Amplified Automatic Gain Control

Simple Circuit Using High-slope Pentodes with Suppressor-grid Injection

By S. W. AMOS,* B.Sc. (Hons) and G. G. JOHNSTONE,* B.Sc. (Hons)

THIS article describes an amplified automatic gain control circuit which can be embodied in a t.r.f. or superheterodyne receiver and is no more complicated than circuits currently used for non-amplified a.g.c.; no additional valves or negative h.t. supply are required. An advantage of the circuit is that the controlled valves may be high-slope pentodes giving higher gain than is available from the variable-mu pentodes used in conventional a.g.c. circuits. The performance obtainable from the suggested circuit can be illustrated by the following measurements made on a t.r.f. receiver with a single controlled valve. As the input was increased from 100 μV to 0.3 volt, the output increased by less than 2 db; in a typical 4-valve superhet with two controlled stages but non-amplified a.g.c. the output increases by approximately 15 db for the same range of input signal amplitude.

The circuit to be described embodies suppressor-grid injection of the control voltage. The application of a negative bias to the suppressor grid of a pentode produces markedly different effects from a negative potential applied to the control grid. The effect of varying the control-grid potential is to vary the total (i.e. anode + screen) current and the curve of total current plotted against control-grid potential is similar to that of a triode. The effect of varying the suppressor-grid potential however is to vary the ratio in which the total current is divided between the screen and the anode. This is illustrated in Fig. 1, which shows that increase in negative suppressor bias decreases the anode current and increases the screen current, and provided the screen-cathode potential is maintained constant the total current is substantially independent of suppressor-grid potential. By negatively biasing the suppressor grid the anode current is reduced but—more important in the circuit to be described—the change in anode current per volt change in grid potential is also reduced in the same ratio; in other words the mutual conductance of the pentode is reduced by application of negative bias to the suppressor grid.

Thus the application of negative bias to the suppressor grid of a pentode has an effect similar to that of applying negative bias to the control grid of a variable-mu valve. There are, however, two significant differences to be noted; first, the mutual conductance of a pentode can be reduced to zero by a sufficiently large suppressor bias but the mutual conductance of a variable-mu valve can only be reduced to a low value by application of control-grid bias. In other words the anode-current/suppressor-potential curve has a sharper cut-off than the anode-current/control-grid potential curve for a variable-mu valve. However, no distortion occurs even if the valve is biased near suppressor cut-off because the control grid is still modulating the full cathode current. Second, the cathode current of a variable-mu valve is reduced to a very low value by a large negative bias on the control grid, but the cathode current of a pentode is substantially unaffected by application of a large bias to the suppressor grid. This constancy of cathode current with variation of suppressor-grid bias plays an essential part in this a.g.c. circuit and is largely responsible for the simplicity of the circuit. The use of suppressor injection enables high-slope pentodes of the EF50 type to be used as automatic - gain - controlled r.f. or i.f. amplifiers.

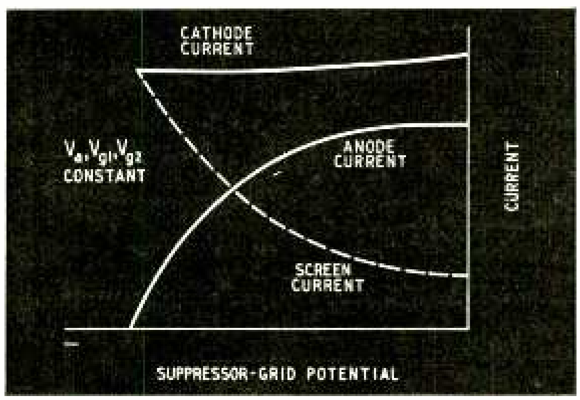


Fig. 1. Typical curves illustrating variation of screen and anode current with suppressor-grid potential.

A disadvantage of suppressor-grid control for valves of the EF50 type is that the suppressor-grid base is usually longer than the control-grid base of a variable-mu pentode, and a.g.c. voltages of the order of 50 volts are necessary for effective control. Thus d.c. amplification of the a.g.c. voltage is essential for effective control. This amplification can be achieved without the use of an additional valve, and successful circuits have been constructed in which the necessary d.c. amplification has been obtained (without effect on their normal function) from (1) an anode-bend detector, (2) the first a.f. stage following a diode detector, (3) a second r.f. stage, and (4) an i.f. stage.

A.g.c. amplifiers are, of course, commonly used in superheterodyne receivers more elaborate than the typical 4-valve type but most circuits require a negative h.t. supply to enable the quiescent anode potential of the d.c. amplifier to be slightly negative with respect to the controlled-valve cathodes. The provision of a negative h.t. supply is disadvantageous because it necessitates additional smoothing components, and possibly requires a further rectifier.

The necessity for such a supply can be avoided if the cathode potential of the controlled valve is made

* B.B.C. Engineering Training Department.

sufficiently positive with respect to earth potential for the control electrode to be directly strapped to the anode of a d.c. amplifier with its cathode approximately at earth potential. A high value of controlled-valve cathode resistance is required to give a high cathode potential and if the a.g.c. voltage is applied to the control grid the large negative d.c. feedback produced by the cathode resistor to a large extent nullifies the d.c. amplification. In other words when the control grid is driven negative the cathode current falls, resulting in a fall of cathode potential so that the resulting change in grid-cathode potential is relatively small. It is for this reason that negative h.t. supplies are almost universally employed in receivers where it is desired to amplify the a.g.c. If, however, the bias is injected into the suppressor grid, loss of a.g.c. voltage does not occur, because the cathode current is virtually independent of the suppressor bias.

Thus a skeleton form of a practical circuit is that shown in Fig. 2 in which the d.c. amplifier V2 is also an a.f. voltage amplifier, the a.g.c. voltage being injected into the control grid as shown. V1 is a controlled valve and its cathode potential together with the anode potential of V2 are both in the region of 100 volts. This high cathode potential has two advantages : it avoids the necessity for a negative h.t. supply as already explained ; it also reduces the screen-cathode potential of V1 to a value (150 volts approximately) at which the maximum safe screen dissipation is not exceeded even when all the cathode current is flowing to the screen, i.e., when the receiver is tuned to a strong signal. A separate screen dropper must not be used because the screen potential will then vary with suppressor bias and the cathode potential will also vary. If for any reason additional decoupling is required, the dropper should be made to carry the anode current in addition to the screen current to maintain correct performance.

If V2 is to have high d.c. gain its associated component values must be chosen with care. Typical values for anode load, screen dropper and cathode resistor for an a.f. voltage amplifier are 100kΩ, 330Ω and 2kΩ respectively. A stage with such components will probably have an a.f. gain of approximately 150 times, but the d.c. gain would be very much less and is unlikely to exceed 10 times. This low d.c. gain is due to negative d.c. feedback caused by the resistance in the cathode and screen circuits. When a negative potential is applied to the grid the cathode current falls, producing a drop in cathode potential and the screen current falls causing a rise in screen potential. These potential changes both tend to maintain the anode current constant, thus reducing the d.c. gain.

The ratio of the d.c. gain to the a.c. gain in such a circuit can be calculated from the expression applicable to any negative feedback circuit, namely,

$$\text{d.c. gain} = \frac{\text{a.c. gain}}{1 + gR}$$

The degeneration due to the cathode resistor R_k can be calculated by putting $R = R_k$ and $g = g_m + g_s$ where g_m is the mutual conductance and g_s is the screen conductance, i.e., the change in screen current for a 1-volt change in screen potential. If $g = 1$ mA/V and $R_k = 2\text{k}\Omega$, practical values for a high-slope pentode with a high-value of load resistor, the d.c. gain is one-third the a.c. gain.

In addition, however, there is degeneration due to the screen circuit and to calculate this g is put equal to g_s and R to R_{sg}, the screen feed resistor. For a voltage amplifier typical values for g_s and R_{sg} are 0.01 mA/V and 330 kΩ respectively. Substituting these in the above expression gives the d.c. gain as 1/4 of the a.c. gain. If degeneration is present in both cathode and screen circuits as in practice, the d.c. gain is likely to be only about 1/13 of the a.c. gain. This calculation, based on conservative values of g, emphasizes the need for low values of screen and cathode resistance in d.c. amplifiers.

The cathode potential may be stabilized by using a low-value cathode resistor through which is passed a constant bleed current which, to avoid unnecessary waste of h.t. current, can sometimes be the cathode current of other valves. The screen potential may be stabilized by feeding it from a low-resistance potential divider connected across the h.t. supply but this again implies waste of h.t. current. This loss may be avoided but the stabilizing effect retained by using a potential divider already present in the circuit, namely, the controlled valve V1 and its cathode resistor. The cathode current of the controlled valve may also be used to stabilize the cathode potential of V2. By this means a d.c. gain of the order of 100 times may be obtained from a voltage a.f. amplifier.

For correct polarity of a.g.c. voltage the grid of the d.c. amplifier must be driven positive by the d.c. component of the incoming signal; in other words if a diode detector is used its cathode must be connected to the control grid of the d.c. amplifier. For this reason a multiple valve such as a double-diode-pentode cannot be used for detection and d.c. amplification. A full version of this a.g.c. circuit is given in Fig. 3 in which the first a.f. amplifier also functions as d.c. amplifier.

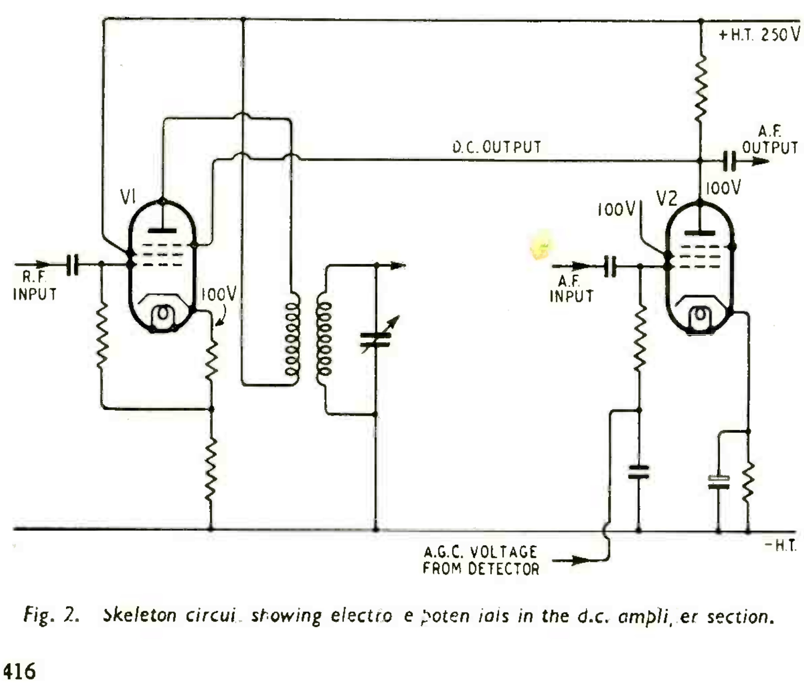


Fig. 2. Skeleton circuit showing electrode potentials in the d.c. amplifier section.

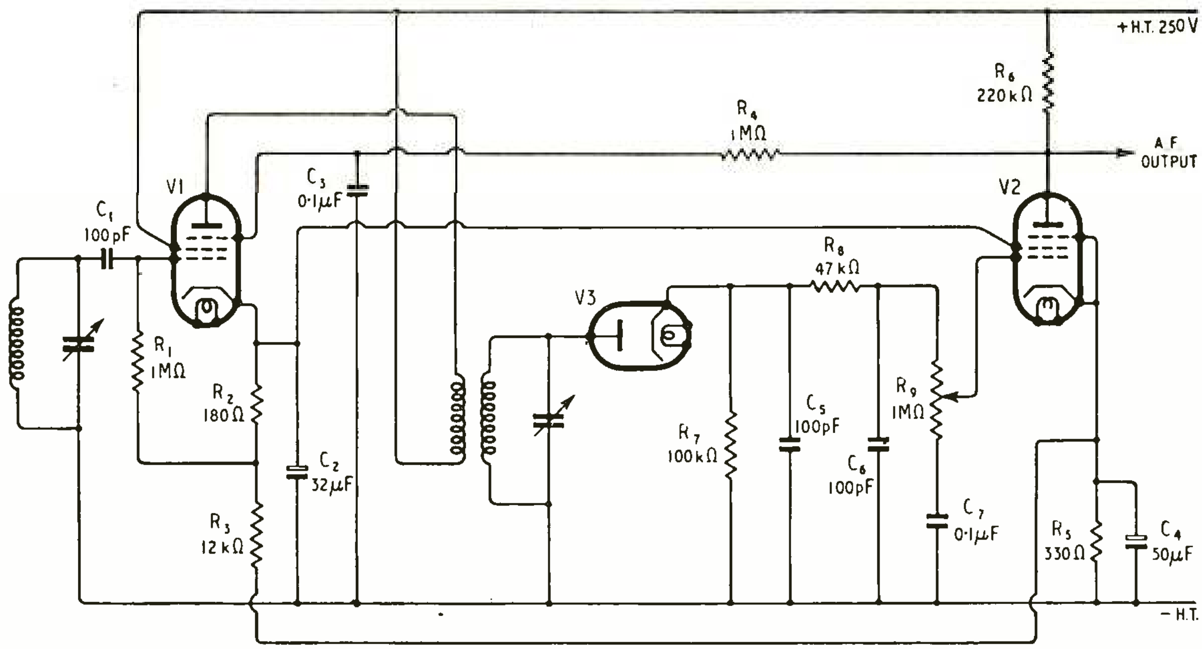


Fig. 3. One version of the amplified a.g.c. circuit in which an a.f. amplifier also functions as d.c. amplifier.

The screen of the audio amplifier must be decoupled by a large capacitor to minimize hum in the output and a 32-μF, 200-V component C_2 is used for this purpose. Moreover, to prevent a.f. signals being impressed on the suppressor grid of V1 an a.f. filter R_4C_3 is included in the suppressor feed. The detector circuit is unusual in that the d.c. blocking capacitor C_7 is included at the earthy end of the manual audio volume control R_9; by means of this circuit arrangement the full d.c. output of the diode is always applied to the grid of V2 irrespective of the setting of the volume control. R_5, the cathode resistor of V2, carries the cathode currents of V1 and V2 and is somewhat larger than might be expected; the larger value is necessary to offset the positive bias of approximately 1 volt which is given by the diode even in the absence of r.f. signals. The value of R_5 or R_6 is critical because between them they determine the quiescent anode potential of V2 and thus the suppressor potential of V1; one of these, preferably R_6, should be capable of adjustment and should be set so that the suppressor-cathode potential of V1 is zero with no signal input.

For successful results V1 must have a suitable, i.e., relatively short, suppressor grid base and the EF50 has been found satisfactory, the cut-off voltage being approximately – 50 volts. The SP41 and SP61 are unsuitable because the suppressor base is longer and the anode potential of V2 cannot, without extensive rearrangement of quiescent circuit potentials, be made to change sufficiently to give adequate control. The anode potential of V2 will not normally go less than about + 20 volts with respect to earth and a suppressor bias of – 80 volts is insufficient to cut off an SP41. Valves with specially short suppressor grid bases such as the 6F32 would appear to be suitable for use in this type of a.g.c. circuit (possibly without d.c. amplification) but have not been tried because they are of lower slope than the EF50.

Valve manufacturers generally stipulate an upper safe limit to the cathode-heater d.c. potential and this is sometimes as low as 50 volts. It is therefore advantageous to keep this potential as low as possible and in the circuit of Fig. 3 the cathode-heater potential can be reduced by connecting the heater winding of the mains transformer to a tapping point on the cathode load of V1. If the winding is centre-tapped it is not usually necessary to decouple the junction point to earth, but if there is no centre tap decoupling may be necessary; a 4-μF capacitor is generally adequate. An advantage of biasing the heater supply is that it frequently brings about a welcome reduction of hum in the a.f. amplifier.

In an alternative form of the circuit V2 may be an r.f. amplifier, in which case R_6 may be a decoupling resistor. Alternatively V2 may be an anode-bend detector and this brings about a considerable circuit simplification; it is hoped in a subsequent article to describe a sensitive t.r.f. receiver using such a circuit. V2 cannot be a leaky-grid detector because the anode potential of such a valve goes positive when an r.f. signal is applied to the grid.

The chief disadvantage of this circuit is that the signal-handling capacity of V1 is limited, and not increased by a.g.c. bias as in conventional a.g.c. circuits. A straight r.f. pentode overloads for inputs exceeding approximately 1 volt peak, but this disadvantage may be overcome by careful design. For example if V1 is a first r.f. amplifier, overloading on strong local stations may be prevented by the use of wavetraps. If some gain can be sacrificed, overloading in general can be avoided by the use of current feedback in the cathode circuits of controlled valves. Current feedback may be applied in Fig. 3 by tapping C_2 and the screen of V2 down the cathode chain of V1.

The a.g.c. provided by this circuit has an inherent delay; this is due to the shape of the $I_a - V_{g3}$ characteristic (Fig. 1) which is substantially level for the first few volts of negative bias applied to the suppressor grid. Thus there is no reduction in controlled-valve gain until its suppressor bias has fallen to the beginning of the knee of the characteristic. Because of the amplification of the a.g.c. bias this delay is not very great and can be eliminated if not required by arranging for the quiescent anode potential of the d.c. amplifier to be lower than the cathode potential of the controlled valve by a suitable amount. If a substantial delay is required it is probably best obtained from a biased diode in the normal manner, the diode output being d.c. amplified as before.

If the quiescent anode potential of the d.c. amplifier is made considerably less than that of the controlled-valve cathode, the maximum sensitivity of the receiver is limited. Such an adjustment is useful if the receiver is used only for reception of strong signal; the adjustment can be made in the circuit of Fig. 3 by suitable choice of the value of R_5 or R_6.

This circuit lends itself readily to the operation of an S-meter, which can be included in the screen circuit of the controlled valve. On weak signals the screen current is low and on strong signals it is high; thus the meter is forward-reading.

This circuit arrangement and other similar ones using the same principles are the subject of a patent application.

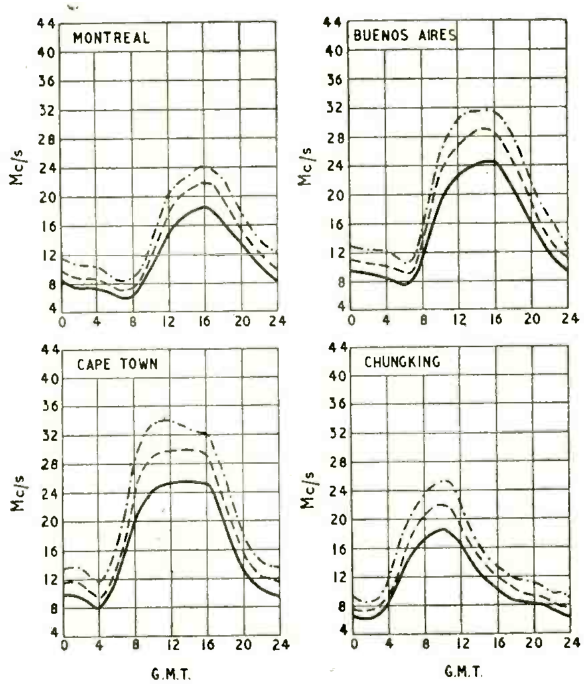
SHORT-WAVE CONDITIONS

August in Retrospect : Forecast for October

By T. W. BENNINGTON *

DURING August the average daytime maximum usable frequencies for these latitudes remained at about the same value as during July, instead of increasing as had been expected. The failure to increase was, no doubt, occasioned by the large amount of ionospheric storminess which occurred. The nightime m.u.f. decreased considerably, in accordance with expectations.

Daytime working frequencies were relatively low. Frequencies above 17 Mc/s were seldom well received over east/west paths, and over north/south paths 24 Mc/s was about the highest frequency for regular reception. Occasionally, of course, frequencies somewhat higher than these were workable over certain circuits. At night 11 Mc/s was usually workable till midnight or somewhat later, but frequencies as low as 7 Mc/s were necessary in the early morning hours.



——— FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE ON ALL UNDISTURBED DAYS
– – – – – PREDICTED AVERAGE MAXIMUM USABLE FREQUENCY
–·–·–·– FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE FOR 25% OF THE TOTAL TIME

Sporadic E was less prevalent than during the previous month, though on a few days 28 Mc/s communication to European countries was noted as taking place by way of this medium.

Sunspot activity was, on the average, considerably lower than during the previous month.

August was a very disturbed month, and conditions remained subnormal for some very long periods. The ionospheric storms giving rise to these conditions occurred during the periods 1st-3rd, 13-15th, 17-18th and 20-28th. Only one Dellinger fadeout was reported and that was of minor intensity.

Forecast.—During October there should be a continued increase in the daytime m.u.f. for these latitudes, with a continued decrease in the nightime m.u.f.

Daytime working frequencies should, therefore, be relatively high, though on east/west circuits 17 Mc/s should remain about the highest regularly used frequency, with those up to about 23 Mc/s sometimes usable. Over north/south circuits frequencies up to about 26 Mc/s should be regularly usable during the daytime, and those up to about 33 Mc/s occasionally so. The medium-high frequencies will, however, have to be used for a larger proportion of the daily time, due to the decrease in the hours of daylight. At night 9 Mc/s will generally be of use till about midnight, but after that time 7 Mc/s, or perhaps 6 Mc/s, will be about the highest regularly usable frequency.

Sporadic E should decrease sharply in the frequency of its occurence, and little communication on high frequencies is likely by way of this medium. It is unlikely that the E or F_1 layers will control transmission at any distance and medium distance communication will therefore be by way of the F_2 layer, being possible on slightly higher frequencies by day and on lower ones by night, than during September.

The curves indicate the highest frequencies likely to be usable over four long-distance circuits from this country during the month.

* Engineering Division, B.B.C.