silcor II, 0.018-in, laminations, and Silcor III, 0.02-in. With core materials of greater loss, the original circuit is preferred.

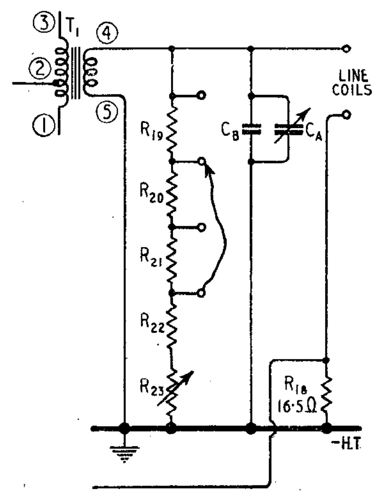



**Fig. 2. Modified damping circuit recommended for transformer cores having low losses. The references are as on the original diagram except for the new parts $C_A$, $C_B$.**

# ELECTROMEDICAL STIMULATORS

## Application of Radar Circuit Techniques for Diagnosis and Treatment

By O. B. SNEATH, B.Sc., and E. G. MAYER, D.Sc.
(Multitone Electric Company)

IN physiotherapeutic and orthopaedic practice it is desirable to have available certain electric currents for the purpose of stimulating muscle and nerve tissue. The reactions enable assessment to be made of faults in these tissues, and the stimulation can also be used to treat muscle and nerve fibres.

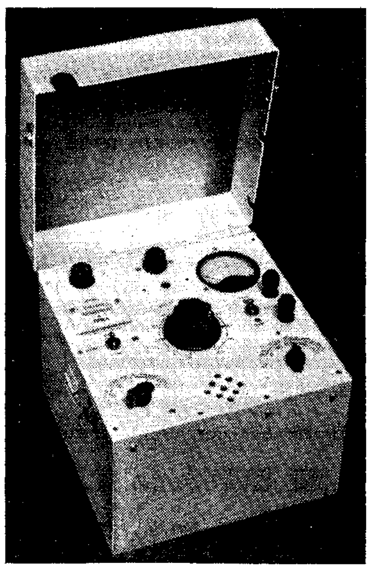
By passing a d.c. current through muscle tissue, a contraction is obtained and in the past this was supplied by a source of d.c. passed through a mechanical interrupter to the electrode in contact with the body surface. The interrupter usually took the form of a " Metronome " having a curved crossbar at the top of the inverted pendulum, one end dipping into a bottle containing mercury at each stroke and thus gave the necessary interruptions. The speed of the " Metronome " controlled the length of shock and repetition rate, the shock length varying approximately from 0.3 to 1.0 second.

Nerve tissue responds to comparatively short shocks which have no effect on muscle fibre, and for nerve stimulation damped waves were used. These were produced by means of a " Faradic Coil " which consisted of a vibrator interrupting a direct current through the primary of a double-wound transformer, the current in the secondary being applied to the tissue where stimulation was required.

For diagnosis, muscle reactions to d.c. shocks were noted and also the effect of damped waves applied to motor nerves controlling the muscle. Large variations were encountered due to varying " Metronome " and vibrator speeds, neither of which was generally very accurately measured; at best the results of tests were described by different people in varying ways.

The current generators described above were more satisfactory in the treatment of muscles and nerves by stimulating exercise, but were extremely uncomfortable to the patient and the constant increase and decay of current to contract and relax a muscle at a natural speed in many cases caused such pain that sufficient output could not be tolerated.

Experiments were carried out by Professor A. E. Ritchie, now at St. Andrews University, who found that long pulses of approximately 0.1 millisecond duration produced with electronic square-wave generators, stimulated muscle tissue, whilst shorter pulses affected motor nerves only, the reaction falling away rapidly at 0.01 milliseconds. Professor Ritchie collaborating with O. B. Sneath, then perfected a muscle stimulator producing pulses of 100, 10, 1, 0.1 and 0.01 milliseconds which could be applied to muscle tissue in such a manner that the voltage necessary at each pulse length to produce a given stimulation could be accurately read. A graph could then be drawn showing voltage against pulse length. This makes accurate diagnosis possible by showing the point when response falls, and the degree of nerve and muscle degeneration.



Ritchie-Sneath stimulator unit for diagnostic work.

It was found that the square voltage pulse output of the electronic apparatus was comfortable compared with older forms of mechanical generators and interrupters, and so this technique was applied to treatment units. These units first took the form of multi-current generators producing long, medium and short duration pulses, sinusoidal and uninterrupted direct currents, all of which could be applied in various modified forms with fast and slow repetition rates, and surged from zero to a pre-set maximum strength. A further instrument has been produced to replace the old " Faradic Coil." The output is 0.3 millisecond pulses repeated 50 times per second, and this output can be surged at five different rates with adequate spacing between surges for muscle relaxation.

Pulse generators have been extensively developed in connection with radar and telecommunication, and whilst electromedical apparatus is not generally required to give as great a stability of pulse length and interval as radar, it is necessary to be able to obtain, by switching, pulses of widely varying lengths and repetition rates. The pulse lengths commonly employed are 1 second, 100, 10, 1, 0.1 and 0.01 milliseconds, though the longest and shortest of these are often dispensed with. The repetition rate required may be 50 per second to produce continuous nerve excitation and also one or several per second as does the " Faradic Coil."

**Electromedical Stimulators**

In apparatus for treatment, as distinct from diagnosis, the rapidly repeated pulses require to be surged or varied from zero to maximum at a rate from 6 to 100 times per minute depending on the treatment. Spacing between surges should be sufficient to allow muscle tissue to relax and though this may be achieved manually by moving the strength control, it is far more convenient and accurate for this to be done automatically.

It will be realized that the long pulse lengths and slow repetition rates involved raise problems of coupling and decoupling somewhat different from those usually met with in radio practice.

In addition to square pulses, sinusoidal and continuous currents may be required, and provision must be made for surging these; a further requirement is to apply alternate surges, either with reversed polarity or to different pairs of electrodes in order to stimulate opposing muscles in turn.

One of the chief problems in the design of the apparatus is to produce this wide variety of currents without undue complexity in the switching, and to avoid the use of an excessive number of valves. In general, the maximum output voltage required is approximately 100 volts at 100 mA, although for short pulses of 1 millisecond or less nearly double these values may be necessary.

The only standard valves suitable for such relatively high current-to-voltage ratios, are output valves designed for use in a.c./d.c. instruments as for example 35L6's or Mullard CL33's in parallel. By connecting the screens to h.t. + through suitable resistors the drop across the anode load, when the grid is not biased, may be made only a fraction less than the h.t. supply voltage. Applying pulses to the grids, which are normally biased to cut-off, causes the valves in the output stage to resemble a high-speed relay in their action.

The best-known circuit for producing square-topped pulses is probably the multivibrator employing two triodes. The output may be obtained from a potentiometer in the anode of one of these triodes, but variations in output load, and most methods of surging the power would, in normal circuits, vary the pulse length.

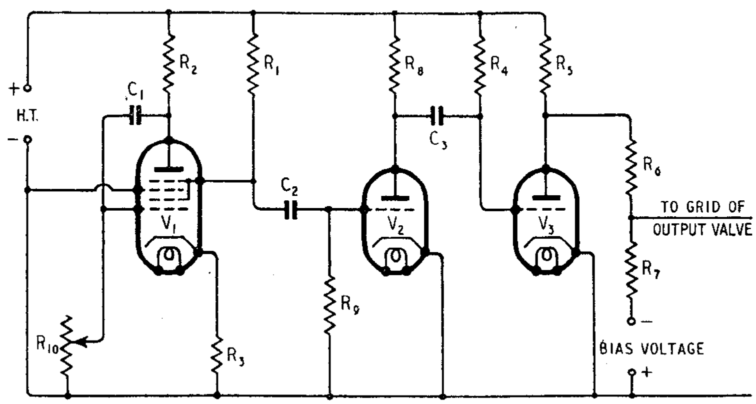



Fig. 2. "Miller integrator" type of circuit using a hexode as oscillator.

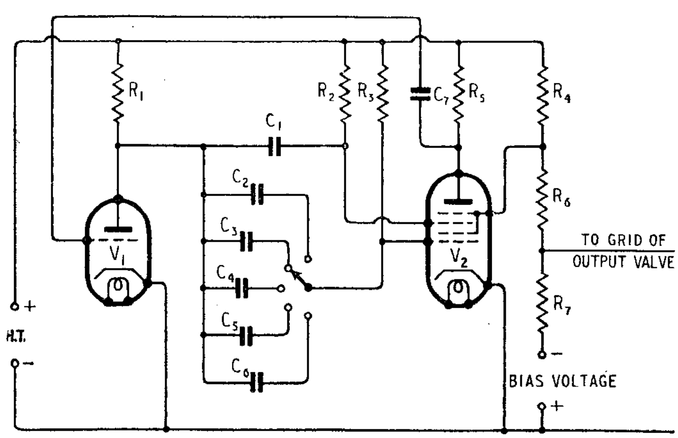
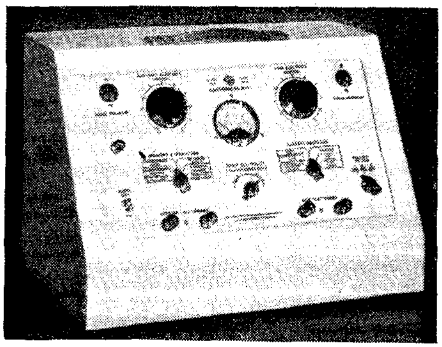



Fig. 1. Modified multivibrator circuit giving variable pulse length and spacing.

To overcome this difficulty the anode of one of the valves in the multivibrator may be coupled to the grid of the output valves which serve merely as an "on-off" switch for the current.

A factor of importance in this application of the multivibrator is the limitation of the maximum ratio obtainable between the periods of the two phases. This depends on the fact that the valve which is "blocked" for the long period has to have its grid condenser recharged during the short period, and is of the order of the ratio of the maximum value of grid leak employed to the anode load impedance. It is not generally satisfactory to use a grid leak of more than 10 megohms; thus if the anode impedance is 10,000 ohms the interval-to-pulse ratio will be limited to approximately 1,000/1 although by using a power valve biased to cut-off on the shorter pulses, a higher ratio can be obtained. There is no need for the valves to be similar, and a high-impedance valve with a higher resistor in the anode circuit could be used in conjunction with the power valve.

A modified form of "multivibrator" circuit devised by O. B. Sneath enabling 10 or 100 microseconds, or 10 or 100 millisecond pulses to be repeated one a second, employs a triode and hexode as shown in Fig. 1. It will be seen that the triode $V_1$ is coupled to both the inner and outer control grids of the hexode $V_2$, the condensers $C_2$ to $C_6$, coupling it to the inner grid, vary the pulse lengths. When the time factor of the coupling to the inner grid is greater than that of the outer, the circuit behaves as a normal multivibrator, the outer grid only being biased when $V_2$ is not passing current owing to the outer grid being biased to cut-off; this condition applies for pulses of one millisecond or longer, using condensers say $C_2$ to $C_4$.



Multi-treatment unit providing all types of current required in physiotherapy.

For short pulses the anode of $V_2$ remains positive for a period depending on the coupling to the outer control grid, which is adequate to charge the condenser $C_2$, but the screen from which the output is taken remains positive for a shorter time depending on the coupling condensers $C_5$ or $C_6$. Under this condition the output from the screen is not a true multivibrator output, as the unbiasing of the inner screen is gradual, but allowing for steepening effect due to the gain of the output valve it can be made satisfactory.

Another method of producing square waves is to apply the output from a multivibrator, both constants of which are longer than the required pulse length, through a capacitor and resistor giving the required period to the grid of an amplifier which serves as or contains a limiter.

Instead of the usual multivibrator circuit, a form of circuit which has been previously described in *Wireless World* under the titles " Miller " or " Blumlein " oscillators may be employed, and is the start of the circuit for producing square pulses shown in Fig. 2. It was, however, found advantageous to employ a hexode rather than a pentode valve, as this enables the anode to be biased off by the outer grid with a lower value of cathode resistor; the outer grid takes the place of the suppressor, and the two coupled screens that of the single screen of the pentode. The behaviour of the circuit is as follows.

Assume the inner grid of $V_1$ to be biased, the bias will gradually decrease as $C_1$ discharges through $R_{10}$ and both screen and anode voltage will fall. The valve $V_1$ causes the time constant of $C_1$ and $R_{10}$ to be multiplied by a factor equal to one plus the effective amplification factor of the valve in conjunction with its anode resistor $R_2$, and hence the circuit is very convenient where long time intervals are required As the current increases, a point will be reached where this increase continues but the anode current falls and anode potential starts to rise, due to the cathode resistor $R_3$ increasing bias on the outer grid; the rise in anode potential is communicated through condenser $C_1$ to the inner grid, and the screen current suddenly rises until the anode current is cut off by the bias on the outer grid, and the inner grid is driven positive.

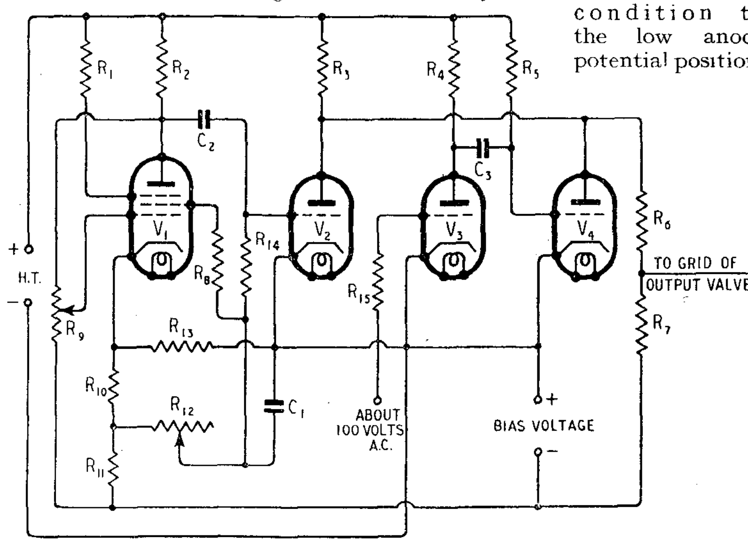



Fig. 3. Circuit for periodic amplitude control or surging of pulses at mains frequency.

This causes grid current to flow as in the normal multivibrator until the anode and inner grid drops back and stability is reached, with the inner grid not completely biased to cut-off.

The resistor $R_{10}$ has one end taken to h.t. − rather than to h.t. +, as this reduces the values of $C_1$ and $R_{10}$ for a given pulse repeat rate, it also reduces frequency instability with varying h.t. supply. The output of $V_1$ is taken from the screen and not the anode, as this circuit has a lower impedance. It is coupled to $V_2$ through a resistance-capacitance coupling of long time constant compared with the pulse length, and this triode $V_2$ has its anode coupled to the grid of $V_3$. The coupling condenser $C_3$ and resistor $R_4$ between the grid of $V_3$ and h.t. + controls the length of the pulse.

The pulse is produced once in each complete cycle of $V_1$, when the screens of $V_1$ suddenly become positive, causing the anode potential of $V_2$ to drop, and $V_3$ to be biased to cut-off. The steepness of the termination of the pulses is increased by raising the gain of $V_3$, so that a smaller change in grid voltage is required to carry it out of the cut-off condition to the low anode potential position.

The most obvious way to surge a series of rapidly repeated pulses at a slow rate would appear to be to vary the h.t. supply to the output valves. This, however,

**Electromedical Stimulators**

involves a duplication of the power valves and also an increase in the initial h.t. voltage and power dissipation, as there is bound to be a voltage drop in the surging valve even when the surge is at maximum. An alternative method is to couple the grid of the output valve to the anode of two valves connected together, one of these valves shown as $V_4$ in Fig. 3 ceasing to pass current during each pulse, and the other, $V_2$, being arranged to pass minimum current at the peak of the surge. The output valve only passes current when $V_4$ is "blocked" out, the amount of such current depending on the state of bias on $V_2$. Fig. 3 shows a circuit for producing surged pulses at a repetition rate of mains frequency. The circuit of the surge-producing valve $V_1$ is capable of producing surges with both a gradual rise and a gradual decay, but the rise and decay may be different, and the ratio is controlled by the value of the resistors $R_{10}$, $R_{11}$ and to a lesser extent by the potentiometer $R_9$.

The cycle of operation is as follows. Assuming the suppressor grid of the pentode $V_1$ is at zero potential or has only a small negative bias, the anode will then pass current, and if $R_2$ is sufficiently high, will be at fairly low potential. Owing to the coupling through the potentiometer $R_9$, the inner grid will, however, be somewhat biased, and the total cathode current fairly low. This will cause the cathode potential to be low, and the negative bias on the suppressor will gradually increase, due to the current flowing through $R_{12}$ charging $C_1$. Finally a point is reached where this bias causes the anode voltage to rise, which causes a rise of the inner grid voltage, and, provided that the screen resistor $R_1$ is low, a sudden rise in total current through $V_1$ and also of cathode voltage. The suppressor grid will now gradually lose bias until the anode becomes relatively negative again and the cycle is completed. Suitable values of resistors to obtain this result are: $-R_2$, 220k$\Omega$; $R_1$, 22k$\Omega$; $R_{13}$, 15k$\Omega$ and $R_9$, a 1M$\Omega$ potentiometer. A source of negative bias is required of about the same voltage as the high tension. The voltage across $R_1$ is fairly steady at a low value for a part of the cycle and at a higher voltage for the remainder, but at the junction of $R_{12}$ and $C_1$ there is a fairly steady rise followed by a fairly steady fall of negative bias. The total length of the cycle is controlled by $R_{12}$. The junction of $R_{12}$ and $C_1$ is coupled through a high resistance to $V_2$ which surges the pulses, these can be produced by applying a.c. at about 100 volts, 50 c/s to the grid of $V_3$ through a resistance. Approximately square pulses are produced on the anode of $V_3$, which is coupled through a condenser $C_3$ to the grid of $V_4$. This condenser and a resistor $R_5$ to h.t. + determine the pulse length.

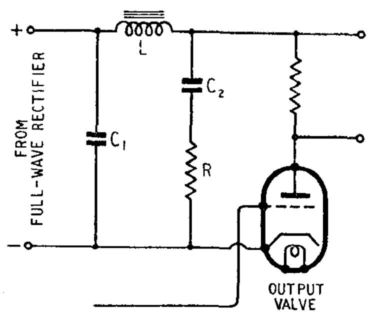



**Fig. 4. Stabilizing circuit for h.t. supply when using long pulses.**

The form of surge usually desired is a gradual rise followed by a rapid decay The surge should be effective for about a quarter of the time, which in general means that it should be over half value for about a quarter of the time. It is of little consequence how much of the remaining time the pulses are absent or at a low level. It is probably almost as effective to have a gradual rise followed by an instantaneous decay, and a number of well-known circuits similar to those used for television scanning can be adapted for this purpose. Many of these circuits can only be varied over the required range, while retaining their wave form, by employing ganged controls.

It is desirable to provide a source of constant amplitude sinusoidal voltage and also to surge this voltage at varying repeat rates. Whilst "continuous sinusoidal" is, of course, obtainable from the secondary of the mains transformer, it is very involved to surge this without entirely changing the waveform. Probably the simplest method is to produce surged square pulses of approximately 10 milliseconds in length at a repeat rate of 50 per second and to connect in parallel with the output a 50-c/s resonant circuit.

A special problem arises in connection with the longer pulses, and is due to the gradual fall in voltage after the start of the pulse owing to the discharging of the reservoir condensers in the power pack. This could be solved by the use of a stabilizing device normally drawing current comparable with that of the pulse, but this would greatly increase the power dissipation of the apparatus, and is therefore undesirable. The arrangement shown in Fig. 4 has been employed,* and the general effect is that when the valve starts to draw current, this is obtained from the condenser $C_2$ charged to open circuit power pack voltage, but the voltage is dropped by the resistor R. As the current continues, the voltage of $C_2$ falls, but the effect is compensated by the rise of current through the choke L. Owing to the non-linear nature of the rectifier resistance and the complication due to the first reservoir condenser, no simple formula can be given for the value of R and perfect squaring is not achieved, but reasonably good results were obtained with R, 600 ohms, L, 12 henrys, $C_1$, 16$\mu$F and $C_2$ 32$\mu$F.

* Patent has been applied for in connection with this arrangement.

REFERENCES

Walter, W. Grey and Ritchie, A. E., "Electronic Stimulators," *Electronic Engineering* 1945, Vol. 17, p. 585.

Ritchie, A. E. "Thermionic Valve Stimulators, *British Journal of Physical Medicine* 1948, Vol. 11, p. 101.

## PORTABLE RECORD PLAYER

DESIGNED for operation from a.c. mains, the "Karrigam" record player made by the V.S.E. Construction Co., 5-7, Denman Street, London, W.1, employs the new all-class B8A miniature valves and the heaters are fed in series through a capacitor. Negative feedback is employed, and the output is 2½ watts to a flat-type 6½in moving-coil loudspeaker mounted in the lid of the rexine-covered carrying case. A lightweight Garrard turntable unit is used and bass-boost at 3.8db per octave is provided in the 3-valve circuit which works in conjunction with a high-quality magnetic pick-up. The weight of the player is 16¾ lb and the price is £26 19s 8d.