VALVE VOLTMETER

without calibration drift

"Infinite-input, Zero-output-resistance" Adaptor for use with any D.C. Voltmeter

By M. G. SCROGGIE, B.Sc., M.I.E.E.

ALTHOUGH much has been written about valve voltmeters (including at least one large book), so far as the writer is aware the subject has never been pursued to its logical conclusion. If we are to attempt to do so, we must consider what a valve voltmeter essentially is—a voltmeter in which the connection of the indicating instrument to the voltage to be measured is indirect, through an impedance-converting device, in order to eliminate as far as possible any loading effect on the source of the voltage.

It will be noticed that no mention has been made of a rectifier. That is because it is required only when measuring alternating voltages, and although most valve voltmeters include provision for doing this it is not a *sine qua non*. The essential element is the impedance converter, and it is the fact that valves can so readily be arranged to have a very high input impedance and relatively low output impedance that makes them so useful for that purpose. Some of the simpler types of valve voltmeter combine rectifying and impedance-converting functions in one valve, but the tendency nowadays is to separate them. The design of the rectifying portion has already received much publicity; in the following article it is intended to confine attention to the impedance converter, which is required for both alternating and direct voltages.

It will be assumed that the indicating instrument is a moving-coil meter, as experience has shown this to be the most generally convenient. Preferably it should not have to be highly sensitive, because microammeters are both expensive and easily damaged.

Clearly, then, the ideal converter would be one with infinite input impedance (for it would then have no effect at all on the voltage being measured), and zero output resistance (for then the whole of the meter circuit resistance could be wire-wound and highly constant). Although one of the main objects in valve voltmeter design has long been to make the output resistance as stable as possible, the results are usually quite a long way from the ideal zero. And it seems to be too often assumed that any valve worked with negative bias has negligible d.c. input conductance. The purpose of this article is, first, to show that where valve voltmeters are concerned this assumption is false, and, secondly, to explain how the output resistance can be made practically zero, so that the valve gear can be added to any ordinary moving-coil voltmeter, to adapt it for "infinite-impedance" measurements, without affecting its calibration. It follows that the calibration is unaffected by even large changes in the valve characteristics, because a variable factor multiplied by 0 remains constant at 0.

First, then, the matter of input conductance. The elementary book or instructor explains that current through a hard valve flows to the cathode only from a positive electrode; so as long as the grid is negative to cathode there is no grid current. For ordinary amplifier and receiver circuits, in which the resistance in series with the grid seldom exceeds 1MΩ, and nobody is any the worse for a slight shift of grid bias, that is a fair enough simplification. The trouble comes if one carries this early impression into work on measuring apparatus, where grid circuits may be much above 1MΩ and an unintended shift of even a small fraction of a volt is intolerable.

In the work which led to this article it was considered that a valve voltmeter in which the reading was noticeably affected by a grid circuit resistance of 20MΩ would be a poor thing. The fact that published valve voltmeter circuits showed rectifier load resistances of anything up to 100MΩ, followed by (as far as could

Fig. 1. Conventional valve voltmeter circuit.

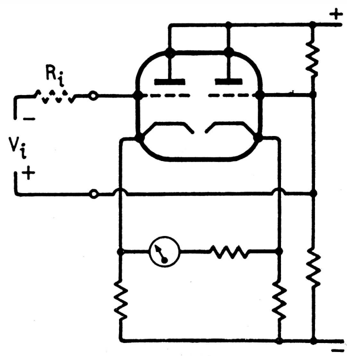
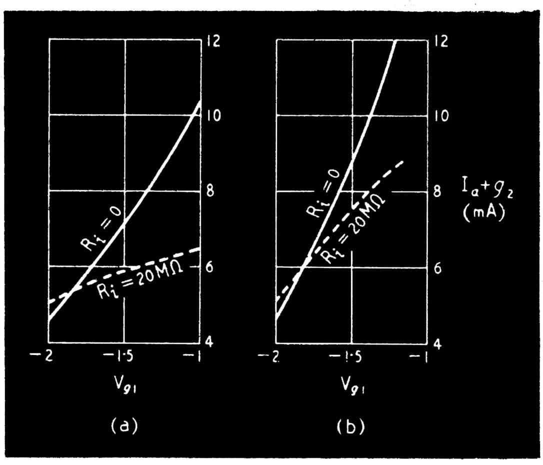


Fig. 2. Portions of characteristic curves of typical triode-connected samples of (a) EF50 and (b) SP61, with and without 20MΩ in series with the source of V_{g1}, showing unsuitability for measuring apparatus with high-resistance input circuit. Results with other types of high-slope valves were, in general, similar.



be seen) ordinary amplifier valves, suggested that there should not be much difficulty in this respect. But there was.

In the original trial circuits, the impedance converter or d.c. amplifier followed the present-day practice of using a twin triode somewhat as sketched in Fig. 1. The merit of this arrangement, of course, is that variations in supply voltage tend to affect both triodes equally, so that the zero setting is not appreciably affected. And the negative feedback due to using the valves as cathode followers has a stabilizing and scale-linearizing influence.

But just now we are considering the input circuit. To observe possible error due to input circuit resistance the effect on the reading of inserting high resistances at R_i was noted. With the original valve, as little as 5MΩ caused a considerably larger change in reading than could be tolerated. A number of other valves were then tried—ancient and modern, separate and combined, triode and pentode. The results varied from indifferent to bad. Even glass-based valves of the EF50 type were not really satisfactory, nor was the valve holder responsible. It must be understood that the types tested were those considered reasonably suitable as cathode followers feeding a meter movement taking 1mA or so full scale. That is to say, they were chosen for a fairly large slope, not less than 3mA/V and preferably more, in the interests of an approach to the other clause of the ideal—zero output resistance. Figs. 2 (a) and (b) show the changes in anode current on inserting 20MΩ in series with the grid bias source, for typical samples of EF50 and SP61 valves respectively. (In fairness to the makers it must be said that such a grid resistance is far above the limit rating.) One notable feature is that with normal heater voltage the grid has to be as much as 1.5-2V negative to cut off grid current. With greater negative bias the error is of opposite sign, due apparently in the main to leakage from the anode. This was especially so with some of the twin-triodes. Assuming that the anode is at +100V, the leakage resistance to make the grid potential 0.02V more positive when 20MΩ is inserted in the grid lead is as high as 10^5MΩ; which shows that the demands of a valve voltmeter are not easy to meet, at any rate with mass-produced high-slope valves.

The conclusion from this was that to fulfil the input requirements it would be necessary to use a type of valve such as the EF37, which was known to have an extremely low input conductance, especially when run at low anode voltage and current and somewhat subnormal heater voltage. It had been found possible to use such a valve to measure resistances up to well over a million megohms. Unfortunately these conditions of type and operation are diametrically opposed to those for low-resistance output. There seemed to be nothing for it but to use two pairs of valves —one for high-resistance input, coupled to the other for low-resistance output.

At this stage we can turn our attention to the output resistance. A valve with a slope of 5mA/V under operating conditions has an output resistance as a cathode follower of about 200Ω. In Fig. 1 there are two in series, making 400Ω. Assuming that the lowest range is 0-1V, and the meter is 0-1mA, and (for the moment) that there is a one-to-one coupling between input and output valve grids, the total resistance in the meter circuit on this range would have to be 1,000Ω. Of this, 400 would be in the valves, and therefore varying inversely as their slope. Al-

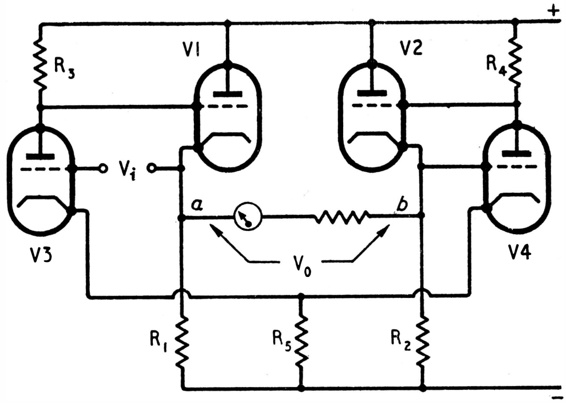


Fig. 3. Outline circuit of system having very high input resistance and almost zero output resistance. V_i is the voltage to be measured, and V_o (the output voltage) is less than 1% lower even when feeding a resistance as little as 500Ω.

though the effect of varying supply voltage on zero setting would largely be balanced out by the symmetrical valve arrangement, the effect on slope and consequently calibration would not, and stabilization would be essential to ensure accuracy under present-day electricity supply conditions. The gradual change in slope with age would also cause error, and valves could not be replaced without recalibration or adjustments of a fairly elaborate kind.

So altogether we seem to be arriving at quite a complicated and expensive piece of apparatus, with a not very attractive performance. A voltmeter of any kind cannot really be considered satisfactory if a substantial part of its resistance on any range depends on the mutual conductance of a valve or valves. One would much prefer to see it all as wire.

Self-stabilizing Circuit

At this point it occurred to the writer that a stabilized power supply can easily be made to have an output resistance (R_0) of a few ohms, or, with a little care, a fraction of an ohm. The type of circuit in mind operates as a cathode follower with amplified feedback. Whereas the output resistance of a simple cathode follower is approximately $1/g_m$, this figure is divided by any voltage gain that is provided in the feedback loop. A valve of the EF37 class is excellent for providing voltage gain. Applying the voltage-stabilizer technique to a balanced valve-voltmeter circuit resulted in the arrangement shown in its simplest form in Fig. 3. V1 and V2 are the output valves. V3 and V4 are the high-input-resistance valves, connected so that their full gain in conjunction with V1 and V2 is fed back. With normal values, the difference between the output and input voltages V_0 and V_i is of the order of only 1 per cent, so a d.c. voltmeter connected between *a* and *b* need not be recalibrated when used to measure voltages applied to the V_i terminals. As will be shown later it is an easy matter to eliminate even this small error.

V1 and V2, being used as cathode followers, are necessarily triode-connected whether or not they are in fact triodes. Seeing that R_0 is now of the few-ohm order, the possible error in calibration due to variations in the characteristics of these valves is quite negligible, and if this consideration were the only one

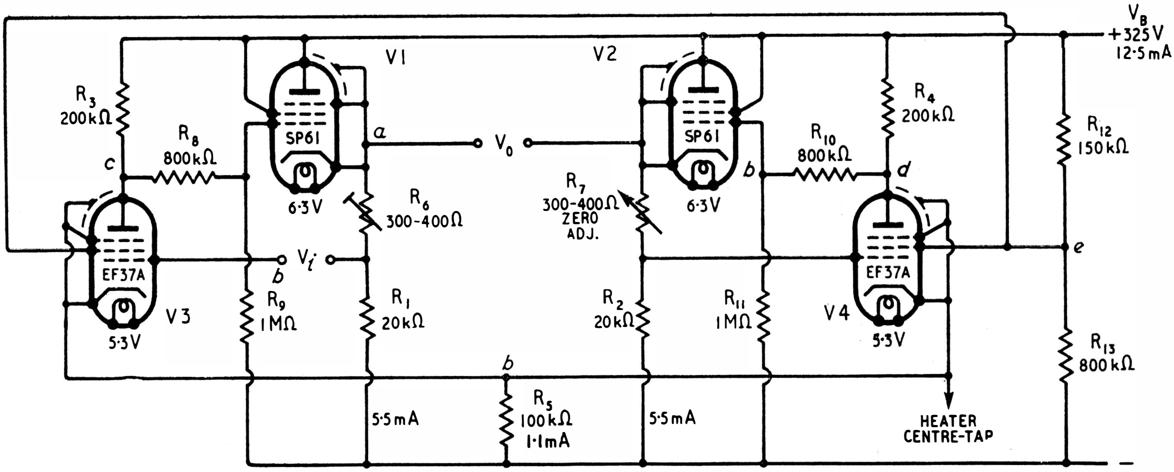


Fig. 4. Circuit diagram of the Fig. 3 type as finally evolved. The heaters of V3 and V4 are underrun to ensure low input conductance.

almost any type of valve could be plugged in without any readjustment or recalibration whatever, even on the lowest range. It is only when designing for the highest range that the characteristics are of much importance. As in designing voltage stabilizers for a large range of output voltage, one aims at high μ in conjunction with low g_m. Two SP61 valves are used in the writer's design, but an ECC33 twin triode can be substituted for them without making any noticeable difference except about 5 per cent. reduction in the highest measureable voltage.

Another advantage of the scheme is that almost any voltmeter can be connected between the cathodes; the full-scale current does not matter, within reason. The meter adopted actually has a full-scale current of just over 3mA. So most makes of multi-range test meters are suitable for use with this "electronic adaptor," which is so stable and error-free that a 5-inch scale can be used to advantage.

The range-changing switch on a multi-range voltmeter is operative when the adapter is in use, but only up to a certain limit depending on the circuit design and the supply voltage. It is here that the action of the circuit must be considered in greater detail.

A full mathematical analysis would probably just be confusing to most readers at this stage. The most helpful way to consider a problem of this kind, perhaps, is as an approximation, supplemented by a good idea of the factors controlling the departures from the approximate. For instance we know already that the output resistance of the system, as a source of d.c. for the meter, is (to a first approximation) zero; and that what resistance there actually is can be made more nearly zero by increasing the total voltage gain round the feedback loop. To a second approximation

$$R_0 \simeq \frac{2}{g_m m}$$

where g_m is the slope of V1 and m is the voltage gain of V3. The 2 comes in because V2 and V4 (assumed to be the same as V1 and V3) introduce a second equal resistance. The gain of a resistance-coupled valve of the EF37 type (including EF36, EF37A, and any similar) is not very large as a triode, but can easily be of the order of 100 in a pentode circuit. Taking $m = 100$ and $g_m = 0.006$A/V, we calculate that $R_0 = 3.35\,\Omega$, which is of the order realizable in practice.

Let us disregard initial voltages in Fig. 3 and take a as the zero point. The usual method of connecting a diode for alternating voltages makes V_i negative; so to begin with let us suppose $V_i = -1$. This negative voltage on the grid of V3 will reduce its anode current, raising the voltage of its anode and lowering that of its cathode. The raised anode voltage will make the grid of V1 (and hence its cathode) more positive, too. This positive movement, being fed back to the grid of V3, will tend to neutralize V_i. In fact, assuming for the moment that the cathode of V3 is somehow kept at a constant potential, and that the gain from the grid of V3 to the cathode of V1 is of the order of 100, only about -0.01V change in the grid potential of V3 would be needed to raise the cathode of V1 by 1 volt. In effect, then, introducing V_i does not lower the potential of its grid end appreciably; it raises its cathode end by practically the full amount. Although we know, therefore, that a change in grid-to-cathode potential necessarily occurs when a voltage V_i is applied, to a first approximation it is negligible compared with V_i.

Now we can consider the drop in cathode potential of V3. It is equivalent to a rise in the grid potential of V4, and tends to increase the anode current of V4. This tends to neutralize the change in common cathode potential of V3 and V4, and also to lower the grid (and consequently cathode) potential of V2. So while point a goes up, b goes down. And since, for the reasons given in connection with V1 and V3, the change in grid-to-cathode potential of V4 is negligible compared with V_i, it follows that the change in potential between the grids of V3 and V4 must also be relatively small. If one were to neglect it altogether then it would follow that V_i plus the output voltage V_0 would always have to be zero; in other words, the voltage reaching the meter would be opposite *and equal* to that applied at the "infinite resistance" input terminals. Which, as Euclid used to say, is what was to be done. The only difference between this ideal and the actual is of the order of 1 per cent caused by the fact that the gain of V3 and V4 is not quite infinite.

We see that when V_i is applied a must rise and b fall, but it is probably not obvious in what proportions. It is obvious that the movement cannot be perfectly symmetrical, *b* falling by the same voltage that *a* rises, because if it were (with matched valves) there could be no change of current through the common cathode resistor of V3 and V4, and therefore V4 would receive no "signal." Which, as Euclid again would have said, is absurd. So it is clear that *b* must fall less than *a* rises, and the common cathode potential of V3 and V4 must fall by about the same amount as *b*. Analysis shows that the rise/fall ratio is approximately $1 + R_3/R_5$, but this result is not very reliable in practice because of complications such as maintenance of g_2 potential in V3 and V4.

These practical complications are now due to be considered. In the first place, Fig. 3 clearly would not work as it stands, because if all the valves had their appropriate grid biases the anode voltages of V3 and V4 would be negative. The first method to be tried for correcting this was to insert resistances of about 6kΩ in the positions R_6 and R_7 in Fig. 4. Fortunately a pentode can work down to a very low anode voltage; say 20. If V_i is always negative, then the initial anode voltage for V3 can be near its minimum, because its movement is always towards positive. The anode of V4 goes negative, but (with typical component values) only about one-third to one-quarter the extent, so if 50 or 60V is to be the top range, it is sufficient to give the anodes of V3 and V4 about 40V, which will leave some margin. But if both negative and positive voltages are to be measured this would not be enough.

The cathode resistor method was found unsatisfactory for at least two reasons. Firstly, if the meter terminals are kept at the cathodes, V_0 becomes substantially greater than V_i and the original calibration does not hold; if they are not, the extra cathode resistors increase R_0 substantially, and the calibration is again upset, though not so badly. Secondly, it was desired to tap off about − 1.5V from R_1 for ohmmeter purposes, and the slight change in the value of R_1 when the ohmmeter terminals are short-circuited shifts the zero slightly when V1 has a cathode resistor "above" R_1. So although it throws away a good deal of the voltage gain, a potential divider (R_8R_9 in Fig. 4) was tried, and worked better than expected.

Although cathode resistors in the positions R_6 and R_7 were abolished because of the disadvantages just mentioned, those disadvantages were due to the high value of resistance (about 6kΩ) required to provide V3 and V4 with anode voltage, rather than to their presence itself. In fact one of the disadvantages—the increase in V_0 for a given V_i—can be turned into an advantage by choosing a much lower resistance for R_6 and R_7. By making R_6/R_1 (and R_7/R_2) equal to the proportionate error $(V_i - V_0)/V_i$ the discrepancy between V_0 and V_i can be completely eliminated. And varying R_7/R_6 is a convenient method of zero setting. The values required for this purpose are too low to cause appreciable trouble in the ohmmeter circuit.

Fig. 4 also shows the g_2 feed for V3 and V4 (R_{12} R_{13}).

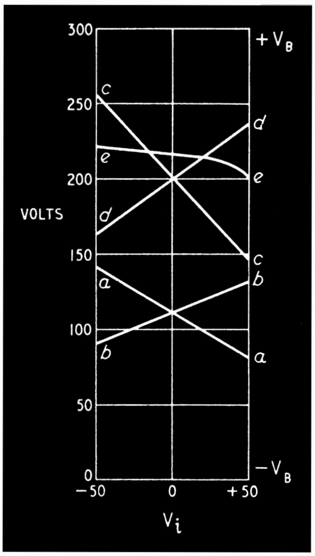


Fig. 5. Voltage diagram relating to Fig. 4, showing how the potentials of the parts marked a to e vary as V_i is varied from −50V to +50V. The differences between grid and cathode potentials of valves, being relatively small, are not shown.

Determining the component values in this circuit to ensure the desired top range, for the minimum supply voltage, is rather complicated, and would distend this article excessively if described in full and in general. But some idea of how a particular specification was arrived at may be helpful. The voltage diagram Fig. 5, obtained by measuring currents in the various leads over a range of positive and negative V_i and multiplying them by the appropriate resistances, is very illuminating. It is drawn for a supply voltage of 300, and shows the potentials of the correspondingly lettered points in Fig. 4. The differences in potential between the various parts marked *b*, which are only a few volts, are neglected in Fig. 5; and similarly for *a*. With the circuit values in Fig. 4, arrived at by several trials, we see that at $V_i = 0$ the anode voltage on V3 (V_{bc}) is about 90, rising to 165 at $V_i = -50$V and falling to 15 at + 50V. Obviously this is the limiting factor for measurement of positive voltage. The limit for negative voltage is set by V3 cutting off, which it would do when *c* reached 270V. The limit in both directions is thus set by V3 because it handles a greater amplitude than V4, so it is desirable to equalize the swing between these valves as much as possible. If, working on the formula $1 + R_3/R_5$, we reduce R_3 we reduce the gain, which is undesirable; so instead R_5 is made as large as practicable. Increasing it increases the current through V1 and V2 and ultimately restricts their swing. An incidental advantage of the potential dividers R_8 R_9 and R_{10} R_{11} is a tendency to equalize the swings; the approximate theoretical ratio is altered from $1 + R_3/R_5$ to $1 + p\,R'_3/R_5$, where p is the step-down ratio $R_9/(R_8 + R_9)$ and R'_3 is the resistance of R_3 in parallel with $R_8 + R_9$. Fig. 5, shows even better equality than theoretical, due presumably to the fact that this simple formula ignores screen-grid current. Note, incidentally, that the "push-pull" operation prevents the screen-grid voltage (V_{be}) from varying as widely as it otherwise would.

Before completing the design it is necessary to check absence of grid current in V3 and V4 at the extremes of input, and if necessary modify component values or V_B if not satisfactory.

A suitable working margin must be be allowed above the maximum scale V_0, to allow for variations in mains voltage and valve characteristics. Beyond the working limits of V_i the system cuts off sharply and there is no further increase in current through the meter, which is thus automatically protected. It is advisable, however, to have it on its highest range when switching on and off, to avoid excessive transient unbalances while the heater temperatures are varying. The type of scale in the meter used made it convenient to have ranges of 0-1.5, 5, 15, and 50V. These were obtained in the usual way by series resistors. An extension of this procedure for higher ranges would have necessitated an unduly high supply voltage, so they were provided by means of an input potential

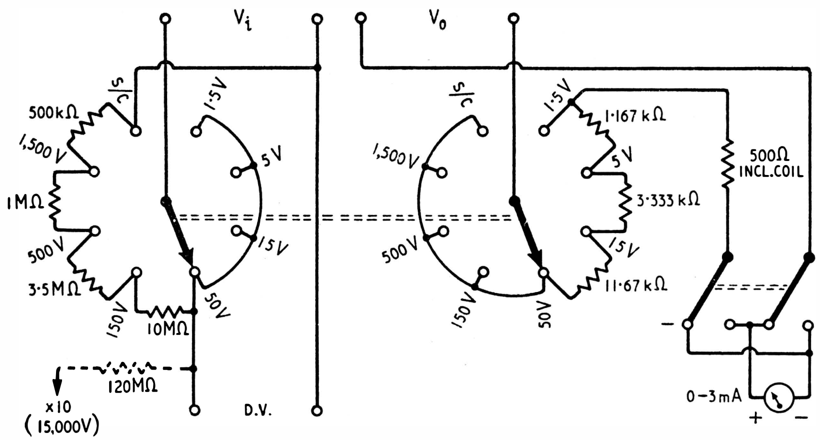


Fig. 6. Details of switching for d.v. (direct voltage) range and polarity, associated with the circuit of Fig. 4.

divider, which places 15MΩ (high-stability carbon) across the input terminals on all ranges. If desired this could be open-circuited on ranges lower than 150V. Fig. 6, which joins up with Fig. 4 at the V_i and V_0 terminals, shows the range switching, including a reversing switch for reading positive voltages when required. The dotted multiplier is suggested for television voltages, etc., but should not be attempted by anyone inexperienced in high-voltage technique, as the chain of resistors totalling 120MΩ must be suitably mounted to avoid corona and leakage.

Resistance Measurements

A valuable feature of a valve voltmeter is the ease with which wide-range ohmmeter facilities can be incorporated. Fig. 7 shows how this was done in the present instance. The arrow marked D.V. leads to the arm of the range switch shown in Fig. 6, the resistance measuring components being cut out in the most clockwise position of the switch above. It is possible to combine the two switches in one, but it would have to be provided with rather a large number of ways. The values in Fig. 7 are based on 5.5mA through R_1, with readings taken on the −1.5V range. This voltage is tapped off R_1 273 Ω from the top, and the 731 Ω shown in Fig. 7 is the amount required to bring it up to 1kΩ (allowing for the rest of R_1 in parallel) to constitute the lowest resistance standard. The other standards are 10kΩ, 100kΩ and 1MΩ. The resistance to be measured is connected to the R_x terminals and forms with the selected standard a potential divider, so the resistance scale is of the usual ohmmeter type in which mid-scale reading corresponds to R_x equal to the standard. www.keith-snook.info

Satisfactory readings are given from 100 Ω to 10MΩ, with rough indications at 10 Ω and 100MΩ. An additional facility is shown for testing insulation: a tapping at −78V in conjunction with 1MΩ across the V_i terminals, giving a minimum reading (at full-scale) with 50MΩ applied to the " Ins." terminals. A separate scale is, of course, necessary for this range. For insulation measurements three termimals are available, for employing the usual guard ring technique to exclude effects due to undesired leakage paths. Satisfactory readings are given up to 1000MΩ, and an indication at 5,000MΩ. The tapping is at 78V to allow for the full-scale drop of 1.5V across the standard, 75V across the 50MΩ, and 1.5V across a safety 1MΩ at the tapping. This last prevents the greater part of R_1 being short-circuited but does not prevent the pointer being driven off the scale, so before using this range a test should be made on the 1MΩ range.

Accuracy of the ohmmeter readings depends on constancy of the current through R_1. Ideally V_B should be stabilized, but since the current can be checked at any time by open-circuiting the R_x terminals it was decided to adjust V_B by hand, leaving enough margin to give full-scale with bottom-limit mains voltage.

Finally, a few data on performance. In spite of half the gain of V3 and V4 being thrown away by the potential divider, the total output resistance is only 4 Ω, so that even on the lowest range and with a meter taking as much as 3mA full-scale, V_0 is only 0.8 per cent less than V_i, and as we have seen this small error is eliminated by R_6 and R_7. A 10 per cent drop in V_B produces a barely perceptible (about 0.004V) shift of zero, and no perceptible change in calibration. A 10 per cent drop in heater voltage produces 0.01V zero shift and no perceptible change in calibration. Even the drastic test of connecting a second SP61 in parallel with V1 was equally satisfactory. The characteristics of V3 and V4 are much more influential; doubling V1 shifted the zero 0.61V, but when this was taken up on the zero adjustment the lowest-range calibration was again not perceptibly altered. In practice the change in characteristics of V3 and V4 even over a long period is likely to be small, and certainly not more than a very small fraction as drastic as this test. Experience so far has shown that even with ordinary carbon resistors, provided that they are conservatively rated and kept away from the heat of the valves, the zero is remarkably stable; but high-stability or wire-wound resistors are to be preferred for the more critical values.

It is hoped at a later date to describe the arrangements for measuring alternating voltage.

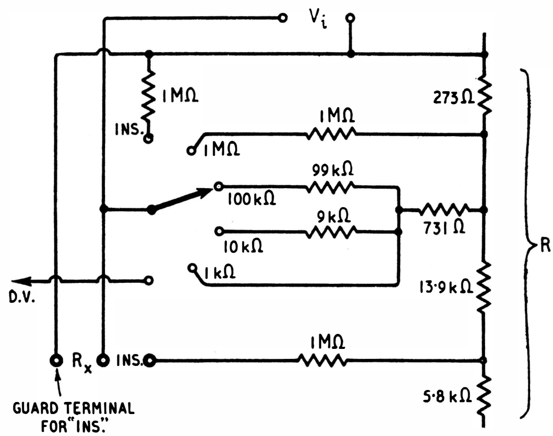


Fig. 7. Details of ohmmeter and insulation testing arrangements associated with Fig. 4.

VALVE VOLTMETER

The Rectifier Section

By M. G. SCROGGIE, B.Sc., M.I.E.E.

Design of Input Circuits for Alternating Voltage Measurement

THE following article is a sequel to the one published in January describing a highly stable valve voltmeter for direct voltages only. The special feature of that instrument, it may be remembered, was that the addition of the valve circuitry to the moving-coil voltmeter did not necessitate any recalibration, since the output voltage of the valve unit was the same as its input, within a fraction of 1 per cent. But although this feature may not be common to valve voltmeters in general, most of them are similar in so far that the rectifier that fits them for measuring alternating voltages is organically distinct from the "d.c. amplifier" (really a resistance converter), so the principles now to be discussed have a much wider application than the particular system already described. The subject, in other words, is the left-hand unit in Fig. 1. This does not, of course, necessarily take the form of a completely detachable unit, though it is often made up as a flexibly connected probe.

Because the circuitry of this part is so very simple, consisting of a diode and one or two resistors and capacitors, it may be supposed that it calls for little consideration. But it is a good rule that anything with a rectifier in it calls for plenty of consideration. What follows does not claim to cover anything like the whole subject of how to rectify with a diode, but it may help to remove some common misunderstandings.

The first thing to consider is the "value" of the alternating voltages to be measured—peak, mean, or r.m.s.? Sometimes one is required and sometimes another, but for most purposes the r.m.s. value is the most appropriate, because it is the one from which the power can be calculated, regardless of waveform. Unless the contrary is specified, alternating voltages and currents are generally assumed to be r.m.s. Unfortunately the types of meter that respond on principle to r.m.s. values generally consume an excessive amount of power for the purposes chiefly in view, or have other limitations and inconveniences. Direct voltages can be measured, using the Fig. 1 technique, with very little disturbance of their source, so for corresponding alternating-voltage (a.v.) measurements the usual practice is to add a rectifier. If a suitable rectifier could be obtained with a natural square-law, the resulting readings would be r.m.s.; but the best that can be found is an approximation to a square-law over a very limited range. So for wide ranges of measurement the r.m.s. ideal is generally abandoned.

Instead, the aim is a "perfect" rectifier—one with zero (or at least negligible) forward resistance and infinite back resistance, like a switch. With such a rectifier connected in series with a high resistance, the average direct voltage across the resistance is equal to the average alternating voltage of the active half-cycles. If the waveform is sinusoidal, the average of a half-cycle reckoned over the period of half a

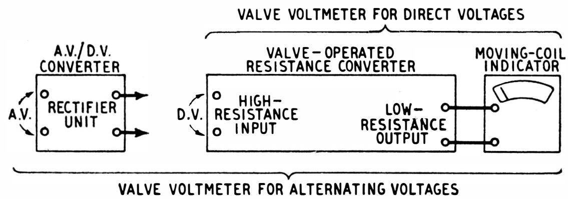
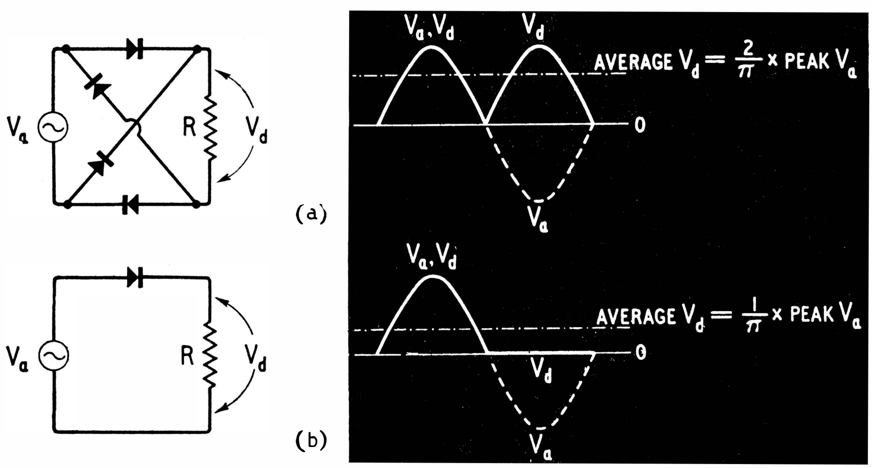


Fig. 1. Functional arrangement of the general class of valve voltmeters considered, showing inputs for direct voltage (d.v.) and alternating voltage (a.v.).

Fig. 2. Comparison of input and output voltages with (a) full-wave and (b) half-wave non-cumulative rectifier. The symbols V_a and V_d, here and in the text, do not imply any particular value—instantaneous, peak, r.m.s., or average.



cycle is $2/\pi$ or 0.637 of the peak value, so is 0.637/0.707 or nine tenths of the r.m.s. value. With full-wave rectification this ratio holds over the whole cycle (Fig. 2(a)) so if the whole voltage across the resistor is applied to a d.v. valve voltmeter and the reading multiplied by 10/9 the result should be the r.m.s. value of the alternating voltage. But if there are harmonics in the waveform the resulting errors are likely to be of the same order as the percentage of harmonics.

Using half-wave rectification (Fig. 2(b)) each half-cycle is averaged over a whole cycle, so the readings are half as much—with sine waves, less than one third of the peak value.

Not only is the reading with a simple half-wave rectifier rather low, but at very high frequencies the stray capacitance across the load resistance (R) retains sufficient charge to be appreciable during the inactive half-cycle, so causing the reading to be higher at high frequencies than at low. The way out of these difficulties is to add so much capacitance (C) across R that the charge is almost fully retained at *all* working frequencies, down to the lowest. The result is the familiar " cumulative " diode rectifier, used in most radio receivers as well as valve voltmeters. If the rectifier really were ideal and had no forward resistance, C would charge up to the full peak value during the first half-cycle. From then until the next half-cycle it would leak through R, but provided that R and C are so large that their time-constant RC is very long compared with the period of a cycle the loss is small and the reading is very nearly equal to the peak value, which with sine waves is $\sqrt{2}$ or 1.414 times the r.m.s. value.

It might seem, then, that all that need be done is to connect a suitable rectifier in series with R and C (Fig. 3) and apply 70.7 per cent. of the voltage across R to the direct-voltage instrument described in the January issue, which would then read r.m.s. values (so long as the voltages being measured were sinusoidal).

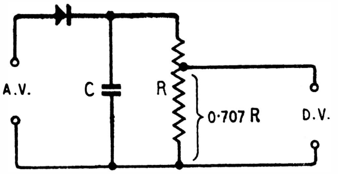


Fig. 3. Theoretical circuit of cumulative rectifier giving r.m.s. readings of sinusoidal voltages

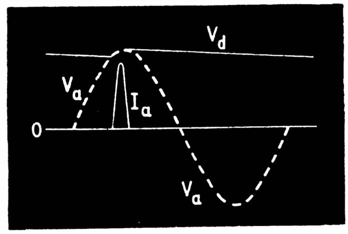


Fig. 4. Input/output diagram for cumulative rectifier, when the charging resistance (r) is much less than the discharging resistance (R).

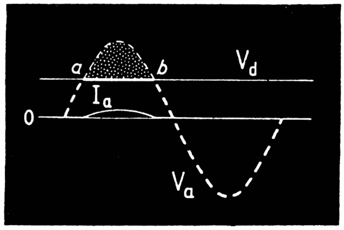


Fig. 5. If r is adjusted to reduce the output to one half, the diagram (compare Fig. 4) is altered to this.

Effective Input Resistance

In practice it is not quite so simple, however. The requirement from which all the complications spring is that the measuring system must present a very high impedance to the a.v. source. A perfect rectifier is at all times either conducting completely or not at all. Nearly all the time the charge on C biases it back so that it is not conducting, and except for any stray capacitance in parallel with the rectifier the input impedance is infinite. But at the very peak of the active half-cycle the input voltage momentarily exceeds the voltage to which C is charged (because since the last peak there has been a slight leakage through R), so the rectifier conducts and restores the charge to full peak voltage. If the forward resistance of the rectifier really were nil, the charging process would be instantaneous and the charging current infinite ; in reality, of course, charging takes an appreciable time, but as it is small compared with the whole cycle the current is correspondingly large compared with that which discharges through R (Fig. 4). For example, if the peak V_a were 10V, and R were 100kΩ, the leakage current would be nearly 0.1mA ; and if charging occupied one twentieth of the cycle the charging current would have to average $20 \times 0.1 = 2$mA over that period. So although most of the time the input impedance would be infinite, during 0.05 of each cycle it would be about $10/2 = 5$kΩ.

If one insists on a single figure for input resistance, the most reasonable reckoning is on a basis of power dissipation. The forward resistance of the rectifier being neglected, the only dissipating component is R, and the wattage is V^2_d/R. V_d being practically equal to the peak V_a, or (with sinusoidal input) $\sqrt{2}$ times the r.m.s. V_a, V^2_d/R can be taken as equal to $2V^2_{a(r.m.s.)}/R$. If the effective a.c. input resistance is denoted by R', it must by definition be such that the dissipation is $V^2_{a(r.m.s.)}/R'$. By equating the dissipation expressed in these two different ways, R' is seen to be equal to R/2, subject to the approximation assumed.

When the source of V_a is a sharply resonant circuit, the flywheel effect of the circuit keeps the input waveform reasonably sinusoidal, and although the damping due to the voltmeter is concentrated into a small fraction of each cycle, its effect is practically the same as that of a resistor of value R/2. In the example considered, $R' \simeq 50$kΩ, and if the dynamic resistance of the tuned circuit were also 50kΩ the effect of connecting the rectifier circuit would be to reduce V_a by one half, without causing much change of waveform. The tuned circuit is analogous to the pendulum or balance-wheel of a clock, which, though it bears a very uneven load during its cycle, has enough stored energy to maintain a smooth, regular swing.

But when the source is a non-resonant generator with internal resistance, such as the output of a valve with resistance coupling, the effect is quite different. During the active part of the cycle the source resistance (say 50kΩ) limits the charging current to an amount which clearly cannot be many times the discharge current through 100kΩ, so must spread over a considerable proportion of the cycle. Consequently the simplifying assumptions break down and calculation becomes much more difficult. At the same time practically the whole of the enlarged active peak is cut off by the voltage drop in the 50kΩ, so the waveform at the voltmeter input is highly distorted.

Fig. 5 shows in thick line this waveform when the source resistance (call it r) is adjusted to reduce the rectified voltage to one half, CR being assumed sufficiently large to maintain it very nearly steady. Current is flowing from the source during the period a to b, and since a sine wave reaches half its peak at 30° and 150° this period is 120° or one third of the whole cycle. The voltage driving this current through r is represented by the shaded area, and can be averaged over the whole cycle by measuring this area and dividing it by the length representing the cycle period. It is about 0.11 times the sine peak. The mean rectified voltage (driving the discharge current through R) having been adjusted to half the sine peak, the mean charging voltage is 0.22 times the discharging voltage. Since the mean charging and discharging currents must be equal, the ratio of r to R must be 0.22—not 0.5 as with the resonant source.*

Alternative Circuits

Experiments with an oscilloscope confirmed all the foregoing statements, and brought out very clearly that with a non-resonant source the distortion and the drop in output voltage are considerable even when r is less than 1% of R. This fact is of great importance in valve voltmeters, but before following it up the reader might note input resistance as an example of how this apparently simple circuit is commonly misunderstood. In one textbook it is stated that the input resistance (at 60 c/s) is equal to the value of series resistance required to drop the full-scale deflection to one half. Without any guidance to the contrary the reader might easily assume that input resistance found in this way (presumably with a non-resonant source) meant the same thing as input resistance calculated on a power-dissipation basis, or measured with a resonant source. If so he would be seriously misled.

This is not the end of confusion about input resistance as will soon be seen; but in the meantime it should be noted that with the Fig. 3 type of rectifier connection the source resistance r includes the forward resistance of the rectifier. So even if the source itself has no resistance, or it is allowed for by regarding the voltage to be measured as the voltage at the terminals of the instrument, there is still a loss or error due to this forward resistance. Fig. 6, which is a generalization of the half-drop example considered with Fig. 5, shows that to get the rectified voltage within 1% of the peak input it is necessary for R to be something like 2,000 times the rectifier forward resistance. Now although the forward resistance of a thermionic or a germanium diode with inputs of at least several volts may be of the order of 100 Ω, so that for the high ranges this source of error could be kept negligible by making the load resistance R no more than one megohm, at low voltages the rectifier resistance rises to thousands of ohms, so that the values of output and peak input rapidly diverge as the input is reduced, causing the well-known bend at the foot of the calibration curve. Even if one were content to put up with considerable curvature by using a 1-MΩ load resistance, one would have to face the fact that connecting the valve voltmeter would cause an appreciable drop in the voltage to be measured, even when the resistance of the source was in only the hundreds of ohms! So the oft-repeated claim for valve voltmeters in general, that they can confidently be used for measurements on high-impedance circuits, can hardly be accepted without reserve—even before starting to consider disturbance due to input capacitance. Taking this together with the fact, emphasized in the previous article, that only exceptional amplifier valves will allow more than a megohm or two in their grid circuits without appreciable drop therein due to leakage current, one sees that valve voltmeters demand a good deal of care in design if they are to come anywhere near their popular reputation.

Connecting the rectifier as in Fig. 3 has one important advantage: the output voltage is almost perfectly smooth. It is taken for granted that in order to avoid low-frequency error the time constant CR must be many times the period of one cycle; for example, if the lowest frequency to be measured is 20 c/s, the period is 1/20, and the choice of CR would be at least 0.5 and probably 1—say 10MΩ and 0.1 μF. The ripple on the output would then be so small as hardly to need additional smoothing. Fig. 3 is not unusual if the rectifier is a germanium diode, which has many advantages. But unfortunately the backward resistance of germanium diodes obtainable up till now is far too low if the load resistance is required to be of the ten-megohm order, for it is

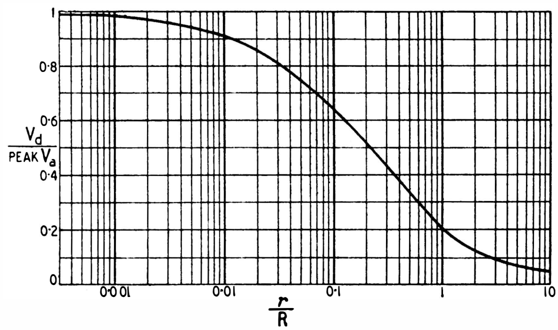
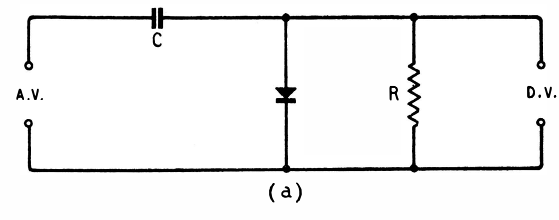


Fig. 6. Effect of the ratio r/R on rectification. (Data by D. A. Bell).



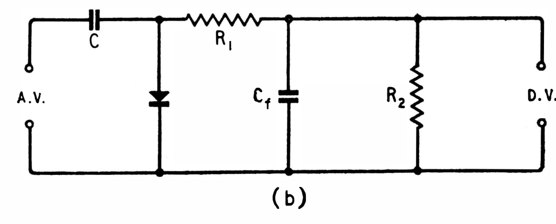


Fig. 7. Series-C modification needed when the voltage source includes a direct voltage or an open circuit. Owing to the large a.v. component passed on, a filter is desirable (b).

* This subject is discussed in greater detail by D. A. Bell in "Diode as Rectifier and Frequency-Changer," *Wireless Engineer*, Oct. 1941, pp. 395–404.

effectively in parallel. For high-impedance instruments the only kind of rectifier is the thermionic diode, hereafter to be called the diode. If it is connected as in Fig. 3, and R is many megohms, there may be undesirable effects from the heater circuit. With a carefully insulated heater winding, however, it should not be impossible to exclude such effects. Stray capacitance is harmless, for it comes across C. But with any type of rectifier the Fig. 3 arrangement is restricted to uses in which there is a highly conductive path through the source, and no direct voltage. To avoid these restrictions the circuit is usually rearranged as in Fig. 7(a). Here the d.v. output is accompanied by the full alternating component. This ought to be removed, because even if one does not measure voltages of sufficiently low frequency to make the pointer vibrate there is risk of overloading the amplifier. So some sort of filter is required. Seeing that there is (as Fig. 3 indicates) some output voltage in hand, there is the obvious possibility of using the series part of R as a filter element (Fig. 7(b)). If rectification were 100 per cent, the ratio of R_1 to R_2 to make the output equal the r.m.s. value of sinusoidal input would be $(\sqrt{2} - 1) : 1$. With rectification less than 100 per cent (as it must be), R_1 could be reduced to compensate—if the rectification percentage would oblige by remaining constant.

Although the Fig. 7 arrangement is broadly equivalent to Fig. 3 the fact that the approximate effective input resistance (to resonant sources) is no longer R/2 is often overlooked. This can be seen in Fig. 7(a) by noting that in addition to providing the same d.c. through R as in Fig. 3, the source also has to provide an a.c. through R (the impedance of C being negligible). By the theorem of superposition these can be reckoned as if they flowed separately. So the effective input resistance is the original R/2 in parallel with R, making R/3.

But when the circuit is complicated by filtering, the input resistance is modified accordingly. Fig. 7(b) is one that might be used to fit the amplifier already described As before, input resistance is half d.c. resistance in parallel with an a.c. resistance. The d.c. resistance is $R_1 + R_2$, and a.c. resistance (C, being large enough for its impedance to be neglected) is R_1. Input resistance is therefore $\frac{R_1\ (R_1 + R_2)}{3R_1 + R_2}$: if $R_1 = (\sqrt{2} - 1)\ R_2$ it is roughly one quarter of R_2.

Some more usual arrangements are shown in Fig. 8. If it is desired to pass the whole rectified voltage to the amplifier, it can be said for (a) that the filter is not appreciably loaded, so the effectiveness of R_1 can be increased by using a higher value. But in (b), where R_1 and R_2 form a potential divider as in Fig. 7, R_3 seems to serve no useful purpose and only reduces the input resistance.

Zero Displacement

Besides having an almost infinite backward resistance and a reasonably low forward resistance (given at least several volts input), the thermionic diode has the advantage of not being easily destroyed by excessive input voltage. The usual maximum rating for suitable diodes is 420V peak inverse, which in practice means $420/2\sqrt{2}$ or 150V r.m.s. input. It is rather useful if the a.v. ranges go up to 250 to include mains voltages, and some makers of valve voltmeters apparently run the risk of exceeding the diode's rating. If the instrument is mainly for high voltages, at not more than about 10Mc/s, the type of diode used for flyback e.h.t. (such as the EY51) is very suitable, as it is quite safe up to at least 8kV peak input.

But it is the low-voltage end that is usually most important, and most difficult to arrange. For not only is there the awkward "bottom bend" but thermionic diodes have the unfortunate complication of "zero current." This is the current that flows in the anode circuit when there is no applied voltage. What matters most is the voltage, negative at the anode, that this current sets up across the load resistance. The current itself varies a good deal from one sample to another of even the same type of diode, but in any given sample depends mainly on the heater voltage. Over a considerable range it is roughly proportional to it, as shown in Fig. 9, which refers to a typical diode. Both current and voltage obviously depend also on the resistance in the anode circuit, and it is an interesting experimental fact that over a very wide range of resistance the voltage is very nearly proportional to the logarithm of the resistance, as shown in Fig. 10 for the same diode. Note that the slope of these nearly straight curves is only very slightly affected by heater voltage.

Now the zero displacement resulting from this effect must obviously be corrected in some way. And since the required correction drifts so steeply with heater voltage,

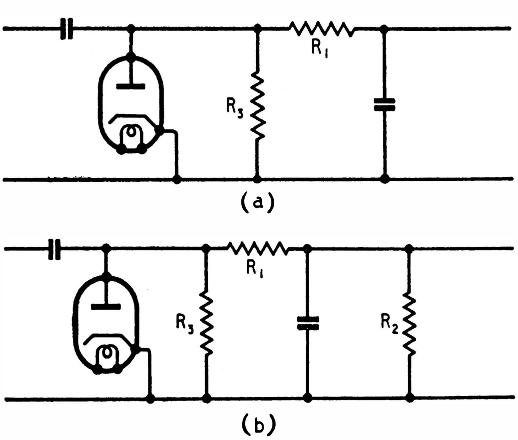


Fig. 8. Alternative load and filter circuits.

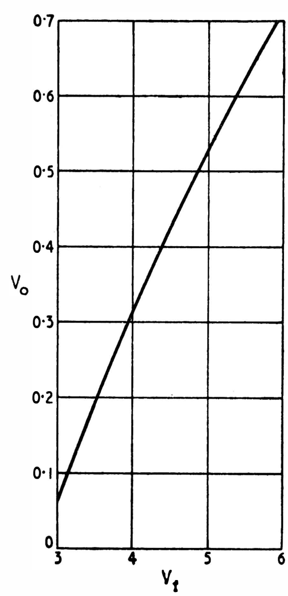
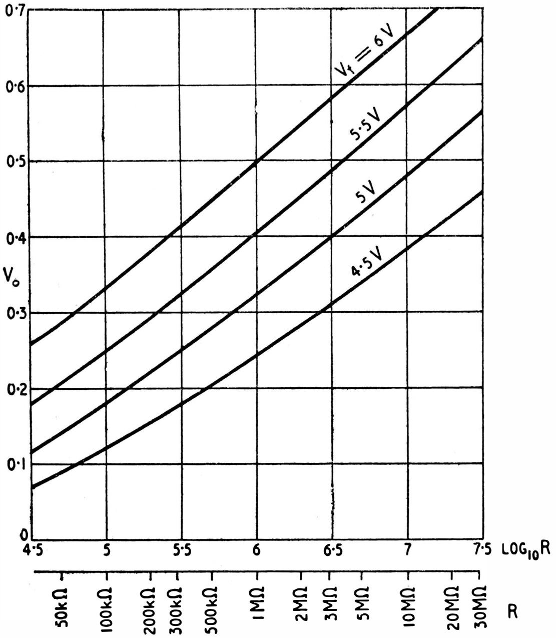
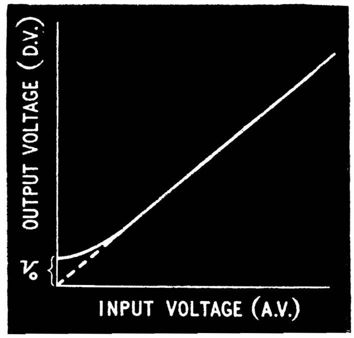


Fig. 9. Variation of "zero voltage" (Vo) with heater voltage (Vf) of typical diode, with a load resistance of 20MΩ.



Left : Fig. 10. Variation of V_o with load resistance (R), for several values of heater voltage.

Right : Fig. 11. Method of arranging for the greater part of an a.v. calibration curve to fit the d.v. scale.



the usual method is to balance it out with another diode of the same type, in conjunction with a preset adjustment to take up any initial difference between the two. This method fits very well into the instrument already described, which is symmetrical. The second diode, which can conveniently be one of the same pair as the first in a double diode, can be connected to the input of the balancing amplifier valve. Fig. 9 shows that the zero displacement voltage can be much reduced by running the heater at, say 4.5V instead of the rated 6.3V, and experiment shows that, with this type of diode at least, and with 10-20MΩ load resistance, the calibration curve is displaced bodily with hardly any change of shape ; but Fig. 9 also shows that *variation* of displacement (which is what matters) is not reduced, so there is really nothing to be gained by underrunning the heater.

Non-linearity at Low Voltages

And now at last it is necessary to deal with the most troublesome feature of all—the bottom bend. This, as we have seen, exists because at low input voltages the forward resistance of the diode cannot be neglected in comparison with the load resistance (in parallel with the backward resistance, if that is not infinite). One result is that even with the highest practical load resistance it is not possible to measure voltages below about 0.1 at all. Another is that up to at least several volts the curvature prevents one from using the same linear scale for all ranges. The only effective remedy is pre-diode amplification, and that calls for very skilful design to avoid drastic reduction in accuracy or frequency range or both. Even with the most modern wide-band amplifier technique one could hardly hope to cover 20c/s to 200Mc/s with good accuracy, or even at all. In any case it is a subject in itself.

An attractive way of dealing with the matter is that proposed by R. Kitai*, who deliberately arranges for a certain amount of zero displacement, v_0 in Fig. 11, so that when the curve does straighten out it will coincide with a straight line through the origin, and by suitable adjustment of the slope can be made to fit the common linear scale, at least from the point where the curve joins the straight line. Mr. Kitai places this point at 0.2V, but this seems over-optimistic, for neither the type of diode specified by him nor any other tried by the present writer has been found to approximate reasonably closely to a straight line down to such a low voltage, even with 20MΩ load resistance. Not only so, but when an accurately-drawn input/output voltage curve is examined it can be seen that there are appreciable differences in slope and zero displacement between the straight lines that best represent even the upper ranges of it. Except for rough work, then, a single preset adjustment of slope and zero displacement for all ranges is not good enough.

Scale Fitting

There are two alternatives. One is to abandon the idea of using the same linear scale for direct-reading a.v. as well as d.v., and to use a calibration curve. If you agree that almost anything is worth while to escape from a calibration curve, you will prefer the second alternative, which is so to contrive the range switching as to fit the curve as nearly as possible to the existing linear scale on all ranges. This will cause the pointer, set to zero for d.v., to move up from zero on a.v. The only range on which this matters is the lowest, because on all the others the first 30 per cent or so of the scale is better served by a lower range. The top readings on the d.v. ranges in the meter already described are 1.5, 5, 15, and 50 ; so the required extent of the ranges is 0–1.5, 1.5–5, 5–15, and 15–50. The rectification characteristic of a normal diode can be fitted with satisfactory accuracy to all except the first, which must be treated as if there were a still lower range by disregarding all below about 0.5 V. For these lower voltages one must either use a calibration curve or embody it in the meter by marking an extra scale.

The first step is to make as accurate calibration curves as possible for at least the first three ranges, using the Fig. 7(b) circuit with R_2 equal to what has already been chosen for the amplifier (15MΩ in the instrument already described) and R_1 equal to $\sqrt{2}-1$ times as much (6.2MΩ in that example). If the diode were a perfect rectifier, and the input were perfectly sinusoidal, all the curves would be straight lines with a 1 : 1 slope, and all the readings would be the correct r.m.s. values of the input voltage. In reality, of course,

* *Electronic Engineering*, Oct. 1950, p. 420.

the slope gradually approaches 1, and the best straight lines representing the curves over the effective parts of the ranges cut the vertical axis at different points above zero. The following are the results of a test on an EB91 :

Range, V	Slope	v_0 (Fig. 11)	Suitable value of R_1 (MΩ)
0.5—1.5	0.92	0.13	4.5
1.5—5	0.97	0.18	5.6
5—15	0.98	0.22	6.0
15—50	0.99	0.31	6.1

The slopes can be brought up to 1.00 by making the range switch reduce R_1 to the values shown in the last column. This, incidentally, varies the zero setting in two opposite ways: by varying the total zero voltage from anode to cathode (see Fig. 10) and by altering the proportion of that voltage passed to the amplifier. The net effect is small, and in any case can be taken care of by the second range-setting adjustment—that needed to provide the appropriate v_0. To do this, and at the same time retain the full compensating effect of the balancing diode, one would have to provide a complete duplicate of the first diode circuit with its range switching, plus a switched preset v_0 from some d.c. source. But fortunately a much simpler system can be devised to give results that are practically indistinguishable even with high-grade equipment.

The slope is adjusted by varying R_1, as suggested; and the appropriate v_0 for each range is derived by making the balancing diode yield just that much less than the zero voltage of the signal diode. This is done by reducing its coupling resistance, in accordance with Fig. 10. It is reduced sufficiently to meet this requirement even though (with the object of making a single untapped resistor do for each range) the whole of the voltage is passed on to the amplifier, and not only about 70 per cent of it as with the signal diode. If the curves in Fig. 10 were perfectly parallel the result would be that the balancing voltage would vary with heater voltage at a rate 40 per cent higher than the zero voltage to be balanced, but it conveniently happens that the converging of the curves towards the left just about offsets this effect over the range of resistance needed to provide v_0.

The circuitry for both diodes, designed on these lines, is shown in Fig. 12, with the adjacent parts of the direct-voltage unit sketched in lightly. Of course there are other ways of arranging the range-adjusting resistors. The values found correct in a particular case are specified as a guide. The procedure for final adjustment is rather like that for tracking a superhet. With a rheostat to provide the balancing diode load (R_b), and another for the variable part of R_1, R_b is set to the estimated value for the range, and R_1 adjusted, if necessary, to make the instrument read correctly at full scale. Then the reading at about one-third scale is checked. If it is high, the slope is not enough, and R_1 must be reduced. It is adjusted until the difference in reading at full and one-third scale is correct; R_b is then readjusted to take up any equal displacement of these points. On the lowest range the best tracking is actually obtained if the instrument reads very slightly high at these points and a similar amount low at about two-thirds scale. When the best settings have been obtained for all ranges the rheostats can be replaced by suitable fixed resistors, unless one prefers the refinement of preset components and can tolerate their bulk and expense. They need not be adjusted very precisely, especially on the upper ranges.

To check the adequacy of heater-voltage compensation with this simplified circuit, calibration was carried out very carefully over the range 0–1.5V, with the heater voltage of the EB91 double diode adjusted first to 6.0V and then to 4.5V. The difference caused by even this 25 per cent drop varied from imperceptible at the ends of the scale to barely perceptible (0.003V) at the middle. This of course is well below the probable error from other causes.

The need for the source of calibrating voltage to have a low resistance is emphasized by the fact that a perceptible drop in reading on the 50-V range occurs when 2kΩ is inserted in series, and 1 per cent with 10kΩ. Errors of several per cent can be caused by waveform distortion in a source transformer, particularly if controlled by a series primary resistance.

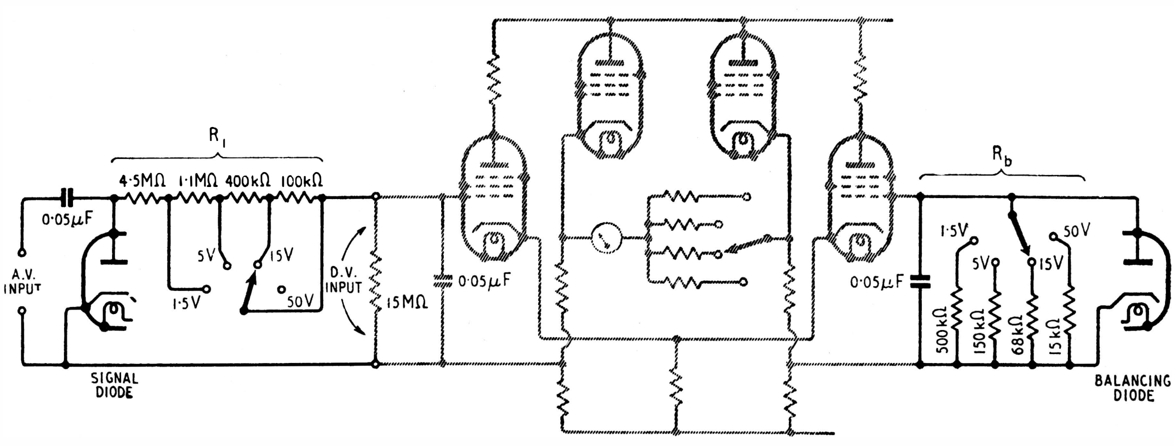


Fig. 12. Complete circuit diagram of a.v. adaptor, showing connections to the partly drawn d.v. voltmeter previously described (Jan. issue). Typical resistor values are shown for the EB91 double diode.