amount during the three periods of acceleration, while Peter ages regularly by 30 years during the complete process: hence, when they meet again, Peter's age is 30 years and Paul's three days. On the other hand, if Peter is the traveller he ages by 1.5 days during each of his outward and return journeys, and by almost 30 years during the change from recession to approach with respect to Paul, while the stationary Paul ages by 15 years during Peter's outward journey, changes during Peter's reversal to a state nearly 15 years before birth, and then ages by 15 years during Peter's return, somehow getting born shortly before Peter arrives. Consequently, when they meet, Peter's age is 30 years and Paul's three days – exactly as in the former case.

We can hardly suppose that Einstein, Born and the others believed that these processes were both actual occurrances, the one entitled to claim reality depending on our preference in choosing to whom to assign the motion, nor did they. What they supposed was that the only *observable* events in the whole process were the separation of Peter and Paul at the beginning and their reunion at the end. Everything that happened in between was regarded as being beyond possibility of observation and therefore demanding compatibility only with theory, not with experience, with which it had nothing to do. This is obviously so important that it is necessary to confirm it by quoting Einstein's own words (in translation), all that needs explanation being that the clock U_1 is Peter and U_2 Paul and that "the right and left hand columns" give the descriptions of the process, as I have described them, when Peter and Paul, respectively, are regarded as moving. Einstein writes[3]:

> "You must bear in mind that exactly the same process is described in the right and in the left hand columns, but the description on the left refers to the coordinate system K while that on the right refers to K'. According to both descriptions, at the end of the process the clock U_2 is retarded by a definite amount compared with U_1. With reference to K' this is explained as follows: it is true that during the stages 2 and 4, the clock U_1, moving with velocity *v*, works more slowly than U_2, which is at rest. But this retardation is over-compensated by the quicker working of U_1 during stage 3. For, according to the general theory of relativity, the clock works the faster the higher the gravitational potential at the place where it is situated, and during stage 3 U_1 is indeed situated in a region of higher gravitational potential than U_2. Calculation shows that the consequent advancement amounts to exactly twice as much as the retardation during stages 2 and 4. This completely clears up the paradox."

What Einstein means here by "the same process" is, of course, everything that is *observable*, while "the description", which differs in the two cases, is wholly a mental construction. The first is unique, for it must be the one thing that would actually occur; the last owes allegiance only to theory, not observation, and can vary within the limits allowed by the theory.

But it is clear, beyond possibility of question, that Einstein's "descriptions" relate to what is observable, and cannot therefore both be permissible; and furthermore, as the credentials of both are exactly the same, it is impossible to decide which must be rejected. Paul could be accompanied by a nurse, of such an age as to become 30 years younger without losing her power of intelligent observation, and she would report on return whether it was a baby or a teenage boy who arrived at the planet, and whether or not a baby was born during the return journey, even if she were unable to confirm the antenatal age of the being whom the planet left. The question I asked Dr Wilkie was, in effect, whether what the nurse would observe would admit of both of Einstein's "descriptions", or whether a theory that required it to do so must be abandoned. I am not surprised at his reluctance to commit himself to a choice; nevertheless, it is imperative that scientists shall make a choice if the ethical demands of science are not to be jettisoned.

"... Mathematical consistency, though a necessary condition, is not a sufficient one for the truth of a physical theory."

What is the net result of all this? As I have said, it throws no light at all on what would happen if the experiment were made, for it is an analysis, not of a physical process that has never occurred, but of the requirements of a theory that purports to accord with physical processes, and I think it shows beyond doubt that the special relativity theory at least must be wrong. If the motion can be ascribed equally rightly to either twin, it cannot make them age at different rates; if it makes them age at different rates, there must be an absolute standard of rest to provide a criterion for distinguishing the faster from the slower developer. The special relativity theory requires different rates of ageing to result from motion which belongs no more to one twin than to the other: that is impossible.

It is impossible to exaggerate the importance of this result, for this theory is, by common consent, "taken for granted" in Max Born's words, in all modern atomic research, and it determines the course of practically all current developments in physical science, theoretical and experimental, whether concerned with the laboratory or with the universe. To continue to use the theory without discrimination, therefore, is not only to follow a false trail in the investigation of nature, but also to risk physical disaster on the unforeseeable scale, modern atomic experiments being what they are. It should therefore be a point of honour with those on whose authority atomic research is now being conducted to acknowledge at once the untenability of the theory, and to take without delay the necessary steps to discover where the theory falls.

That does not necessarily mean complete abandoning of its use, but it does demand the determination of the limits of its usefulness. It has already proved its effectiveness in many respects, and this has been mistaken by physicists for evidence of its truth. What the many successes of the Lorentz transformation equations have shown is that those equations are an effectual corrective of the imperfect classical electromagnetic equations within a limited range of experience. But it is now clear that the interpretation of those equations as constituting a basis for a new kinematics, displacing that of Galileo and Newton, which is the essence of the special relativity theory, leads inevitably to impossibilities and therefore cannot be true. Either there is an absolute standard of rest – call it the ether as with Maxwell, or the universe as with Mach, or absolute space as with Newton, or what you will — or else *all* motion, including that with the speed of light, is relative, as with Ritz. It remains to be determined, by a valid experimental determination of the true relation of the velocity of light to that of its source, which of these alternatives is the true one. In the meantime, the fiction of "space-time" as an objective element of nature, and the associated pseudo-concepts such as "time-dilation", that violate "saving common sense", should be discharged from physics and philosophy, and the fact realised that mathematical consistency, though a necessary condition, is not a sufficient one for the truth of a physical theory. Only thus can the scandal of more than half a century of confusion about the meaning of our own creations be ended.

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Audio gain controls

A survey of the methods used to achieve acceptable control of gain in audio amplifiers.

by **Peter Baxandall** B.Sc. (Eng.), F.I.E.E., F.I.E.R.E., M.A.E.S., F.B.K.S.T.S.

The design of gain controls is by no means as simple as it might appear. Peter Baxandall examines the difficulties in design and comments on many circuits which have appeared over the years, from very simple types in which compromises must be accepted, to those used in high-performance equipment.

An ideal audio amplifier with variable gain would have the following characteristics:

(i) noise output voltage = (source Johnson noise voltage) × (gain)

(ii) ability to deliver its full output voltage even at very low gain settings, which may be less than unity. The amplifier is therefore capable of handling very large input voltages at the lowest gain settings.

The simplest way to achieve (i) is that shown in Fig. 1(a), but this simple technique obviously fails lamentably with regard to (ii), for the maximum input voltage that can be handled without overloading is the same at all settings. The arrangement of Fig. 1(b), on the other hand, achieves (ii) perfectly, but fails with regard to (i).

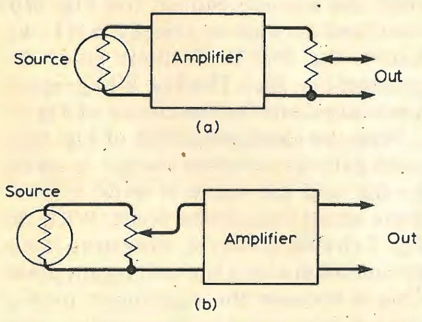

Fig. 1. Two small gain controls which do not fulfil both main requirements. Circuit (a) gives low noise, but will overload at low gain settings, while (b) introduces additional noise.

By using sufficiently subtle gain-control systems, it is possible to satisfy (i) and (ii) concurrently and almost perfectly, but the simple and widely used arrangement shown in Fig. 2 provides a compromise solution which is very satisfactory for many practical purposes.

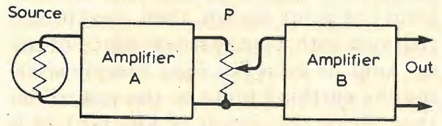

Fig. 2. This arrangement combines circuits of Fig. 1(a) and (b) to give compromise performance.

An ideal amplifier would give a variation of output noise voltage with gain setting as shown by the full-line graph in Fig. 3, whereas the Fig. 2 scheme gives a characteristic as depicted by the broken line. Below a certain setting of P, the noise level from amplifier A at the output of P becomes less than the noise level of amplifier B referred to its input, so that the noise of amplifier B becomes the dominant contribution, establishing the level of the broken-line "plateau".

Now there is obviously no practical advantage in achieving an output noise level which is a long way below audibility at very low gain settings, so that a characteristic of the broken-line type is normally perfectly satisfactory, provided the level of the horizontal plateau is low enough. For a given overall maximum gain requirement, the product of the voltage gains of amplifiers A and B in Fig. 2 is fixed, but there is a choice with regard to the apportionment of this gain between the two amplifiers. The higher the gain of A is made, the lower is the position of the Fig. 3 plateau, but there is the disadvantage that the maximum signal input that can be handled without overloading amplifier A is reduced.

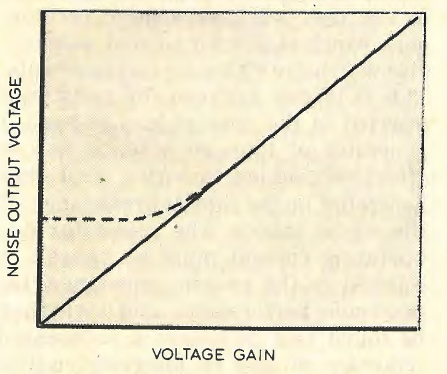

Fig. 3. Dotted line shows gain variation given by circuit of Fig. 2, where residual noise from amplifier B is predominant at low gain settings.

In domestic audio control units, the Fig. 2 arrangement is usually used. A suitable choice for the gain of amplifier B is normally such that full output level is delivered to the following power amplifier for an output level from the pot. slider of about 100mV r.m.s. If the wideband noise of amplifier B, with P set to zero, is equivalent to a noise input voltage to B of $0.5\mu V$ rms, which is fairly readily achievable, the zero-volume-setting noise output from B will then be 106dB below the full signal output level. (It may be added, however, that if the gain of B is made high enough to cope with the least sensitive of power amplifiers, which may require an input level of several volts, then the signal level at the pot. slider for full power output when used with a very-high-sensitivity power amplifier, will be much less than 100mV rms, and a figure much less than the 106dB mentioned above will then apply. Thus, for versatile use, it is desirable to provide a preset gain adjustment within amplifier B, or in the form of a simple passive attenuator after this amplifier.)

A closer approximation to concurrently satisfying conditions (i) and (ii) at the beginning of the article may be obtained, on the same principle as in Fig. 2, by employing three amplifiers with ganged gain-control pots. between them, but in general it is much better, instead, to employ schemes in which variable negative feedback provides much of the gain variation.

Variable-feedback gain control offers advantages both with regard to achievable performance (noise and signal-handling) and often with regard to economy of circuit design. Variable feedback alone cannot normally reduce the gain to zero; for 100% voltage feedback reduces the gain to unity rather than zero. Thus, it is usual to combine feedback variation with passive gain control, sometimes using a ganged pot. and sometimes using the parts of the track either side of the slider, in an ordinary single pot., to perform these functions. There are many possible schemes, of which some have been known for thirty years or more.

One of the simplest schemes is that shown in Fig. 4. The pot. resistance can be made quite low, e.g. 1kΩ, since it is driven by the op. amp., not the signal source. This results in a good noise performance at all settings. Disadvantages of the circuit are:

(a) the minimum gain is unity, not zero, and

(b) a floating signal source is required.

Disadvantage (b) is of little consequence when an input transformer is used, and (a) may be overcome by taking the signal output from the pot. slider. The latter change, of course, sacrifices the virtue of very low output impedance possessed by the Fig. 4 version.

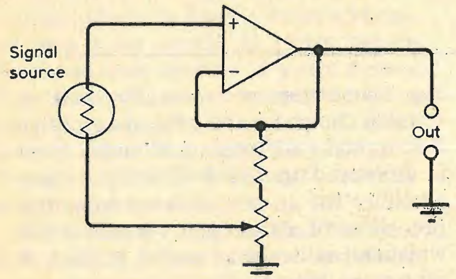

Fig. 4. Simple feedback gain control.

In assessing the pros and cons of various circuits, it is very helpful to appreciate the relationships between the circuits in the most vivid possible way, rather than relying purely on formal analysis. Very often the differences between circuits are much smaller than they appear to be, involving merely the choice of earthing point and/or the way of drawing the circuit diagram, rather than differences of more fundamental significance. Sometimes, in redrawing circuits employing op. amps. to facilitate better understanding of them, it is helpful to replace the op. amp. symbols by ordinary single-transistor symbols — an unfamiliar-looking circuit may then suddenly be recognised as an old friend! At other times, replacing a detailed transistor circuit by the op. amp. equivalent may reveal its true nature in the best way.

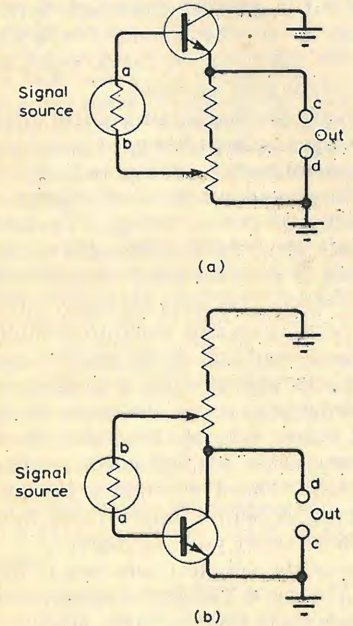

Fig. 5. Single transistor equivalent to Fig. 4, neglecting d.c. conditions. Rearrangement in (b) shows circuit to be easily recognizable.

On replacing the op. amp. in Fig. 4 by a transistor, the circuit of Fig. 5(a) is obtained. Though the collector would in practice be taken to a positive supply line, it is here shown as earthed, for in the present context we are concerned only with a.c. aspects and it is best to omit irrelevant details.

Shifting the earthing point to the emitter of the transistor, but making no other changes, leads to Fig. 5(b), which ios a simple common-emitter amplifying stage with adjustable feedback.

If the output in the Fig. 4 circuit is taken from the pot. slider instead of from the point shown, then the circuit, redrawn with a transistor in place of the op. amp., is as in Fig. 6(a). Merely shifting the earthing point to the pot. slider then yields the circuit of Fig. 6(b). It is now evident that moving the slider to the right has two separate effects — it increases the amount of resistance in the emitter lead, thereby increasing the amount of negative feedback, and it reduces the collector load resistance. Both these effects contribute to reducing the gain, which becomes zero with the slider fully to the right.

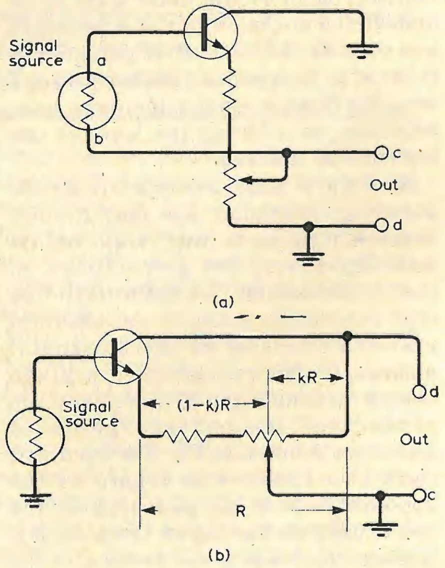

Fig. 6. Fig. 4. Circuit with output taken from pot slider and rearranged at (b) to show dual function — varying emitter resistance and varying feedback.

Employing just a single transistor, as in Fig. 6(b), will give a noise performance which is inferior to that achievable with more elaborate arrangements. This is largely because the resistance inserted in the emitter lead is itself a generator of Johnson noise, which is effectively added in series with that generated in the internal resistance of the signal source. The transistor d.c. operating current must be chosen in relation to the source impedance, for good noise performance, and it will then be found that to obtain a substantial reduction of gain by inserting emitter resistance, the amount of resistance needed will give considerable degradation of the noise figure.

The above noise difficulty may be solved by replacing the single transistor by a suitable pair or triple, having a much higher mutual conductance than the single transistor but whose input stage operates at a similarly low current. The increased mutual conductance and output current permit the resistance values associated with the gain-control pot. to be made much lower, with a correspondingly reduced effect on the noise performance at low gain settings. The well-known configurations for pairs and triples as used in audio class 'B' output stages may be adapted to the present application, but an interesting alternative is that shown in Fig. 7. Here the supply connexions to the op. amp. are used as the equivalent of the transistor collector in the Fig. 6(b) circuit — a way of using an op. amp. which perhaps deserves to be more widely borne in mind.

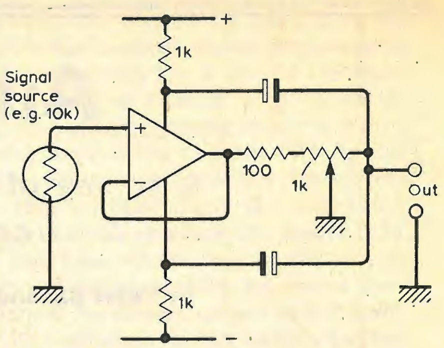

Fig. 7. Fig. 6(b) using an op. amp.

Assuming infinite mutual conductance, the voltage gain of the Fig. 6(b) idealized circuit is simply $k/(1-k)$. Expressing this in decibels gives the graph of Fig. 8(a). The Fig. 8(b) graph is a measured one for the circuit of Fig. 7.

With the idealized circuit of Fig. 6(b), unity gain occurs when the pot. is set for $k=0.5$, and the curve is quite symmetrical about this centre point. With the Fig. 7 circuit, however, the curve is not symmetrical about the unity-gain point. This is because the right-hand part of the pot. is shunted by the parallel value of the two 1kΩ resistors going to the supply lines.

Another very simple feedback gain-control circuit is shown in Fig. 9. With high forward gain in the op. amp. itself, this circuit gives a gain, between the input and output terminal pairs shown, accurately equal to $k/(1-k)$. (This formula, as for the Fig. 6(b) case, may be prefixed by a minus sign if it is desired to allow for the fact that phase inversion occurs.)

The Fig. 9 circuit, unlike those previously discussed, has the feature that the current in the gain-control resistance chain is supplied by the signal source. This makes it impossible to achieve a good noise figure over a wide range of gain adjustment, no matter how the resistance values are chosen in relation to the signal-source impedance. That this must be so can best be understood as follows. Negative feedback as such never has any effect on the signal-to-noise ratio, at a given frequency, of an amplifier circuit to which it is applied, though the resistors introduced for the purpose of providing the feedback may do so. Thus the output signal-to-noise ratio of the Fig. 9 circuit is the same as that of the circuit shown in Fig. 10. If R is made low, say equal to the internal resistance of the signal source, it will degrade the signal-to-noise ratio at the source terminals*, whereas if R is made much higher, a large amount of resistance is introduced into the op. amp. input circuit at intermediate slider settings, with correspondingly large Johnson noise and maybe noise from the op. amp. equivalent current-noise generator.

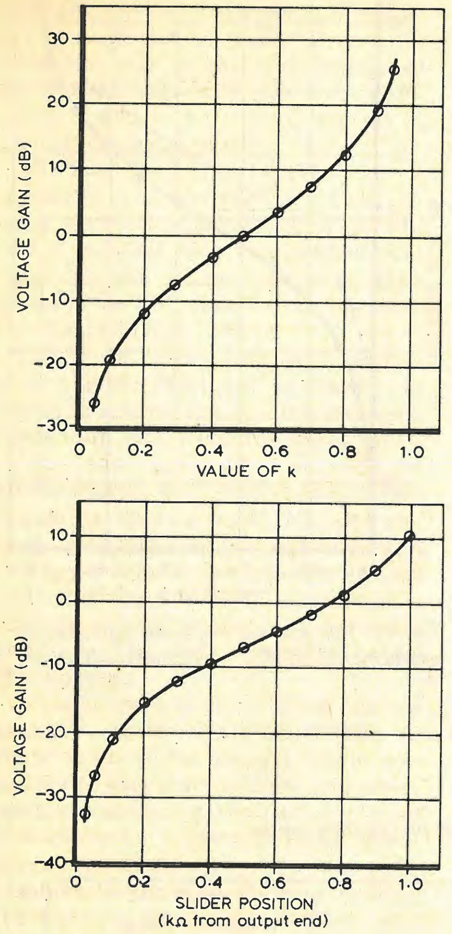

Fig. 8. Gain variation of Fig. 6(b) circuit is at (a). Measured performance of equivalent op. amp. circuit of Fig. 7 is shown at (b).

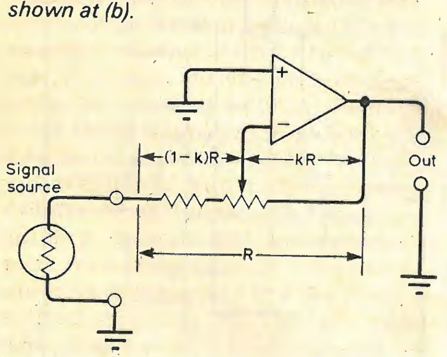

Fig. 9. Feedback gain-control circuit, which has disadvantage of source-fed resistor chain, giving poor noise figure over wide range.

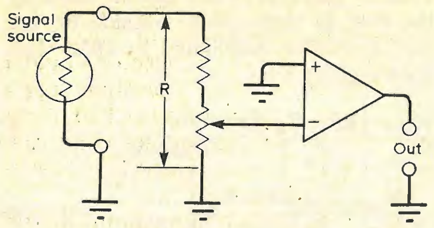

Fig. 10. Circuit of Fig. 9 gives same noise performance as circuit shown here.

Comparing Fig. 10 with Fig. 1(b) might suggest that the Fig. 9 circuit is no better than that of Fig. 1(b) as regards noise performance. This is not so, however, for to effect a given number of decibels reduction of gain below maximum, the slider in Fig. 9 has to be moved a smaller fraction of the way from the signal-source end of R than is necessary for the same gain reduction in the Fig. 1(b) circuit. The noise performance of Fig. 9 is better than that of Fig. 1(b), but is nevertheless not very good.

Another feature of the Fig. 9 circuit which makes it undesirable for some applications is that the loading of the signal source varies with the pot. setting. If the signal source has a complex internal impedance, the overall frequency response will vary with the gain setting.

This undesirable characteristic of the Fig. 9 circuit may, to a large extent, be overcome by inserting an emitter-follower (or op. amp. follower) between the signal-source and the left-hand end of the resistance chain. With a 50kΩ signal-source, for example, R could be made about 5kΩ, giving reduced Johnson noise from R but nevertheless subjecting the signal-source to negligible loading.

As already mentioned, the Fig. 9 circuit as it stands produces the gain-control characteristic shown in Fig. 8(a), which is symmetrical about the unity-gain point. Over a range of about 30dB, and using an ordinary linear pot., the scale shape obtained approximates fairly reasonably to the desirable one having uniformly-spaced decibel divisions, though for many applications a gain of more than unity would be preferred at the centre of this control range. The modification shown in Fig. 11 provides an increased gain at the point of inflexion of the control characteristic, but has the weakness that the gain cannot be reduced right down to zero. Provided R_a and R_b are made much lower in value than the pot. resistance, however, the minimum gain may be made sufficiently low for many purposes.

If a stud type pot. is used, and assuming there is complete freedom in the choice of its law and total resistance value, the Fig. 11 modificiation gives no advantage, the required performance being obtainable with better economy of components by adopting the Fig. 9 arrangement.

The circuit of Fig. 12 possesses a combination of several good features. It employs only one op. amp., has a high input impedance, the feedback network can be of low resistance for good noise performance, and the values of R_a and R_b can be chosen, in relation to R, to make the point of inflexion in the control characteristic occur at a gain of much greater than unity, as sometimes desired.

Analysis shows that the gain of the Fig. 12 circuit is given by:

$$\frac{V_{out}}{V_{in}}=\frac{R+\dfrac{R_b}{1-k}}{R+\dfrac{R_a}{k}} \qquad 1.$$

$$\text{or}\ \frac{V_{out}}{V_{in}}=\frac{k}{1-k}\times\ \frac{R(1-k)+R_b}{R_k+R_a} \qquad 2.$$

Thus, if R_a and R_b are each much greater than R, the gain is approximately proportional simply to $k/(1-k)$, and is approximately equal to R_b/R_a when $k=0.5$. Thus the control characteristic is fairly closely as in Fig. 8(a) but shifted upwards. For lower values of R_a and/or R_b the characteristic is of modified form, covering a smaller number of decibels with reasonable linearity.

The curve shown in Fig. 13 is the result of a measurement using the Fig. 12 circuit with the following values:

$$R=1\text{k}\Omega\ R_a=330\Omega\ R_b=3.3\text{k}\Omega$$

Comparison of this curve with Fig. 8(a) shows that it gives a poorer

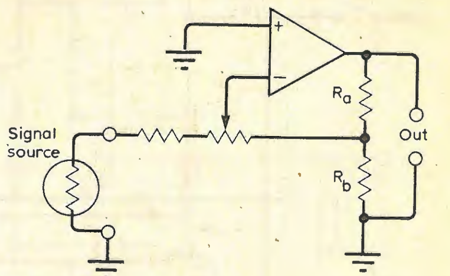

Fig. 11. Variation of Fig. 9, giving increased gain at halfway position of slider.

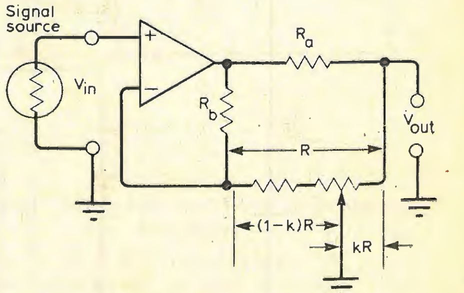

Fig. 12. Circuit featuring only one amplifier, high input impedance, low-resistance feedback chain for low noise and flexibility in choice of inflexion point.

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*When a resistive source of internal resistance R is shunted by a load resistance equal to R, the signal voltage is halved, but the Johnson noise voltage is reduced by a factor of only 2. The signal-to-Johnson-noise ratio is therefore worsened by 3dB.

approximation to the ideal linear shape for values of k above about 0.2. (The ideal curve would not, of course, remain linear down to $k=0$, for this would make it impossible to fade a programme down to zero volume. For most audio purposes, the ideal characteristic would cover about 40dB linearly, curving down to "minus infinity dB" below about $k=0.2$.)

Another circuit combining feedback and passive gain variation by means of a single linear pot. is shown in Fig. 14.

This, in essence, is the circuit used by the BBC in their OBA9 outside broadcast amplifier, published in 1952. The gain is given by:–

$$\frac{V_{out}}{V_{in}}=\frac{kR+R_a}{R_a}\times\frac{R_b}{(1-k)R+R_b} \qquad 3.$$

$$\text{or } \frac{V_{out}}{V_{in}}=\frac{1+kR/R_a}{1+(1-k)R/R_b} \qquad 4.$$

The Fig. 14 circuit cannot give zero voltage gain, the gain with $k=0$ and 100% negative feedback being $R_b/(R+R_b)$. Though not an ideal feature, the minimum gain in the BBC design is nearly 90dB below the maximum gain, and is stated to be "effectively nil in normal conditions of use"[2].

The full-line curve in Fig. 15 is a calculated result for the Fig. 14 circuit, using the values $R=100\text{k}\Omega$, $R_a=330\Omega$ and $R_b=3.3\text{k}\Omega$. For the broken-line curve, R_b was changed to 10kΩ. (The values in the BBC design were $R=1\text{M}\Omega$, $R_a=390\Omega$ and $R_b=100\text{k}\Omega$.)

Figure 15 shows that with an ordinary, linear 100kΩ pot. in the Fig. 14 circuit, a control law not departing by more than 2dB from the ideal linear decibel scaling is obtained over an approximately 40dB range. In the BBC design[2], a stud type of 1MΩpot. was used, giving 38 steps of 2dB each and two larger steps at the low-gain end. Of course, if the luxury of stud pots. is allowed, any of the circuits here discussed may be given whatever control law is desired.

Though there is much to be said on grounds of economy, especially in stereo systems, for using a single pot. section to vary the feedback and effect passive attenuation, the use of ganged stud type pots. to perform these operations separately gives the designer greater freedom of choice in optimizing the design in all its aspects. This technique was used in the BBC OBA8 outside broadcast amplifier, designed well over forty years ago[1]. Starting at the maximum-gain setting, anticlockwise rotation of the knob first simply applied increasing negative feedback to the first stage, by raising the effective value of the feedback resistance in the cathode circuit. When this purely local feedback had been increased sufficiently to give a gain reduction of 16dB, further rotation of the knob maintained this first-stage feedback constant but proceeded to insert increasing passive attenuation between the first stage and the second (output) stage. In this way the two-valve amplifier was made capable of delivering full output level to line, at low distortion (abour 1%) for peak microphone input levels extending over a range of 56dB. (It is evident that the designers of this amplifier and the associated units gave high priority to keeping the number of valves used down to the absolute minimum necessary number. This is understandable enough, bearing in mind that the AC/SP3 television pentodes used were physically large and consumed four watts of heater power each. Now that high-gain devices are very small and cheap, and consume relatively tiny amounts of power, the designers of today are justified in adopting a very different outlook, often exploiting the plentifulness of gain to eliminate, or reduce the size of, transformers and also to achieve lower distortion levels in equipment of very much smaller size. Now that it has become fairly easy and cheap to obtain very low distortion levels, there is little argument for doing otherwise, whereas when the OBA8 was designed, lower distortion would have meant more valves, higher power consumption and shorter operating time on standby batteries. The designers were therefore justified in making the distortion just comfortably low enough, but no less, though they were doubtless quite capable at that time of achieving much lower distortion levels had this been thought desirable. In most circumstances of use, it is doubtful whether the subjective quality of the OBA8 could be distinguished from that of the best modern equipment. The weakest feature of the design is that the secondary of the input transformer,

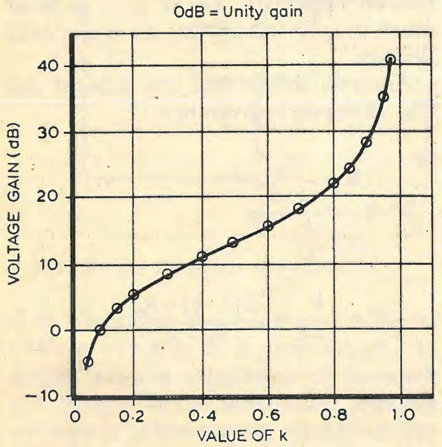


Fig. 13. Curve of circuit in Fig. 12.

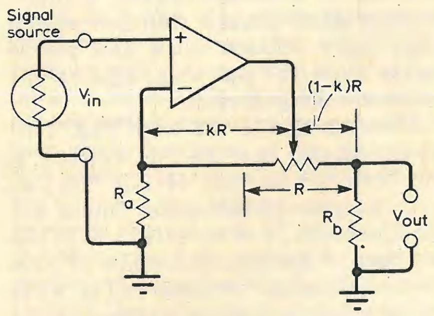


Fig. 14. Circuit providing feedback and passive control in one pot.

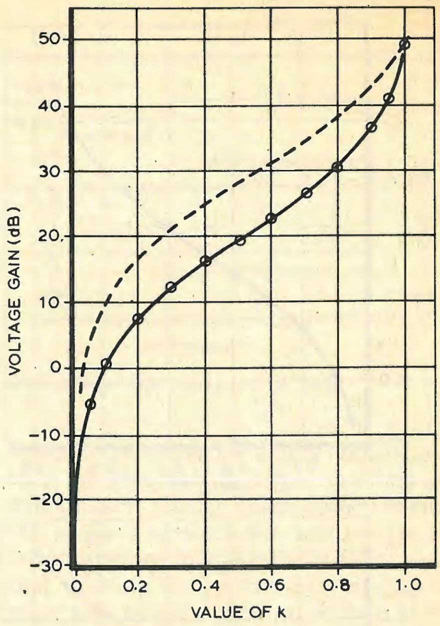


Fig. 15. Full-line curve shows calculated performance of Fig. 14 for two values of R_b.

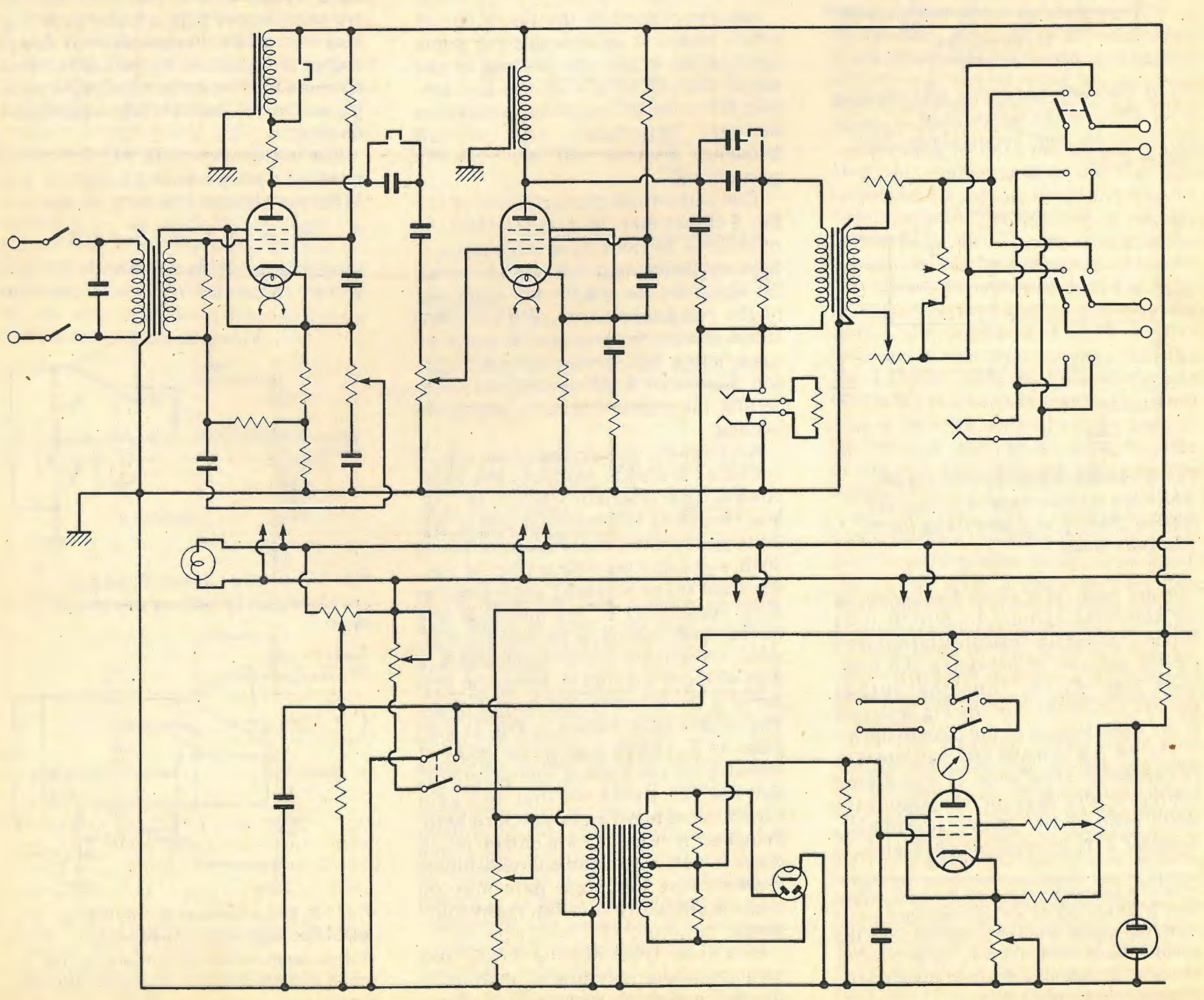


Fig. 16. BBC OBA8 circuit of 1939, with peak programme meter.

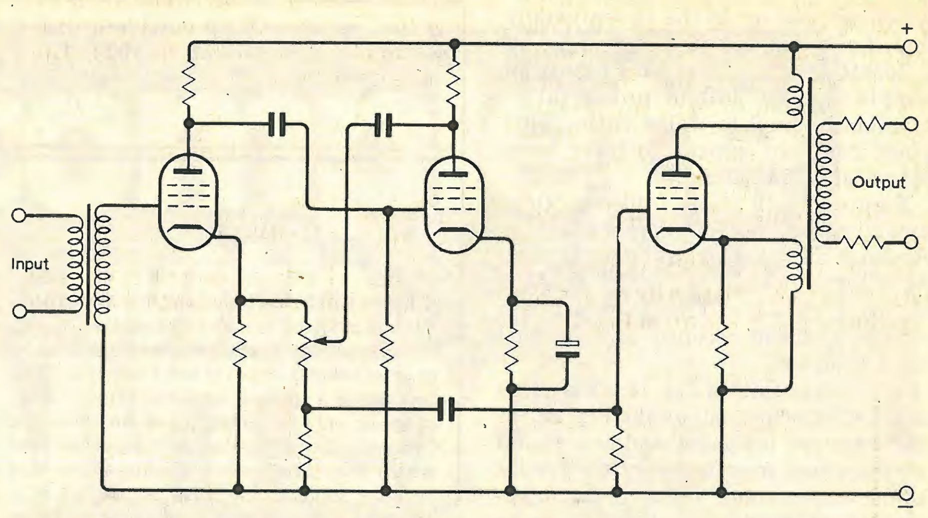


Fig. 17, BBC OBA9 circuit, designed in 1952.

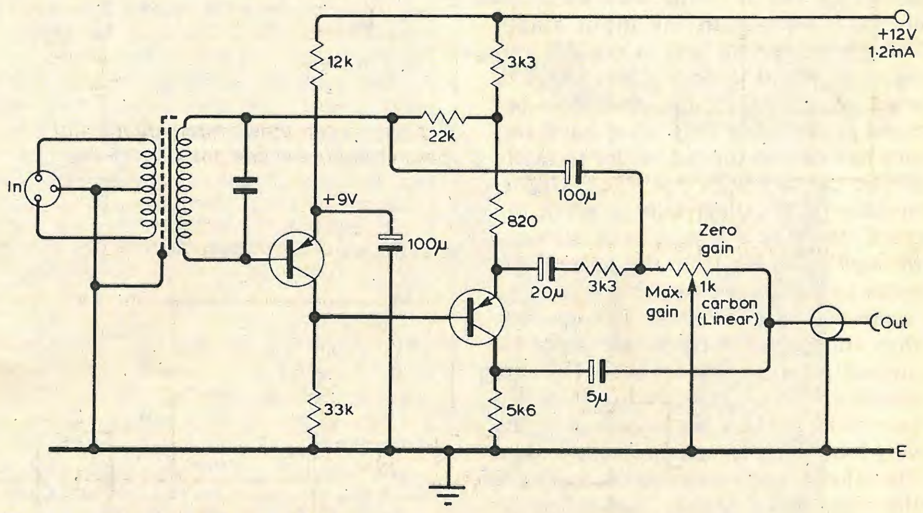


Fig. 18. Author's design of 1961.

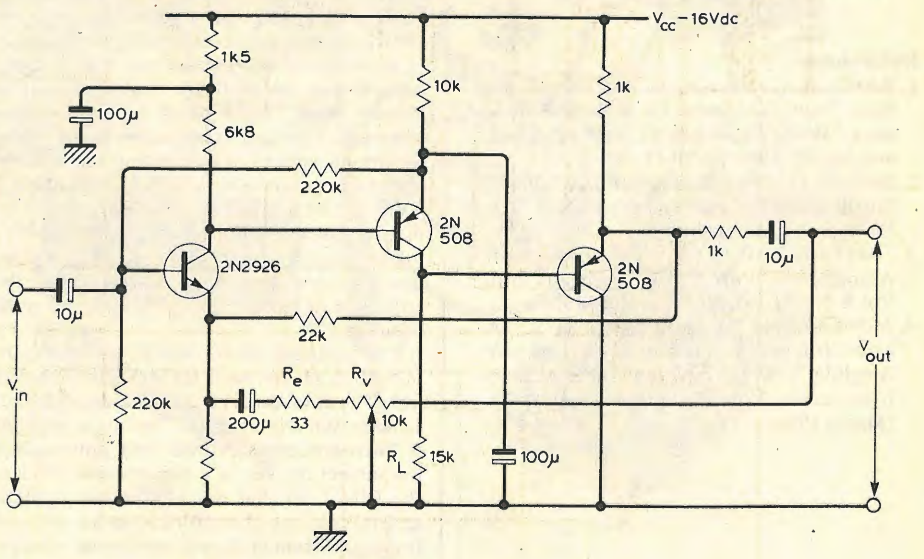


Fig. 19. Circuit by McWhorter of 1966.

which stepped up to the exceptionally high impedance of 300kΩ, is shunted by a 300kΩ resistor, thus sacrificing, in simple theory, 3dB of potentially-available signal-to-noise ratio. This point does not appear to have been appreciated at the time.)

Figures 16-19 show four practical amplifiers which use a combination of feedback and passive gain control. The McWhorter design[4] of Fig. 19 employs the basic circuit of Fig. 12, which has also re-surfaced recently in a Philips tape recorder[8].

My own circuit[3] of Fig. 18 is the same in broad principle, but unlike Fig. 12 has the negative feedback and the signal output taken from different electrodes of the output stage. This permits injection of the feedback voltage in series with the transformer secondary, thus obviating the introduction of local emitter feedback in the input stage. Though this circuit was in regular and very successful use for some years, a weak point in its design ultimately became evident, but only after hard service had caused the pot. slider to make erratic contact with the track. Unfortunately, if the slider fails to earth the track, there is a signal path straight through the track from the output collector to the input base. This is positive feedback and is of greater magnitude than the negative feedback from the output emitter. Violent oscillation therefore occurs during moments of poor slider contact, with accompanying very loud noises from the loudspeaker! The other circuits described do not have this weakness – a point worth bearing in mind.

To be continued

References

1. Barrett, A. E., Mayo, C. G. and Ellis, H. D. McD. "New Equipment for Outside Broadcasts." *World Radio* July 21, 1939, pp. 12-13 and July 28, 1939, pp. 10-11.
2. Berry, S. D., "New Equipment for Outside Broadcasts", *The BBC Quarterly*, Vol. VII, No. 2, pp. 120-128. (Summer 1952).
3. Baxandall, P. J., "Low Distortion Amplifiers – Part 2", *B.S.R.A. Journal*, Vol. 6, No. 11, pp. 246-256. (Nov. 1961).
4. McWhorter, M. M. and Warner, G. S., "A Low-Noise Transistor Microphone Amplifier", *IEEE Trans. on Audio & Electroacoustics*, Vol. AU-14, No. 1, pp. 27-31. (March 1966). ☐

BOOKS

A Users Guide to Copyright, by Michael F. Flint, is intended to make clearer the subject of copyright "to enable people whose jobs — or even hobbies — cover any copyright field, to acquire a general understanding". It is, however, only a reference book, and does not cover all the more complex legal aspects which may arise when dealing with this intricate subject. The book is laid out in a manner which will enable its reader to obtain the relevant piece of information quickly, and each chapter is sub-divided into well defined sections, each with a reference number and a bold sub-heading. Part 1, the first 14 chapters, is a general explanation of the copyright law, whilst the second part gives a more specified description of copyright in practice, with chapters directed at publishers and printers, advertising agencies, the music industry etc. The book is published by Butterworth Law Publishers Ltd, 88 Kingsway, London WC2B 6AB, and its price is £8.50 in limpback form.

Microcomputers are responsible for a great number of paperbacks, mainly from the USA, and the pace of publication does not appear to be slackening. Three such books have reached this office recently, among others too numerous to mention, each slanted in a different way.

The first is by a British author, Robin Bradbeer, and is entitled **The Personal Computer Book,** published by Input Two-Nine at £5.25 and distributed by MCB Publication, 198/200 Keighley Road, Bradford, West Yorks. BD9 4JQ. This one assumes no knowledge of computers — not even enough to know what computers will do — and, accordingly, the first two chapters are extremely basic. The rest of the book is an attractively written explanation of the more important aspects of computing techniques and of computers, a very useful feature being a survey of equipment currently on the market. Several appendices provide information which is quite difficult to find elsewhere in one place, such as bus standards, addresses of clubs, manufacturers and publications.

The second book, by E. A. Parr, is published by Bernard Babani (Publishing) Ltd., The Grampions, Shepherds Bush Road, London, W6 7NF at £1.75. This one is entitled **A Microprocessor Primer,** and approaches the subject by way of a hypothetical device, the DIM-1, so that the author can explain general features of microprocessors without being constrained by any particular design. Having gone through this process, he then sets out to study the Z-80. This is a small book (75 pages) but within its scope achieves its purpose.

Thirdly, there is **Introduction to Microcomputers for the Ham Shack,** by Harry L. Helms, Jr., published by Howard Sams and distributed by Prentice-Hall International, 66 Wood Lane End, Hemel Hempstead, Herts HPZ 4RG at £3.20. Also a small book, this is concerned with the application of micros to amateur radio. Three chapters are allocated to the basics of micro operation and programming, after which two chapters describe present and future operations using micros to send and receive Morse, to convert slow-scan tv to fast-scan for ordinary viewing, to store frequencies, in digital modulation, and in several other roles.

Early Radio Wave Detectors, by V. J. Phillips, gives a comprehensive account of various radio wave detectors used before the advent of the crystal and thermionic valve. Among the types described are spark-gap, electrolytic, magnetic, thin-film and capiliary detectors, as well as tickers, tone wheels, heterodynes and coherers, the type of detector which makes use of "a phenomenon which occurs in a poor electrical contact, the sort of contact which the engineers of today would call a 'dry joint' ".

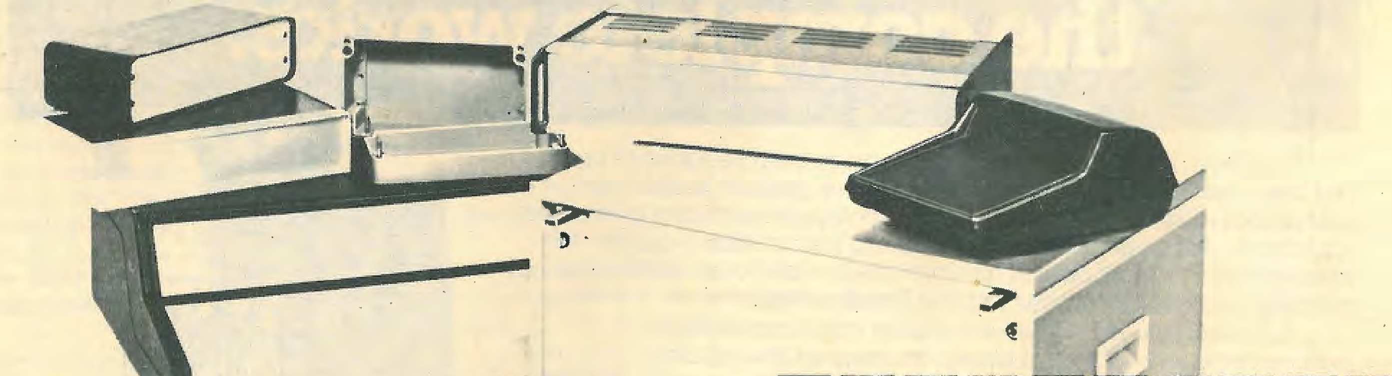
Among the items described under the heading "Miscellaneous detectors," are the 'physiological' receiver, which made use of the electrical sensitivity of a frog's leg to displace a pointer on the smoked surface of a rotating drum, and the use of a human brain as a coherer, the description of which is supplemented by a photograph for which an advisory note is given for the benefit of "readers of delicate sensibilities". Be forwarned, however, the note appears at the bottom of the page, and the photo at the top!

The last chapter, entitled "And so to the modern era," covers the early crystal and thermionic valve type detectors and how they were used — an appropriate finale to an interesting and well-illustrated book. The publishers are Peter Peregrinus Ltd, Marketing Dept, Station House, Hitchin, Hertfordshire SG5 1RJ, and the price of the book in hardback form is £16.

Digital Techniques and Systems, by D. G. Green, is intended as a first course book for students with a basic knowledge of electronics and telecommunication transmission techniques, but the combined coverage of basic techniques used in modern digital circuits, and elementary principles of data communication, laid out in a logical sequence, make it useful for anyone wishing to gain insight into this field.

Chapter 1 gives a concise introductory description of a few of the uses of modern digital applications to which he may put the knowedge that he is about to learn. The second and third chapters cover the operation of electronic gates of all kinds and the remainder of the book, which includes chapters on digital modulation, data-links and pulse code modulation, is devoted to the subject of data transmission over telephone lines.

Worked examples are included in the text, and each chapter concludes with exercises, some of the questions of which have been taken from past C and G examination papers. Multiple-choice questions are also provided at the end of the book, which is priced at £4.95 and published by Pitman Books Ltd, 39 Parker St, London WC2B 5PB.



Audio gain controls

2 – Obtaining equal gains in the two channels of a stereo pair

by **Peter Baxandall,** B.Sc. (Eng.), F.I.E.E., F.I.E.R.E., M.A.E.S.

Continuing his survey of gain control problems and solutions, Peter Baxandall discusses tracking volume controls in stereo amplifiers, concluding with a proposal for an unusual design of control.

Stereo gain control tracking

Connected with the problem of obtaining a satisfactory scale-shape for the volume-control law in stereo control units, is that of achieving an accurately equal gain in the two channels at all knob settings. Preferably, the channel gains, if adjusted to be equal at one volume control setting, by means of the balance control or otherwise, should remain within ±1dB of equality at all other settings of operational significance. This is quite likely not to be the case if cheap types of carbon-track, ganged log. pots. are used.

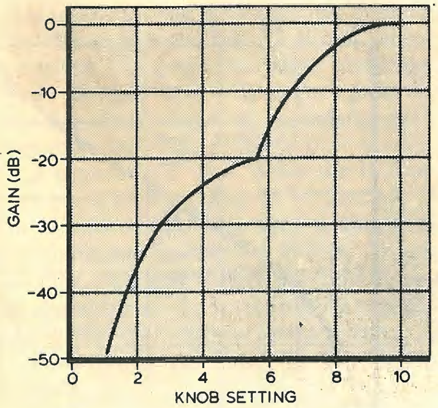


Fig. 20. Approximation to log. law obtained by changing resistivity of halves of carbon-track pot.

Figure 20 shows the measured gain-variation law on one channel of a very high quality, commercial control unit, having a simple, passive volume-control circuit, using the above type of pot. The very rough approximation to a logarithmic (linear-in-dB) law is obtained by making the two parts of the pot. element of different surface resistivities, the resistivity changing suddenly from one value to another at half-rotation of the knob. At the point of change, there is a severalfold change in slope, which is a quite undesirable feature. Though some quite cheap commercial pots. give a better approximation to a logarithmic law than that of Fig. 20, there is clearly much to be said for employing a type of gain control circuit which inherently gives a smooth and nearly logarithmic law without needing pots. with a non-linear resistance law. It ought to be easier to make ganged linear pots. with accurate matching between sections than to make ones with non-linear laws and equally good matching, though unfortunately, limited experience in measuring the departure from linearity of cheap so-called linear carbon-slider pots. has shown that undesirably large errors often occur.

One solution to the problem of obtaining a good scale shape and accurate tracking is, of course, to employ ganged, stud-type volume controls. These should give not more than 2dB per stud, at the most, and should have a click mechanism to make sure they are never left in an unsatisfactory half-way state between one stud and the next. Then, provided their internal resistors are accurate and stable, very accurate tracking will be obtained.

Careful measurements have been made of the resistance versus knob-position relationship for eight specimens of R.S. Components 10kΩ linear "slide tandem" pots, and Fig. 21 shows the results for three of these. It will be seen that:

(a) none of the specimens has a truly linear law;
(b) the departure from linearity, though of somewhat different nature for the three specimens, is nevertheless of fairly accurately the same shape for the two halves of each specimen, and this is the case also for the other five specimens;
(c) there are considerable differences between the absolute total resistance values of the specimens, and, in the case of specimen number 3 particularly, between the two resistance elements in one specimen.

For normal audio control-unit applications, minor departures from the nominal volume-control law are unimportant, provided they are equal for the two channels. Differences in the absolute resistance values for the two elements in a stereo pot. may or may not cause gain mistracking, dependent on the nature of the associated circuit.

Consider first the circuit of Fig. 22(a), which gives a range of gain well suited to most control-unit applications. (The circuits of Figs. 12 and 14 are better suited to microphone-amplifier applications, where the higher maximum gain given is advantageous.) It is necessary in practice to insert a resistor R_1 in series with the input end of the pot. to limit the maximum value of k obtainable to, say, 0.9 or 0.95, otherwise – see Fig. 8(a) – the characteristic becomes too steep at the high-gain end. Note that k is defined as

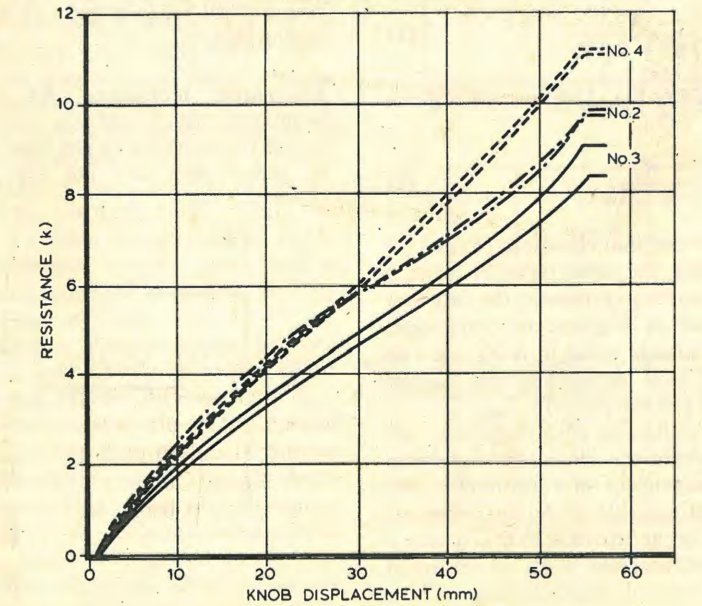


Fig. 21. Samples of characteristics of dual linear pots.

shown in Fig. 9, and is not the same as k' in Fig. 22. The reason for introducing k' is that it enables a more straightforward comparison to be made between the behaviour of the (a) and (b) circuits in Fig. 22 – k' is a measure purely of the knob position, whereas, as shown in Fig. 9, k involves also the value of the fixed series resistor.

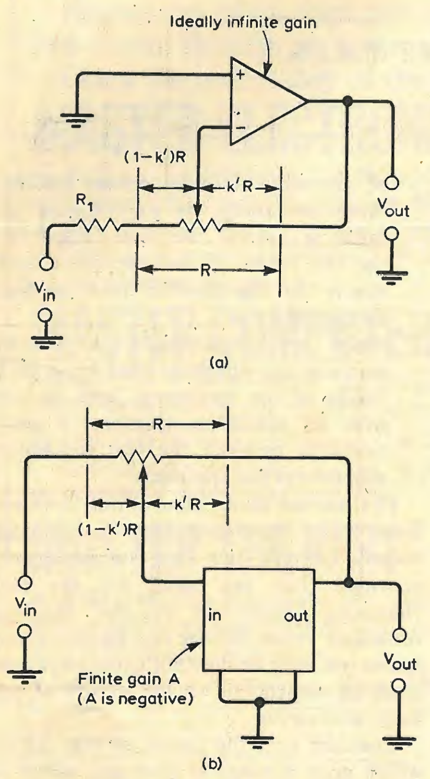


Fig. 22. In circuit (a) the total resistance of R compared with R₁ varies the control curve, whereas the circuit at (b) is independent of track resistance.

The gain of the Fig. 22(a) circuit is given by:–

$$\frac{V_{out}}{V_{in}} = -\frac{k'R}{(1-k')R+R_1}$$

$$= -\frac{k'}{1-k' + R_1/R} \qquad (5)$$

The gain of the Fig. 22(b) circuit is given by:–

$$\frac{V_{out}}{V_{in}} = -\frac{k'}{1-k' - 1/A} \qquad (6)$$

It will be seen that equations (5) and (6) are of exactly the same form, A being a negative number to represent the fact that the amplifier is a phase inverting one. Thus if A is made equal to R/R_1, the two circuits will have identical graphs relating overall gain to knob position.

Circuit (b) has an advantage over (a), however, in that the control characteristic is quite independent of variations in the absolute resistance R of the pot. element, whereas in (a) an increase in R requires a proportionate increase in R_1 to return to the same control characteristic. Thus, using a pair of circuits of the (b) type in a stereo system, differences in the element resistances in the two halves of the ganged pot., which, as already mentioned, are found to occur in practice, will not affect the accuracy of tracking between the channels, whereas in (a) an increasing discrepancy will occur as the gain setting is increased. It has been assumed that the amplifier input impedance in circuit (b) is very high, so that there is no significant loading on the pot. slider.

To carry out the Fig. 22(b) scheme in practice, an economical recipe is required for a phase-inverting amplifier of high input impedance and feedback-stabilized gain. The simple arrangement shown in Fig. 23(a) is not very good, for to avoid significant loading of the slider, the resistors R_a and R_b must be made very high in value, which then seriously degrades the noise performance. This problem may be satisfactorily solved by inserting a unity-gain follower between the slider and R_a, R_a and R_b now being made of very much lower values. This arrangement is shown in Fig. 23(b).

Amplifier A in Fig. 23(b) has to handle only quite small voltage excursions, even though V_{in} and/or V_{out} may sometimes reach levels of several volts. There is no need to use an op. amp. for A, better economy, with little degradation in performance, resulting if a simple emitter-follower is used. A satisfactory practical design is given in Fig. 24. Over a range of gain adjustment of approximately 30dB, the departure from the ideal straight-line graph is no more than ±1dB. The unity-gain op. amp. follower at the left has been included so that the complete circuit presents a high input impedance to the source of V_{in} at all gain settings – this source may be the tape and radio inputs to a control unit, for example. Without this follower, the input impedance at maximum gain setting falls to 1.09kΩ.

Because the gain of the Fig. 24 circuit is independent of the total resistance of the pot. element, being dependent only on the slider tapping ratio, the tracking error between stereo channels can probably be held within ±1dB limits in production, over a 30dB range of gain, using low-cost carbon pots.

Alternative technique. An alternative technique, which, like the previous one, avoids the necessity to put fixed resistance in series with the pot. to limit the maximum gain, is shown in Fig. 25 in its simplest form.

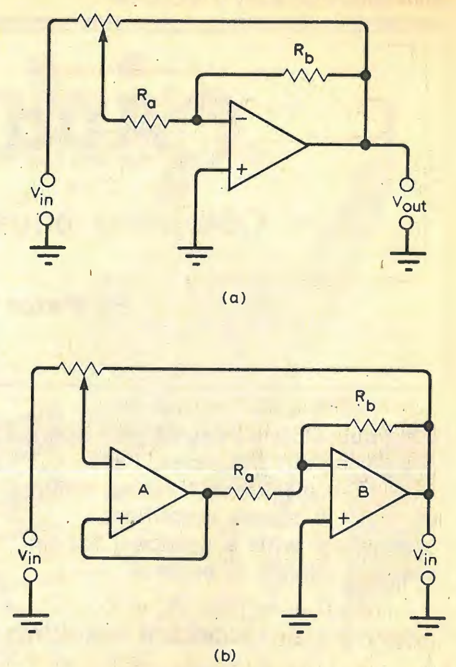
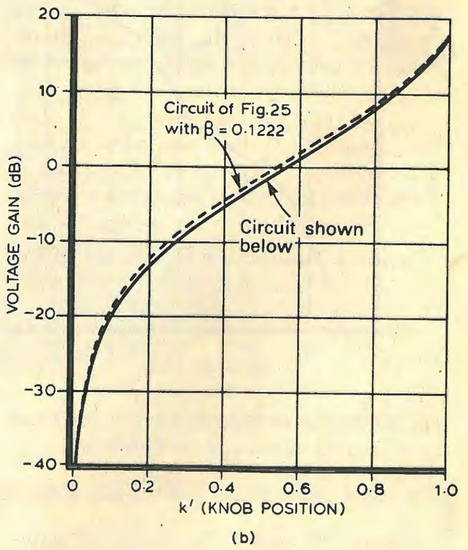
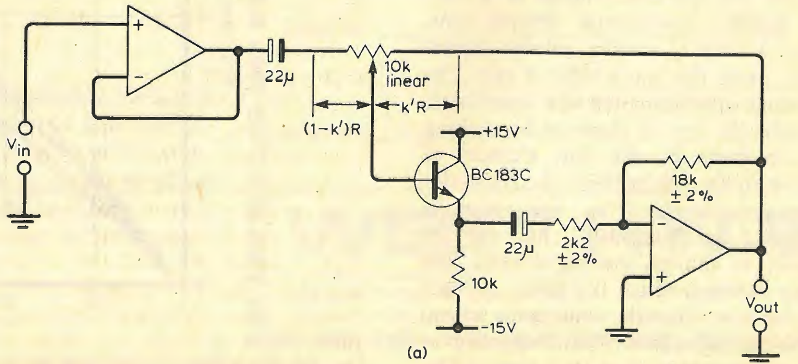


Fig. 23. Two circuits embodying the Fig. 22(b) idea. Circuit (b) uses voltage follower to avoid need for high-value resistors R_a and R_b.

Fig. 24. Practical version of Fig. 23(b) is shown at (a), with its control characteristic at (b) (lower curve).





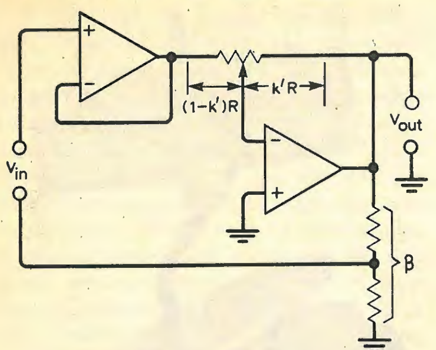


Fig. 25. Feedback amplifier limits maximum gain without use of fixed resistor in series with pot. Characteristic is upper curve in Fig. 24(b).

Here a fraction β of V_{out} is fed back as overall negative feedback in series with V_{in}. The forward gain, A, of this feedback system is $-k'/(1-k')$, so that applying the usual feedback formula gives:

$$\frac{V_{out}}{V_{in}} = \frac{A}{1-A\beta} = \frac{-k'/(1-k')}{1-[-k'/(1-k')]\beta}$$

from which

$$\frac{V_{out}}{V_{in}} = -\frac{k'}{1-k'+k'\beta} \qquad (7)$$

Comparing equation (7) with (5) and (6), it will be seen to be not quite of the same form, for the third term in the denominator of (7) involves k', whereas this is not the case in (5) and (6). Suppose we choose β in the Fig. 25 circuit so that equation (7) gives the same maximum gain, i.e. gain at $k' = 1$, as that given by the Fig. 24(a) circuit in accordance with equation (6). This requires $\beta = 0.1222$, and equation (7) then yields the broken-line curve shown in Fig. 24(b). Looking at these two curves, it is very tempting to conclude that the circuits of Figs. 24 and 25 inherently give slightly different shapes of characteristic, but more careful thought shows that this is actually not the case.

Referring to equation (7), this may be written:

$$\frac{V_{out}}{V_{in}} = -\frac{k'}{1-(1-\beta)k'}$$

$$= -\frac{1}{1-\beta} \times \frac{(1-\beta)k'}{1-(1-\beta)k'}$$

$$= -\frac{1}{1-\beta} \times \frac{k'}{\dfrac{1}{1-\beta} - k'} \qquad (8)$$

Equation (6) may be written:

$$\frac{V_{out}}{V_{in}} = -\frac{k'}{1-1/A-k'} \qquad (9)$$

Comparing (8) and (9), it will be seen that if A and β are so chosen that $(1 - 1/A) = 1/(1 - \beta)$, then the only difference between the equations is that the right-hand side of (8) is multiplied by the constant factor $1/(1 - \beta)$. This means that the curves for the two circuits are exactly the same in size and shape, but that represented by equation (8) is displaced upwards relative to the equation (9) curve by 20 log $1/(1 - \beta)$ decibels.

Thus, the real difference in behaviour between the circuits of Figs. 24 and 25 is that when designed to give identical shapes of control characteristic, the Fig. 25 circuit, at all knob settings, gives a slightly higher gain than does that of Fig. 24.

Passive control using linear pots.

A single linear pot. used as shown in Fig. 1 or Fig. 2 gives a control law which is quite intolerable for normal audio purposes. It is well known that by shunting a load resistor from the slider to earth, a characteristic approximating more closely to the ideal uniform decibel spacing may be obtained, though unfortunately only over a range of some 20dB or thereabouts. Fig. 26, based on calculations I did while a student in 1942, shows what happens as the loading is varied.

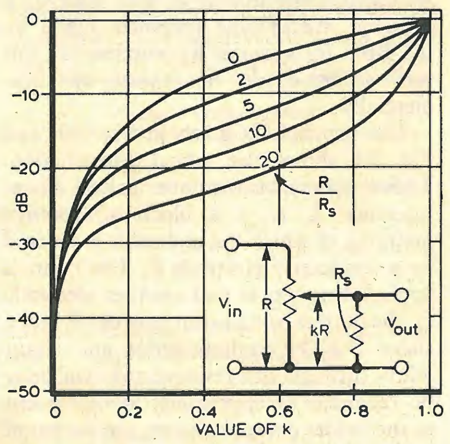


Fig. 26. Family of curves obtained from shunted linear pot. slider.

Very much better results than the above can be obtained with passive circuits using linear pots. if one or more fixed tapping points are provided, and the simplest such scheme is that shown in Fig. 27(a). If the resistors R_a and R_b are made of very much lower value than the pot. resistance, the attenuation with the slider at the tapping position is determined almost entirely by the values of R_a and R_b, and is virtually unaffected by any non-linearity in the law of the pot. element itself. There is, however, a sudden change in slope as the slider passes the tapping point, and a typical characteristic is shown in Fig. 27(b).

By adding a loading resistor between the slider and earth, a much better characteristic can be obtained, and it is possible to choose the value of this resistor so that there is no discontinuity in slope as the tapping point is passed. Fig. 28 shows a practical design employing a centre-tapped linear pot. with the slider output suitably loaded, together with the characteristic obtained. Over a control range of about 35dB, the departure from the ideal straight line is not much more than ±1dB. By having two tapping points on the pot. element – and low-cost slider pots. can be obtained with this feature – the nearly-linear control range can be extended to about 50dB if required, satisfying the most exacting needs.

For instrumentation purposes, the above technique can be extended much

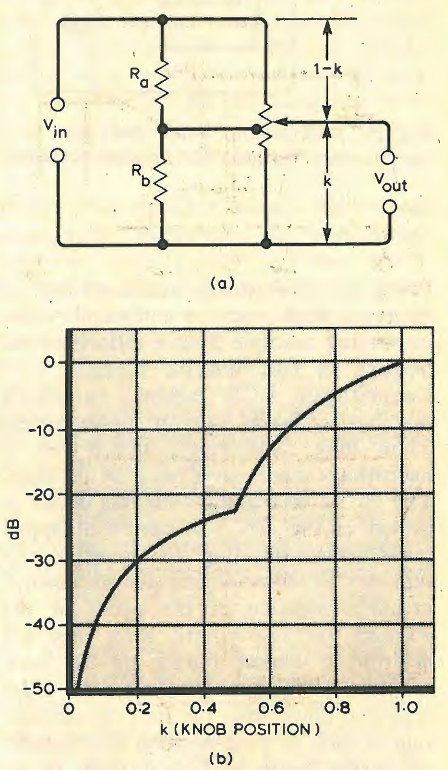


Fig. 27. Tapped linear pot. (a) gives approx. log. characteristic, shown at (b). With R_a and R_b low, gain at mid position is almost independent of track linearity or resistance.

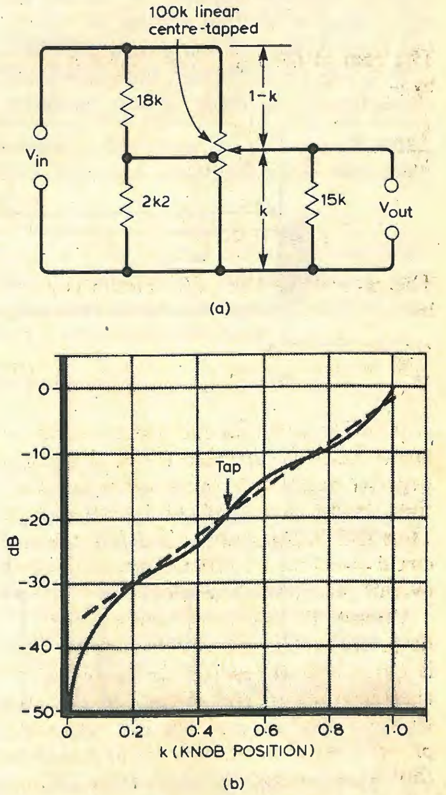


Fig. 28. Practical version of Fig. 27.

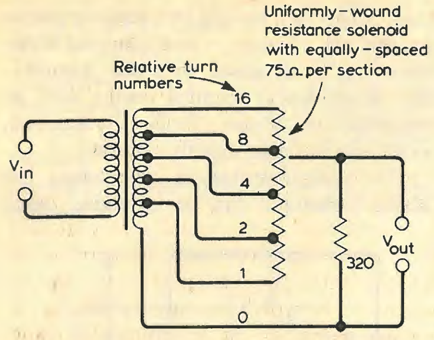


Fig. 29. Multiple-tap linear pot. with transformer-fed taps for precise voltages.

further, providing attenuators of extremely high precision and stability. An interesting example from a different field occurs in the Wayne Kerr B5009 Logarithmic LCR Bridge, in which readings are taken from an approximately 25cm long "slide-rule", which has a logarithmic scale covering a 16 : 1 ratio. The circuit associated with this device is shown in Fig. 29. The use of a tapped transformer winding to energize the tappings on the resistance element ensures extreme precision in the ratios of the voltages at these points, since they are determined almost purely by the turn numbers on the transformer. As the slider is moved down from the top, the attenuation at each tapping position increases by successive factors of 2, or 6.02dB. In the absence of the loading resistor on the slider, V_{out} varies linearly with slider position between tapping points, whereas, for a perfectly logarithmic scale, it is the log of V_{out} that is required to vary linearly. The error amounts to approximately 0.5dB midway between tappings. By adding the right value of loading resistor as shown, this error is reduced to less than ±0.05dB.

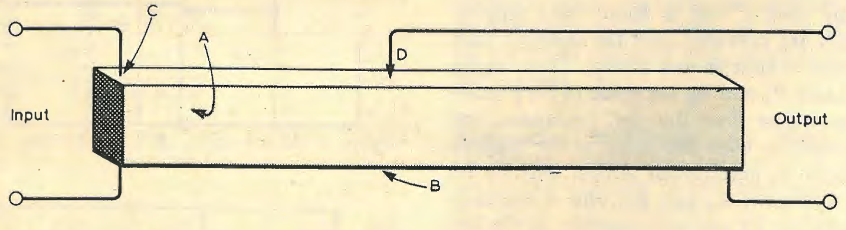
By using a transformer, the attenuation characteristic is made almost perfectly independent of production variations or non-uniformity in the resistance element, provided only that the physical positions of the tappings are accurately maintained. With the Fig. 28(a) type of arrangement, variations in pot. resistance do have some effect, but it may be kept small by making the resistance of the resistor-chain connected to the tapping(s) much less than the resistance of the pot. itself.

For high-grade audio control-unit applications, where the use of slider-type controls is considered appropriate, there would seem to be a strong case for using the Fig. 28 arrangement but with two tappings. By using ±2% resistors to feed the tappings, excellent stereo tracking should be obtained with a most desirable shape of control characteristic.

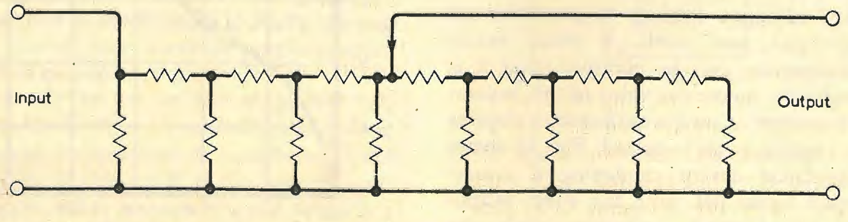
BBC log. attenuator

An interesting and very neat solution to the problem of providing a wide-range gain control having uniformly-spaced decibel scaling was devised in 1946 by C. G. Mayo and R. H. Tanner of the BBC Research Department. It was used in a portable microphone amplifier made by the BBC for acoustic measurements[5], but was unfortunately not taken up commercially.

The principle is given in Fig. 30, and Fig. 31 shows the actual construction. These illustrations are taken from reference 5. A is a block of resistive material, of which the underside is covered by a conductive electrode B. The input is applied between B and another electrode C, the output being taken between B and a slider D. The various series and shunt paths through the resistive material may be regarded as approximately equivalent to the ladder network shown, the output of each successive section of the ladder being a constant fraction of that of the previous section, giving a scaling with uniformly-spaced decibel divisions. The useful range of the model illustrated was about 70dB.



Simplified diagram of attenuator



Analogous circuit

Fig. 30. BBC gain control principle at (a) is 'distributed' equivalent to attenuator network at (b).

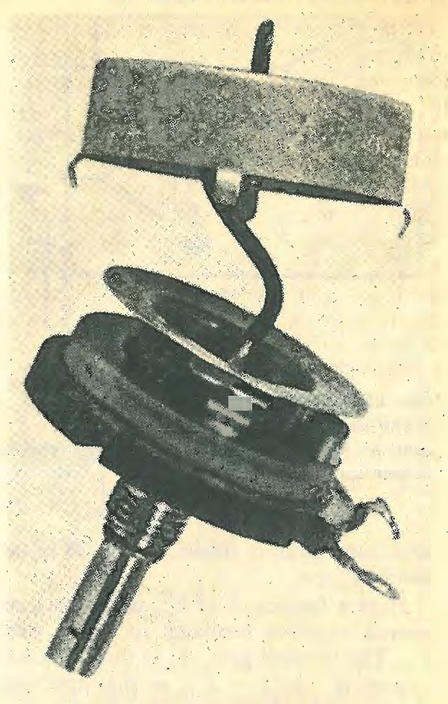


Fig. 31. Attenuator whose principle is shown in Fig. 30. Note screen round output. Photograph by courtesy of Electronic Engineering

It is pointed out that the output impedance of this type of attenuator does not become low when the attenuation is large, so that it is very important to avoid appreciable stray-capacitance coupling between input and output. The output connexion is therefore brought out coaxially, with a screening plate as shown in the photograph.

It has occurred to me that there is no essential need to employ a thick block of resistive material, and that an attenuator based on the same broad principle could be made using carbon-coated s.r.b.p. sheet material of the type commonly used in ordinary carbon pots. To test this idea, a quick experiment was done with the set-up shown in Fig. 32, and yielded the rather impressive result shown in Fig. 33. The very first graph obtained was somewhat inferior, apparently because of unsatisfactory contact between the steel vice jaw and the carbon coating. This was overcome by interposing a strip of polished copper foil between the carbon coating and the vice jaw.

Though an attenuator having a very extended range of operation as in Fig. 33 may fulfil some requirements, it is not ideal for use in control units etc., for the range of control needed in practice covers far less than 100dB, except that an "off" position coming soon after the position giving 40 or 50dB attenuation is really desirable. The Fig. 32 type of construction could readily be modified to provide such a characteristic, by shaping the conductive electrode, or metallic coating, somewhat as shown in Fig. 34. Halving the width of the carbon track, for example, would double the slope of the graph.

It is relevant to consider the suitability of attenuators based on the above principle for stereo purposes, i.e. whether sufficiently accurate tracking would be readily obtainable. Since the slope of the attenuation characteristic depends, to a first order at least, on nothing but the width of the resistive track, it would be important, for stereo use, to adopt a form of construction in which production variations in this width are minimized. The Fig. 34 construction does not appear to be ideal, for it relies on cutting the edge of the carbon material accurately in relation to the position of the metallized coating. The arrangement shown in Fig. 35 would seem much preferable, since accuracy of cutting is no longer involved and the metallized coating could be deposited by some form of printing technique with a very high degree of consistency.

The lower impedances usually used in transistor equipment, compared with earlier valve equipment, ease the problem of keeping the input-to-output stray capacitance sufficiently small, but it is still important to adopt a constructional arrangement which aims to minimize such capacitance. Working at 1kΩ impedance, with a control giving up to 100dB attenuation, the stray capacitance must be kept to less than 0.1pF. The connexion "rail" on which the slider moves must therefore be positioned away from the carbon surface and screened from this and the input connexion by an earthed screening plate.

Another advantage of the Fig. 35 arrangement is that, because of its symmetry, unwanted slight lateral movements of the slider during its traversal would be expected to have less effect on the attenuation than with the Fig. 34 form of construction – though it has been found that even with the latter, movements of about 1mm at right-angles to the direction of traversal produce only a small fraction of 1dB change in output provided the slider contacts the carbon track within 2 or 3mm of its edge.

Other methods of log. control and stereo tracking

- Perfect tracking of stereo channel gains at all settings, without the need for precision gain-control circuits, may be obtained by first producing, from the incoming L and R signals (L + R) and (L − R) signals. If the (L + R) signal is fed to one half of a ganged gain-control circuit, multiplying it by a factor α, and the (L−R) signal is fed to the other half of the gain-control circuit, which multiplies it by a factor β, then the sum of the gain-control circuit outputs is given by:

$$\text{sum} = (\alpha+\beta)L + (\alpha-\beta)R \qquad (10)$$

and the difference of their outputs is given by:–

$$\text{difference} = (\alpha+\beta)R + (\alpha-\beta)L \qquad (11)$$

Thus, though the balance as such is perfect, it is obtained at the price of introducing some cross-talk when α is not quite equal to β. The effects of stereo cross-talk are discussed in detail in reference 6.

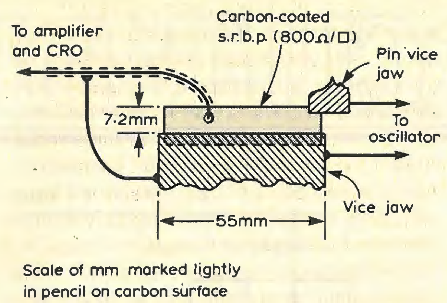


Fig. 32. Experiment using sheet instead of block in Fig. 30.

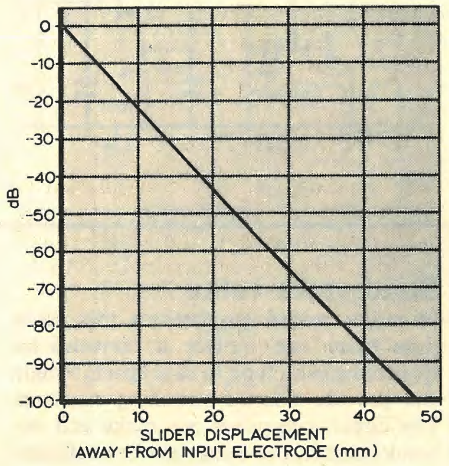


Fig. 33. Measured result obtained with Fig. 32 arrangement.

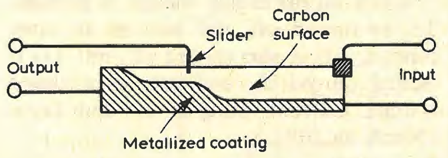


Fig. 34. Suggested form of control using Fig. 32 principle. Characteristic steeper at low-gain settings.

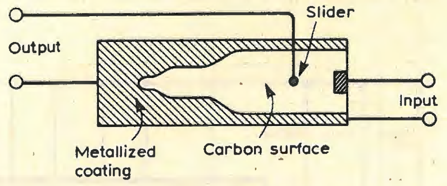


Fig. 35. Symmetrical version of Fig. 34 for improved consistency of performance.

- Perfect tracking without the introduction of crosstalk can be produced if a single gain-control circuit is used to control both channels. This can be done, for example, by first making the L and R audio signals modulate two different r.f. carrier frequencies, the two amplitude-modulated carriers being fed to the same gain-control circuit and being subsequently demodulated in phase-sensitive detector circuits. Though this technique could give virtually perfect results, it would not seem to be very attractive economically.

- Various simple gain-control circuits give a nearly linear relationship between attenuation in decibels and control position over a range of several dB. If a sufficient number of such circuits are put in cascade, and the controls are ganged, an approximately linear relationship may be obtained over any required range. While this technique is not usually very attractive when carried out literally with mechanically-ganged pots., it would appear to be worth bearing in mind as a possible technique for providing electronic gain control with a logarithmic characteristic. The idea is quite old.

- At the present time the most satisfactory technique for wide-range electronic gain control is that which exploits the fact that silicon planar transistors follow with high accuracy the relationship:–

$$I_c = I_o e^{qV_{be}/kT} \qquad (12)$$

where I_c is the collector current and V_{be} is the base-to-emitter voltage. (The other quantities are constants.)

Circuits can be designed in which the gain in decibels is linearly related to the control voltage over a range of about 100dB, and by using the "log-antilog" or predistortion technique, a performance sufficiently good, with respect to distortion and signal-to-noise ratio, to justify the use of such circuits in very high-quality audio systems, can be obtained. A very sound and clear treatment is given in reference 7.

This type of circuit is at the heart of compander systems of the dbx type. It could be used to provide gain control in audio control units, a single pot. varying the control voltage to a pair of such circuits in the two audio channels. The distortion and noise performance, though good, is not quite up to the highest standards sometimes demanded, maybe unnecessarily, in expensive control units, but some further refinement of i.c. versions of these gain-control circuits, including the reduction of residual even-harmonic distortion by the use of more fully balanced arrangements, may take place.

- In a fully digital audio system, gain control with perfect stereo tracking and any desired control law may be carried out on a purely numerical basis.

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