AUDIO HANDBOOK No. 2

Feedback

N. H. CROWHURST, A.M.I.E.E.

An essentially practical work, explaining clearly what feedback can do, where and how, and what its limitations are.

The theory of closed loops is presented in a simple form, progressively developed without involved mathematics.

There are design charts, developed specially for this book, which use a new and simplified approach to the problem.

Examples are given, illustrating the practical application to actual circuits, and showing how the methods can be applied to a wide range of problems and circuits.

With 40 diagrams drawn by the Author

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FEEDBACK

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INTRODUCTION

From the early days of what is still commonly described as "radio," there has been a deplorable gap, noticeable in all branches. The research worker, or advanced engineer, has explored principles and produced new developments, and has written them up in technical language, usually with full mathematical proof. On the other hand, the practical engineer, or experimenter, being under the necessity of doing the job as expeditiously as possible, has neither the time nor the inclination to learn all the "maths." To him much of the technical literature on the subject lacks practical details that can be applied to the job in hand.

The branch of knowledge concerning feedback has been no exception, and this book is published to help to close an obvious gap. Theoretical books explain the action in more or less understandable language ; but fail to give information in a form that can be readily applied to individual problems. Practical books usually give a number of circuits that have usually been tried and are known to work. But often the reader wishes to make up a circuit of his own, to suit his own ideas. When he tries to do this he finds he is still really groping in the dark, and just has to carry on until he stumbles across a workable arrangement.

For the benefit of the reader who is still a little hazy about the whole thing, this book first introduces, one at a time, the various problems set by feedback circuits, explaining them in simple language, so that he may have a good grip of his problems. Then methods of calculating are explained and a useful range of charts is provided to make the job easy. Finally some examples are given to illustrate the methods used.

Having read this book, the reader will be in a position to apply the knowledge he has gained to any circuit of his own. If he wishes, he can try the effect of different values in circuit, no longer groping in the dark; but using the charts to see why certain values become critical and others are unimportant. Such simple experiments will help him to see the practical significance of the conclusions the book draws. He is not confined to the circuits the book gives, and can easily apply the information to his own ideas.

To help application in this way, the more important graphs and abacs have been placed near the end of the book, which, for frequent reference, is considered to be a more convenient arrangement than having them distributed throughout the text.

The author would like to thank the many friends who have—knowingly and unknowingly—helped in the formation of this book, particularly by drawing his attention to pitfalls and fallacies often encountered, so that he can help his reader to avoid or dispose of them.

N. H. CROWHURST.

London, 1951.

I

WHAT FEEDBACK CAN DO

THE material given in this chapter will not be new to many readers who have read about feedback elsewhere. However, as an introduction, it is proposed to commence with the usual feedback " algebra," using it to show what feedback can do.

In this chapter feedback will be taken as being either " positive " or " negative." Beyond these simple ideas of direction, or " sense," the important question of phase is left to be dealt with in later chapters.

Gain

Whatever the reason for which feedback is applied, it always affects gain. As will be seen later, its other effects are directly related to its effect on gain, so this is discussed first as a basis.

Assume that an amplifier has a voltage gain denoted by the symbol A. Voltage from the output is fed back through a network that delivers a fraction, B, back to the input again. Figure 1 shows one possible arrangement that will serve the purpose of illustrating this point. Remember, A is always a large number, and B is a fraction—it may reach unity, but is never greater than unity.

The signal voltage e_g, fed into the input of the amplifier, produces an output voltage A times as large, or $e_o = Ae_g$. The fraction fed back is B times this, or $Be_o = ABe_g$. As this feedback voltage is in series with the input voltage to the whole arrangement, e_i, and is being considered negative, it reduces the effect of e_i as an input to the amplifier, so that

$$e_g = e_i - ABe_g, \text{ or } e_i = e_g + ABe_g$$
$$= e_g(1 + AB) \quad \ldots \quad \ldots\ldots \quad (1)$$

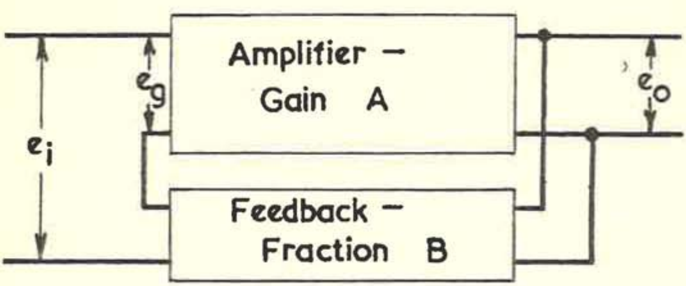

FIG. 1. BLOCK DIAGRAM OF SIMPLE NEGATIVE FEEDBACK ARRANGEMENT

From the user's viewpoint, the input is e_i and the output is $e_o = Ae_g$. Dividing both sides of equation (1) by $(1 + AB)$ gives

$$e_g = \frac{e_i}{(1 + AB)} \quad \ldots \quad \ldots\ldots \quad (2)$$

Using this to find the output in terms of e_i instead of e_g,

$$e_o = Ae_g = \frac{Ae_i}{(1 + AB)} \quad \ldots \quad \ldots\ldots \quad (3)$$

Dividing output by input gives the gain of an amplifier, so dividing the expression for output, (3) by e_i gives the gain with feedback applied as

$$A_f = \frac{A}{(1 + AB)} \quad \ldots \quad \ldots\ldots \quad (4)$$

The gain without feedback was A, so the connection of feedback has had the effect of reducing the gain by dividing by (1 + AB). The product AB represents the overall gain from the input to the amplifier, through the amplifier and back through the feedback to the input again, and is commonly called the "loop" gain. Adding 1 to this product gives the figure by which gain is reduced, or divided, when feedback is connected, so this whole expression, (1 + AB), is called the "feedback," or "feedback ratio." Converted into decibels it is called "db feedback."

When big feedbacks are used, expression (4) can be simplified by an approximation. In using an approximation, it must always be remembered that going beyond the realm where the approximation is near to the truth can lead to wrong results ; but carefully used, approximations can be very helpful. A big feedback means that adding 1 to AB will not noticeably alter its value : (101 is not much greater than 100). So, multiplying top and bottom of (4) by B,

$$A_f = \frac{AB}{(1 + AB)B} \quad \ldots \quad \ldots\ldots \quad (5)$$

and since AB is very nearly the same as (1 + AB), this is nearly the same as

$$A_f \doteqdot \frac{(1 + AB)}{(1 + AB)B} \text{ or } \frac{1}{B} \quad \ldots \quad \ldots\ldots \quad (6)$$

The sign $\doteqdot$ is used by mathematicians to represent "nearly equal to," when an approximation is used. Expression (6) shows that when large feedbacks are used, the gain is nearly independent of the amplification without feedback, A, and almost equal to $\frac{1}{B}$. Thus if one tenth of the output voltage is fed back, the gain will be almost exactly ten.

It has been assumed so far that the feedback is "negative." If it is positive the product AB will change sign and become negative, so *for positive feedback*, the gain with feedback is

$$A_f = \frac{A}{(1 - AB)} \quad \dots \quad \dots\dots \quad (7)$$

If the product AB, *i.e.* the loop gain, is less than 1, so that the voltage fed back is smaller than the input voltage, e_g, positive feedback will increase gain, owing to the fact that A is divided by (1 — AB), which is a fraction. To illustrate with figures : Suppose the loop gain is $\frac{3}{4}$; then (1 — AB) is $1 - \frac{3}{4} = \frac{1}{4}$, and the gain with feedback is $A_f = \frac{A}{\frac{1}{4}}$ or 4A. If the loop gain reaches 1, with positive feedback, the factor (1 — AB) becomes O, so the gain goes up to $\frac{A}{O}$ = infinity. In other words, the arrangement oscillates. AB being 1 means the voltage fed back is equal to the original input voltage. When feedback is connected like this, the fed back voltage *is* the input voltage, so the signal keeps on going through the amplifier in self-maintained oscillation.

Decibels or Gain Ratio ?

Most readers will be familiar with the usefulness of decibels in working out problems with amplifiers. It is quite possible to apply these units to feedback, but in some ways it is easier to think of what is happening in feedback circuits in terms of ratios. Decibels could still be used as terms of reference, and often are, but the corresponding ratio must be remembered to understand what is happening,—*e.g.* 6 db is 2/1, 12 db, 4/1, and so on. This being the case it may be better to work in ratios and convert to db when necessary.

A simple " slide-rule " scale is provided by Figure 27, showing db alongside corresponding ratios, given for convenience both as numbers larger than 1 and as fractions.

In some of the illustrations, scales of db are used and in others ratios are presented. To make thinking easier in both terms, the db scales in Figures 7, 8, 11, 17, 18 and 22 are marked off in 10 db steps, and, also, in 3 db steps.

The reason that ratio is the better term to think in will be particularly clear from the last part of the previous section on positive feedback. Suppose the loop gain rises from $\frac{3}{4}$ to $\frac{7}{8}$: then the amplifier gain with feedback will jump from $\frac{A}{\frac{1}{4}}$, or 4A, to $\frac{A}{\frac{1}{8}}$, or 8A. The change in loop gain is only a ratio of $\frac{7}{6}$, corresponding to about 1.3 db, but the change in amplifier gain resulting from it is 6 db. In fact a further rise in loop gain of about 1.2 db will make the amplifier gain rise to infinity by causing it to oscillate. A gain of another 1 db brings the loop gain to a ratio of about 0.98, raising the amplifier gain to $\frac{A}{.02}$, or 50A, a step-up of about 16 db from 8A. From these figures it is seen that the db relations are not very helpful : it is the actual ratios that count.

Distortion

Suppose that the arrangement of Figure 1 produces an output of $Ae_g + De_g$ without feedback, where De_g is some form of distortion, different in frequency from the fundamental. When feedback is connected, some distortion, BDe_g, will be fed back, as well as the fundamental, ABe_g. But as there is no distortion in the input signal, e_i, the amplifier input, e_g, must be altered by the distortion fed back, so that instead of being simply e_g, it could be written $e_g' = e_g - Fe_g$, where $-Fe_g$ is a negative distortion component due to the feedback. (It will not be $-BDe_g$, owing to the fact that changing the input will also change the output). So the output for this new input is $Ae^g + De_g' = A(e_g - Fe_g) + D(e_g - Fe_g)$.

Multiplying this out, the fundamental term is Ae_g, as before; the distortion terms are $-AFe_g$ and De_g, combined as $(D - AF)\ e_g$.; and the other term, $-DFe_g$, represents a distortion of distortion. As the original distortion is *usually* small, and is reduced by feedback to still smaller proportions, any distortion of distortion will be reduced to infinitesimal proportions, and can be neglected. The fed-back voltage can now be written, ABe_g, fundamental, and $B(D - AF)e_g$, distortion. The input voltage e_i is $e_g + ABe_g$, fundamental, as before distortion entered the discussion, and the $(BD - AF)e_g$ fed-back distortion must be the same as the Fe_g

negative component assumed in the input e_g. Thus

$$Fe_g = B(D - AF)e_g \quad \ldots \quad \ldots\ldots \quad (8)$$

Rearranging this, $Fe_g + ABFe_g = BDe_g$

and dividing through by $e_g(1 + AB)$

$$F = \frac{BD}{(1 + AB)} \quad \ldots \quad \ldots\ldots \quad (9)$$

The new output, with feedback, is Ae_g fundamental, as before, together with distortion given as,

$$(D - AF)e_g = \left(D - \frac{ABD}{(1 + AB)}\right)e_g$$

$$= D\left(1 - \frac{AB}{(1 + AB)}\right)e_g$$

$$= D\left(\frac{1 + AB - AB}{1 + AB}\right)e_g$$

$$= De_g\left(\frac{1}{1 + AB}\right) \quad \ldots \quad \ldots \quad (10)$$

As the output of fundamental, Ae_g, is the same as assumed in the absence of feedback, the reduction of distortion De_g, by dividing by $(1 + AB)$, shown in (10), means that distortion is divided by the feedback ratio, the same as the gain is.

Maximum Power Output

Before going further it will be well here to correct a common error. From the algebra in the previous section, it would appear that because feedback reduces distortion, given enough feedback, bigger outputs could be obtained from the same output valves with the same distortion. For example, an amplifier may give an output of 10 watts with 5% distortion, or of 15 watts with 20% distortion. Then it would appear that feedback to cut the gain by 4 (*i.e.* 12 db) should enable this same amplifier to give 15 watts with only 5% distortion. This *may* be true, or it may not : it depends upon the type of output used, and the kind of distortion it causes. The algebra in the previous section assumed that " Fourier analysis " of a waveform (*i.e.* analysing it into component sine wave frequencies) tells the whole story. But sometimes it doesn't.

Take the example of an amplifier that is practically distortion-free until " square-topping " starts. It is true that negative components due to the square-topping will be fed back to the input, which will provide further drive to try and off-set the square-topping. But, if the valves causing square-topping cannot be driven any further, this extra drive can do no good. As well as trying to provide more drive on the peaks, the feedback will tend to pare down the sides of the wave, and this latter effect may reduce the width of the square top. But it will only be equivalent to reducing the actual output. So, when an amplifier is square-topping, *for the same output level*, negative feedback cannot reduce distortion. But it will reduce distortion before square-topping sets in. Therefore the amplifier will actually distort more suddenly with negative feedback than without it.

If the distortion is of the " round-topping " variety, then feedback may help somewhat. For example, where triodes are driven in Class B_2 or AB_2, grid current drawn from the drive stage, as maximum output is approached, causes round-topping of the wave. Negative feedback, applied from the output back to before the drive stage, will cause the drive stage to produce a waveform made more peaky to off-set the round-topping. This process naturally has limits, as a point is reached where the drive stage can no longer produce the extra power needed to supply the rapidly increasing grid current at the extreme peak, and the amplifier starts to square-top instead.

Thus the effect of negative feedback on maximum power output is broadly the same in all cases : it can clean up the output as maximum is approached, but it always makes overloading occur more suddenly.

Hum

In the section headed " Distortion," if e_g is omitted from the terms representing distortion components, they could equally well represent fixed hum voltages injected within the amplifier. So it would appear from equation (10) that feedback will reduce these hum voltages in the same proportions as it reduces gain.

But hum is usually injected in the earlier stages, and the addition of feedback requires more gain to offset that lost by feedback. If the hum is caused in the earlier stages, then extra gain will increase the hum level in proportion to the increase in gain, and feedback, to bring the gain down to its original value, will also bring the hum back to its original level.

Valve Hiss

Following the reasoning used for hum, one would not expect feedback to reduce valve hiss, because this always starts in the early stages. In fact, however, negative feedback can actually increase valve hiss.

Naturally the extra gain necessary to offset feedback means that valve hiss coming from early stages will be amplified more.

If the feedback neutralised this amplification of valve hiss in the same way as it does hum, then two amplifiers of the same overall gain, one with and the other without feedback, would have the same amount of valve hiss in the output. But feedback does not neutralise valve hiss in quite the same way as it does other forms of signal started in the amplifier itself. Distortion of a signal passing through the amplifier, and hum voltages, are both of what is known as "periodic" waveform : they can be analysed, Fourier fashion, into one or more pure sine waves (of different frequencies). But the waveform due to valve hiss is not "periodic": it never repeats itself, but is a completely random succession of changes in voltage. This does not mean that feedback completely fails to reduce the higher valve hiss caused by higher gain. But to produce 100% reduction, so that valve hiss would be no higher than for an amplifier without feedback of the same gain, the feedback would need to work instantly, almost as if it could anticipate the coming changes, which of course it cannot.

Another way of viewing this matter is on the basis of frequency analysis. Hiss is made up of all frequencies from zero to infinity, although the amplifier only works over, say, the audio range. To cut down valve hiss by the same amount as increased gain has boosted it, the feedback ratio $(1 + AB)$ would need to be maintained throughout the range of frequencies covered by the hiss—an obvious impossibility. It might be thought that provided the feedback ratio is maintained throughout the audio range, then audible components of the hiss will be reduced in the same proportion as feedback reduces gain. But even this is not true.

No amplifier handles transients perfectly. Consequently transients outside the audio range can cause sounds within the range to be generated in the amplifier. A common proof of this fact is to be found in the reception of impulse noise ("static") by any radio receiver. The duration of the impulse is so short that it possesses no audio frequency components in the modulation envelope. But the failure of the audio section to handle it properly as a transient brings its effect into the audio range. A similar thing happens with valve hiss when feedback is applied, so that the amount in the output is increased, if the overall gain is maintained when negative feedback is applied.

Gain Stabilisation

This property of negative feedback has been hinted at by means of equation (6), showing that large amounts of feedback make the overall gain principally dependent on the feedback fraction, B. But it is useful to know how well this stabilisation of gain holds.

Gain is usually quoted in db units, which are logarithmic. So to find how much change in amplifier gain affects the overall gain with feedback, the method, which will be understood by readers with a knowledge of differential calculus (those without need not bother—the result is the important part), is as follows : Using equation (4) :

$$\frac{d \log A_f}{d \log A} = \frac{d \log A_f}{dA_f} . \frac{dA_f}{dA} . \frac{dA}{d \log A}$$

$$= \frac{1}{A_f} . \frac{1 + AB - AB}{(1 + AB)^2} . A$$

$$= \frac{(1 + AB)}{A} . \frac{1}{(1 + AB)^2} . A = \frac{1}{(1 + AB)} \quad \ldots \quad (11)$$

This means that the change in gain, expressed in db or other logarithmic units, is divided by the feedback ratio. For example, suppose there is 10 db feedback, corresponding to a feedback ratio of 3.162. Then a change in amplifier gain of 1db will cause a change of 1/3.162 or .3162 db in gain of the whole arrangement with feedback connected. If there is 20 db feedback, corresponding to a ratio 10/1, 1 db change in amplifier gain will be cut to .1 db change in the complete arrangement.

Frequency Response

The gain of an amplifier varies at different frequencies throughout the audio range. This variation is called its frequency response. Since negative feedback stabilises gain owing to such causes as variation in valve characteristics, changes in supply voltage, etc., (as shown in the previous section), it would be expected that negative feedback will also reduce variations in gain due to the signal changing its frequency, so that the frequency response would be made "flatter."

Variation of gain at different frequencies is due to the presence of "reactances" in the various circuits, in the form of inductances and capacitances. Not only do these reactances alter the voltage passed on at different frequencies, when the same voltage is put in, but they also shift its phase. This phase shifting can eventually result in turning negative feedback into positive feedback, and the result of connecting feedback will then be that, at a frequency where the amplifier showed a reduction in gain without feedback, it shows a marked rise or peak. It may even oscillate at some frequency, usually at the low or high end of the "spectrum," perhaps outside the audible range.

As the remaining chapters are largely devoted to various ways of understanding and calculating these effects, the reader is asked to leave the question of frequency response until he has read Chapter 2, and those following.

Input Impedance

Take first the arrangement of Figure 1, where the fed-back voltage is connected in series with the input to the amplifier. Usually the impedance of the feedback network is small compared to the input impedance of the amplifier, so its effect does not enter the consideration.

Suppose Z_i is the input impedance of the amplifier without feedback, due to a grid leak resistor, blocking capacitor, if used, and stray capacitance from grid to earth line or cathode of the first valve. According to Ohm's Law, if e_g is the input voltage, the current taken by this impedance will be $\frac{e_g}{Z_i}$. When feedback is connected the input voltage becomes e_i as given in equation (1); but the part of this actually across Z_i is still e_g, so the current taken from the input source, as the circuit is a series one, will still be $\frac{e_g}{Z_i}$. Writing Z_f to stand for the input impedance with feedback, it is given by input voltage divided by input current, or

$$Z_f = \frac{e_g(1 + AB)}{\dfrac{e_g}{Z_i}} = Z_i(1 + AB) \qquad \text{... ......} \quad (12)$$

So input impedance is multiplied by the feedback ratio, when series connection is used at the input.

In the circuit of Figure 1 it has been assumed that a voltage only is fed back. No current flows through the input source to which the amplifier is connected, and Z_i, due to the fed-back voltage ABe_g. If some current were to flow, then e_g itself would be modified in addition to the voltage ABe_g appearing across the output of the feedback network. For most circuits of this type, the assumption is well justified, and complications are avoided by being able to make it.

The feedback connection could be applied in parallel with the input instead of in series with it. When this parallel arrangement, shown at Figure 2, is used, it is best to consider the input on the basis of the currents passed through the input impedance, due to the input source, and also due to the feedback network. The combined current will produce a voltage which will be the input voltage for the amplifier with feedback connected.

If the current through Z_i is unchanged, the voltage, e_i, across

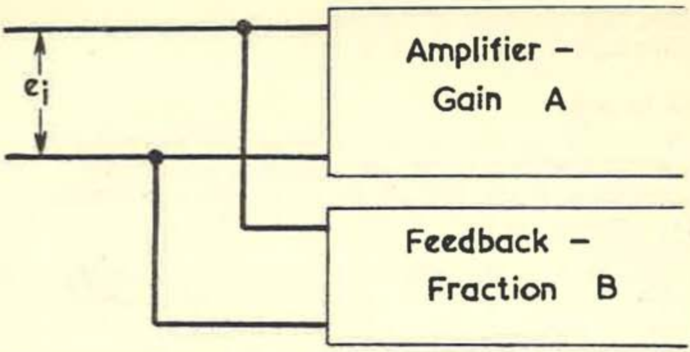

FIG 2. ALTERNATIVE CONNECTION AT INPUT END—PARALLEL INSTEAD OF SERIES

it must also be unchanged. Without feedback the current is $\frac{e_i}{Z_i}$. Due to feedback, AB times this current is passed in the opposite direction through Z_i. But since the total current flowing through Z_i is unchanged, the current taken from the input source by Z_i must be (AB + 1) times its value without feedback. So in this case the input impedance is effectively divided by the feedback ratio.

Output Impedance

In the circuit of Figure 1, the voltage across the output is fed back into the feedback network, *i.e.* it is a parallel arrangement. Circuits of this type are often called " voltage feedback."

Imagine an ideal, or perfect, amplifier (where the valves are supposed to have no curvature in their characteristics) without feedback. With no output load connected there is a certain open circuit output voltage—call it e_o. If the output is short-circuited, the current will be limited by the a.c. resistance of the output stage, r_A, to a value given by Ohm's Law as $\frac{e_o}{r_A}$.

Now suppose the same amplifier has feedback applied, cutting down the open circuit output voltage by the ratio (1 + AB). So the open circuit output voltage is now $\frac{e_o}{(1 + AB)}$. If the output is short-circuited, there is no longer any output voltage at the terminals, so there can be no feedback left. This means the short circuit current will be the same as it was when there was no feedback, $\frac{e_o}{r_A}$. But with feedback, the open circuit output voltage has been reduced so that from the viewpoint of the output circuit this short-

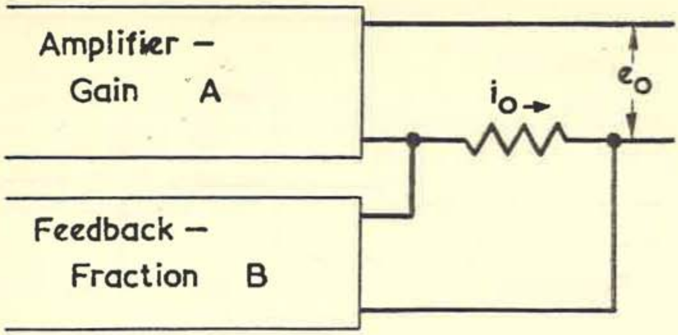

FIG. 3. ALTERNATIVE CONNECTION AT OUTPUT END—CURRENT FEEDBACK INSTEAD OF VOLTAGE FEEDBACK

circuit current is no longer due to a voltage e_o, but due to one $\frac{e_o}{(1 + AB)}$. So the *effective* a.c. resistance has been reduced to

$$\frac{r_A}{(1 + AB)}.$$

Thus voltage feedback has the effect of reducing the output a.c. resistance of the amplifier by the same ratio, $(1 + AB)$, as that by which gain is reduced.

There is another basic arrangement, called " current feedback," shown at Figure 3. The voltage fed back is due to the current in the output load, instead of the voltage across it. The feedback is taken from a position in series with the output.

In this case, when the output is open circuit, feedback will have no effect, so the output voltage e_o will be the same with or without feedback. But on short circuit, the effective output voltage is reduced by the feedback ratio, $(1 + AB)$, so the current is $\frac{e_o}{(1 + AB)}$ divided by the a.c. resistance, or $\frac{e_o}{(1 + AB)r_A}$. With current feedback, the a.c. resistance is therefore effectively multiplied by the feedback ratio $(1 + AB)$.

Summarising, voltage feedback reduces, and current feedback increases, the effective a.c. resistance, by the feedback ratio, *i.e.* the same ratio as that by which the gain is reduced.

For some jobs, such as amplifiers feeding into transmission lines, it is important to keep distortion down to a very low figure, needing a large amount of feedback ; but it is also important that the a.c. resistance should be some particular value, not too high and not too low, usually equal to the load resistance ; for the dis-

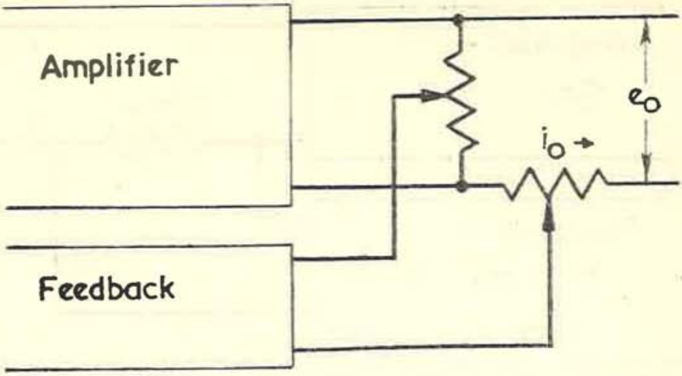

FIG. 4. BRIDGE OR HYBRID FEEDBACK ARRANGEMENT

tortion requirements, enough voltage feedback would make the a.c. resistance too low, while enough current feedback would make it too high ; so " hybrid " or " bridge " feedback is used, the arrangement for which is shown at Figure 4. Correct balance of each kind satisfies all needs.

How Far to Feed Back

This problem is made more complicated by the question of phase, discussed in later chapters. But there is also an aspect of it closely linked with the matters discussed in this chapter.

If feedback is carried over the whole amplifier, as shown in Figure 1, then the impedances connected to both input and output effect the amount of feedback. It was seen that voltage feedback into short-circuit, or current feedback into open circuit, results in no feedback at all. Similarly, at the input, for series injection, Figure 1, the source impedance is considered low, so that feedback voltage does not affect e_i. If the source impedance were infinite (fortunately a physical impossibility), there could be no feedback because there would be no return path for the fed-back voltage to get to the amplifier input. In the case of shunt injection, Figure 2, it is clear that zero source impedance would short-circuit the fed-back voltage. Therefore it is clear that the value of source impedance can affect the feedback ratio—whichever arrangement is used—just as the value of output load impedance does. Overall feedback results in both input and output impedances affecting the feedback ratio. This means output load impedance affects the amplifier input impedance, and conversely, the generator or source input impedance affects output a.c. resistance, according to the feedback arrangement used.

There are four possible combinations, excluding " hybrid " or " bridge " types : (1) series input and output ; (2) shunt input and output ; (3) series input and shunt output ; and (4) shunt input and series output. Each has its own properties, but the one of most practical use is (3).

Feedback need not be connected over a whole amplifier. The impedances appearing between stages are fixed by the circuit values, and are not affected by input or output connections to the whole amplifier. Thus, by applying feedback over only part of the amplifier, this interaction between input and output impedances is avoided, and by using separate feedback loops including the input and output circuits respectively, input and output impedance of the amplifier can be controlled quite independently.

It was shown in the section on valve hiss that feedback tends to increase valve hiss in an amplifier of given overall gain. By using feedback over only the later part of the amplifier, where no more valve hiss can come in, feedback will not increase valve hiss, because the hiss already in will be treated in exactly the way as any other signal handed on from the earlier stages. Feedback is not needed for reducing distortion in the earlier stages of a high gain amplifier, because the signal there is too small to be distorted. The only useful purposes feedback can serve over the early stages in a high gain amplifier are to control input impedance ; to stabilise gain ; or to control frequency response.

II

THE QUESTION OF PHASE

THE amplifying action of a valve is the same at all frequencies, and does not itself cause any phase shift—just reversal. This means that change of voltage in a positive direction on the grid causes a change in a negative direction at the anode.

The simplest case of phase shift—and probably the most common—occurs with the circuit shown at Figure 5. A voltage applied to the left hand side of the arrangement charges the capacitance through the resistance. If the voltage is steady d.c., the voltage across the capacitance reaches the same value after a time fixed by the relative values of the resistance and capacitance. If the voltage is a.c., the voltage across the capacitance follows the applied voltage somewhat later. Also, owing to the fact that the applied voltage has started to fall before that across the capacitance reaches its maximum, this maximum is not as high as the maximum of the applied voltage. Thus, the output voltage is less than the input voltage, and, also, it lags behind it by some portion of an alternating cycle.

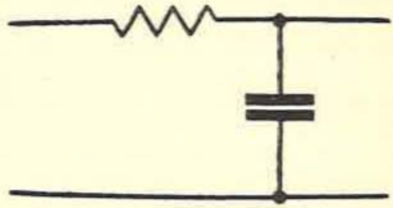

FIG. 5. BASIC HIGH FREQUENCY CUT-OFF ARRANGEMENT

Now consider the effect of applying different frequencies : the higher the frequency, the more quickly will the input voltage fall from its maximum toward zero, and the less charge will the capacitance have time to take before the input voltage falls below that on the capacitance, so that the direction of charging is reversed. This moment, when the input voltage passes the charge reached on the capacitance, is the point of maximum voltage on the capacitance. Figure 6 shows the waveforms at two different frequencies. At (*a*) the voltage across the capacitance is nearly as large as that at the input ; at (*b*), representing four times the frequency, the voltage across the capacitance is lower. The current will always be proportional to the difference between the two voltages, being controlled by the resistance value. The output voltage at (*a*) is lagging about $26\frac{1}{2}°$ behind the input voltage, while at (*b*) it has reached about $63\frac{1}{2}°$ behind.

Voltage and Phase Graphs

Waveforms could be plotted for all different frequencies, but this would become very tedious, so some more direct method of showing the effect is required. There are two principal methods. The commoner is to plot the output voltage on one graph and the

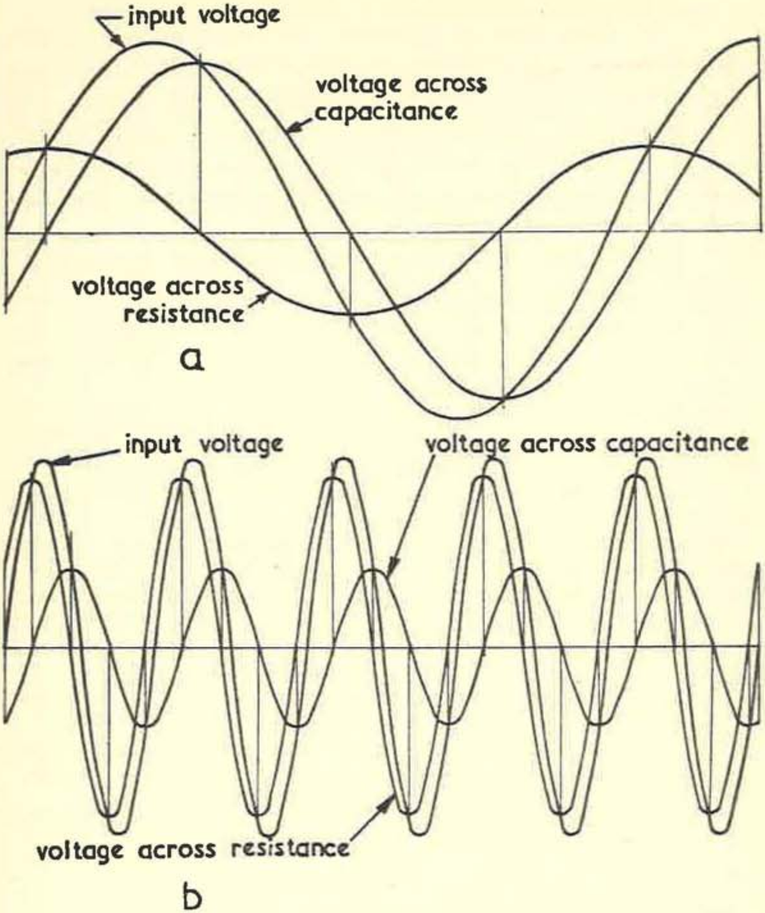

FIG. 6. WAVEFORMS ILLUSTRATING VOLTAGE REDUCTION AND PHASE SHIFT DUE TO ARRANGEMENT OF FIG. 5 AT DIFFERENT FREQUENCIES

phase shift on another, both against frequency. This method is shown at Figure 7. At (*a*) the actual voltage is plotted, as a fraction of the input voltage, while at (*b*) the voltages have been converted into db by means of the fractional ratios from Figure 27. At (*c*) the phase shift angle is shown. From these curves the drop in voltage and the phase shift angle at any particular frequency can be read off.

One particular frequency is of interest. It is the one shown in the centre of Figure 7, in this case 1000 cycles. The output voltage has fallen to .707, or 3 db, and the phase shift is 45°, half the maximum phase shift ever reached by this arrangement. The

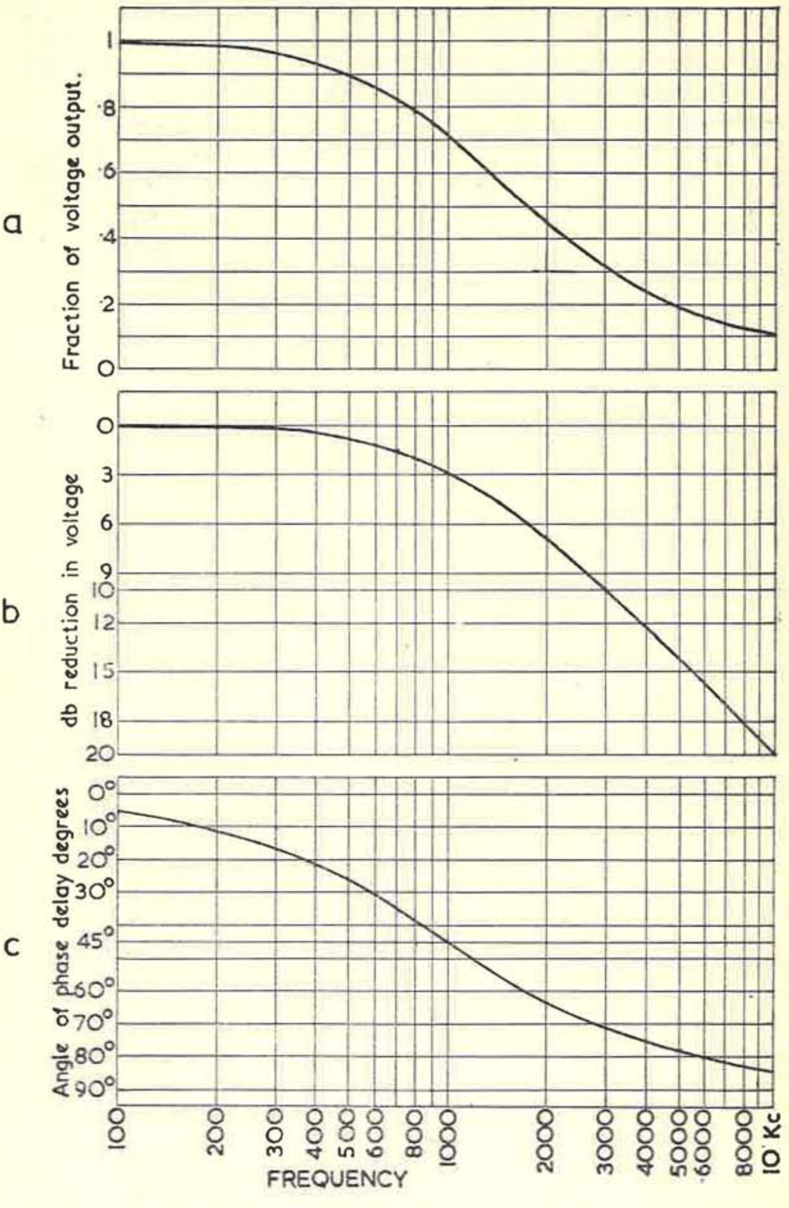

FIG. 7. RESPONSE CURVES, PLOTTING VOLTAGE REDUCTION AND PHASE SHIFT AGAINST FREQUENCY, HIGH FREQUENCY CUT-OFF

useful thing about this frequency is that it occurs where the reactance of the capacitance is equal to the resistance in value.

The chart given at Figure 29, page 54, enables the reactance of any capacitance or inductance to be calculated for any frequency ; or the frequency at which the reactance of the inductance or capacitance has a certain value may read off equally easily.

At any other frequency, having a particular ratio to this centre frequency, the drop in voltage and the phase shift will always be the same. For example, at 10 times the centre frequency, the voltage is always down to .1, or 20 db, and the phase shift just over 84°. This means that, by using a logarithmic frequency scale, which makes a certain horizontal distance always represent the same frequency ratio, all possible responses of this type, both in voltage and phase, can be drawn with a template made to the shape of the curves of Figure 7, according to the scale chosen. This is not only a useful aid to drawing them, it also helps in imagining possible variations.

In practical amplifiers, many of these simple "cut-off" networks occur, centred about different frequencies, as shown in Figure 8, and they can be drawn merely by sliding the template along to the required positions horizontally. The advantage of using a db scale for voltage is also seen here. The overall effect of a succession of these "cut-offs" is obtained by multiplying all the voltage ratios together for each frequency. By using db, equivalent to the logs of the ratios, it is only necessary to add the readings, instead of multiplying them. The total phase shift is likewise obtained by simple addition.

Vector Curves

Figure 9 shows the other method of presenting the "response" of the same arrangement, the network of Figure 5. It has the advantage of combining all the information in one "graph," instead of requiring two as at Figure 7, one for voltage and one for phase. It is based on vectors. The vector OA represents the input voltage, and the vector OB represents the output voltage at one particular frequency (about 640 cycles). The length of OB is proportional to the output voltage to the same scale as used for OA for input voltage. The angle AOB is geometrically the same as the value read off from Figure 7(*c*) for the frequency concerned. By drawing a number of these vectors for different frequencies, and joining up the points they give, a curve is made, relating the voltage to phase shift, voltage being the distance from O, and phase shift is the actual angle between OA and a line from O to the point on the curve.

Although this method gives a better idea of the performance of the network at a glance, it is not so useful for combining with other networks in the manner shown at Figure 8 for curves of Figure 7 type. Frequency markings can be added to each vector

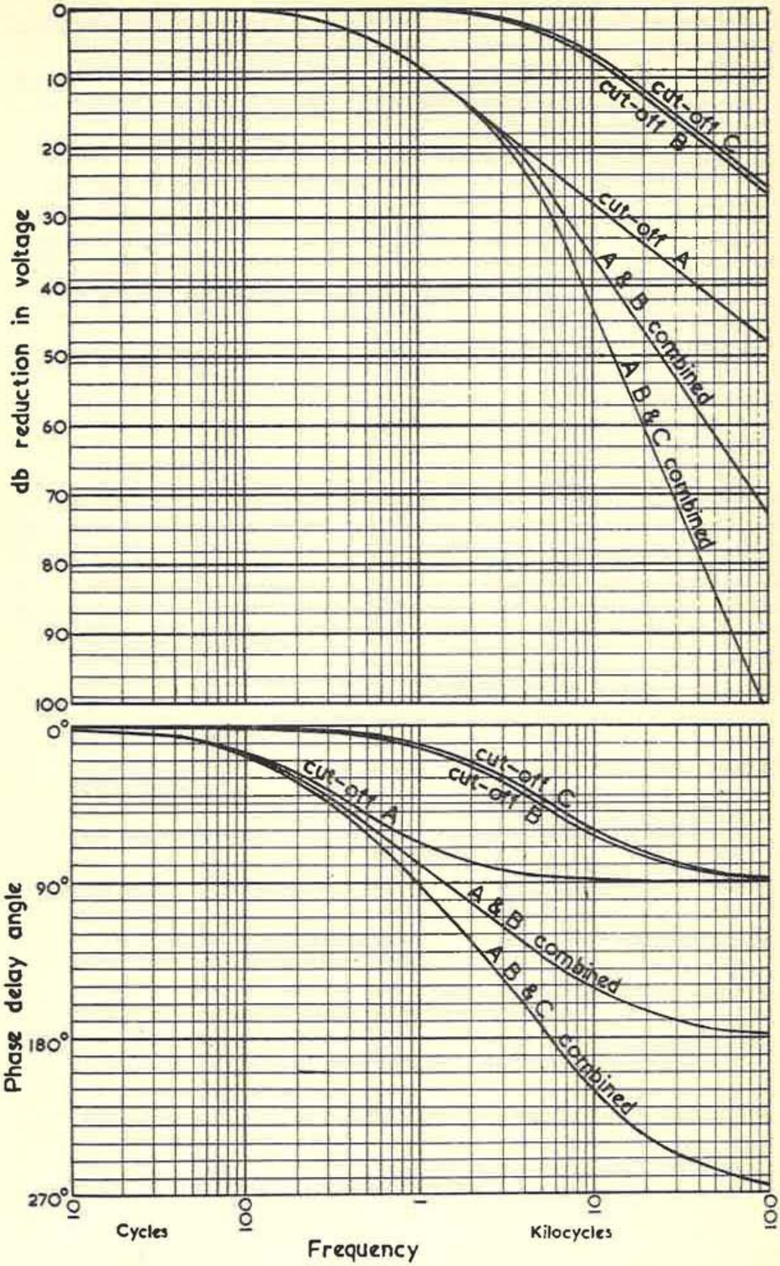

FIG. 8. USING A LOG FREQUENCY SCALE, db FOR VOLTAGE REDUCTION, AND PHASE ANGLE IN DEGREES, ALL HIGH FREQUENCY CUT-OFFS HAVE THE SAME SHAPED RESPONSE: AND THEY CAN BE COMBINED BY SIMPLE ADDITION

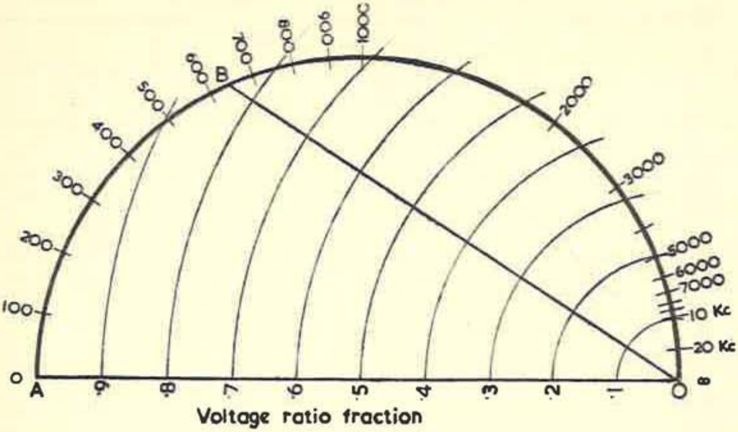

FIG. 9. VECTOR CURVE EQUIVALENT OF FIG. 7, FOR HIGH FREQUENCY CUT-OFF

curve as a scale along the curve itself, as shown in small figures at Figure 9 ; but to construct a curve for the combined effect of two or more such networks, the voltage ratio fractions due to each network must be multiplied together for each frequency, and the phase shift angles must be added together, from which a combined vector can be laid down for one point on the combined curve. The curve for a simple network with only one reactance is, conveniently, a semi-circle, which makes it easy to draw, but combined curves are more complicated in shape. The simplest way to draw a vector curve for a complete amplifier would be to use the method of Figure 8 for combining, and then convert the voltage ratio and phase angle curves into a combined vector curve, as explained for the simple curve with Figures 7 and 9.

The circuit of Figure 5, and all the curves shown so far, are for high frequency cut-offs. This simply means that low frequencies are passed without reduction or phase shift, and higher frequencies suffer progressively greater reduction and phase shift, until they disappear below the limit of audibility or measurability.

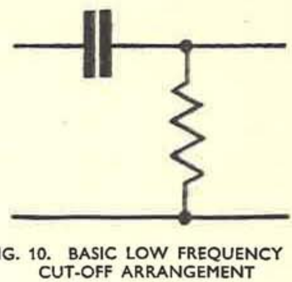

FIG. 10. BASIC LOW FREQUENCY CUT-OFF ARRANGEMENT

The coupling circuit, shown in its simplest form at Figure 10, commonly made by the coupling capacitor between stages, causes a low frequency cut-off. Its frequency response takes the form

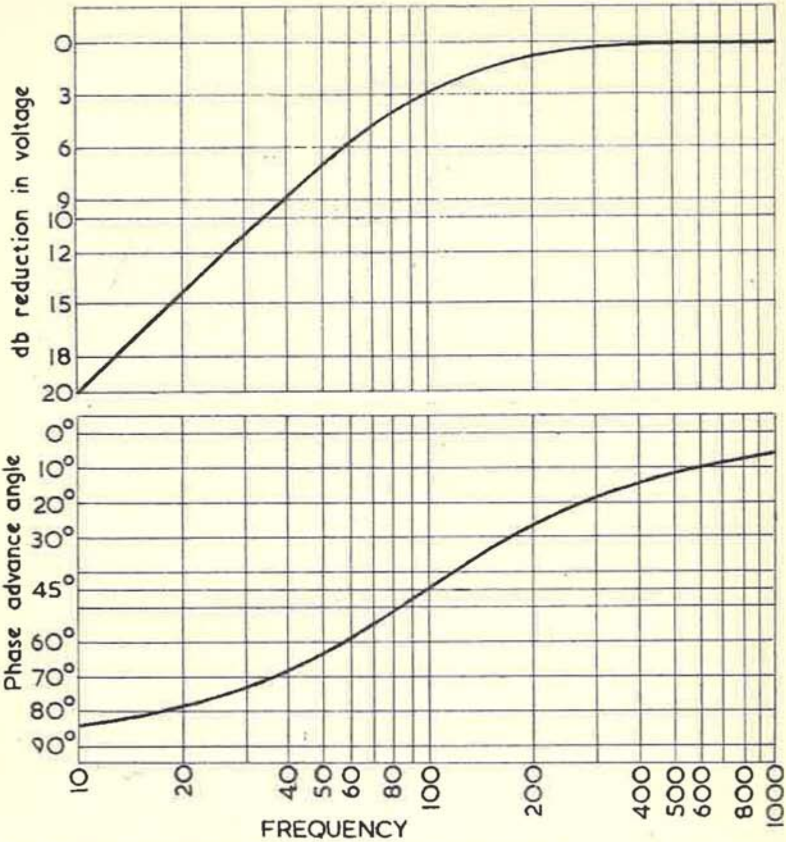

FIG. 11. RESPONSE CURVES, PLOTTING VOLTAGE REDUCTION IN db AND PHASE SHIFT, AGAINST FREQUENCY, LOW FREQUENCY CUT-OFF

shown at Figure 11 in voltage and phase, or at Figure 12 in vector curve style. Notice that the phase shift is in the opposite direction. A high frequency cut-off causes a phase delay, *i.e.* the output is later in phase than the input, but a low frequency cut-off causes a phase advance, so that the output is earlier in phase than the input. From this it appears that the output has to anticipate what the input is going to do. This is true. Proper transfer with a low frequency cut-off, only takes place when there is a steady note. When the note starts, the circuit is thrown " off balance," and it is a short while before it settles down to the correct steady state. However, for the purpose of analysis in this book, only steady notes will be discussed, and for these the output is a little ahead of the input in phase.

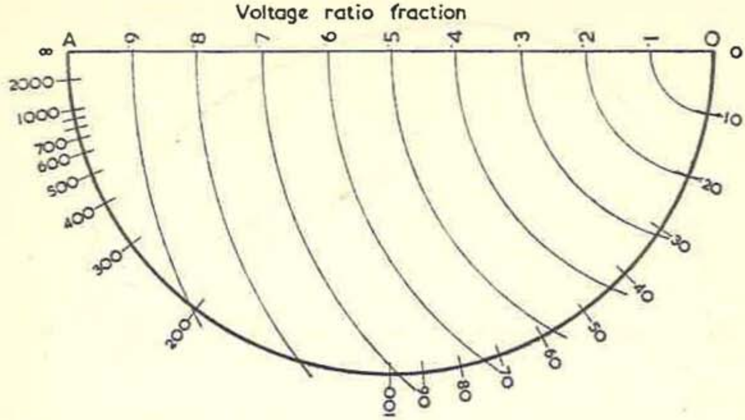

FIG. 12. VECTOR CURVE EQUIVALENT OF FIG 11, FOR LOW FREQUENCY CUT-OFF

Nyquist Diagram

Combining all these components that go to build up the response curve of an amplifier, into a special form of vector curve, makes up what is known as the " Nyquist diagram," leading to the well-known " Nyquist Criterion of Stability." It helps to interpret the meaning of equation (4) when feedback is not just simple negative, but includes a phase shift.

Figure 13 is a possible form for a Nyquist diagram. The length between the points 0 and 1 is used as the unit of measurement. Using this unit of measurement, the length OA is measured off to represent the value of AB in equation (4) for the amplifier concerned, at a frequency such that the feedback is purely negative, with no phase shift, *i.e.* where the capacitances and other reactances in the amplifier have no effect, or just balance one another's effects out. The vector curve for all the networks round the loop, through the amplifier and back through the feedback to the starting point, is drawn with A as the starting point and O as its centre. The curve going from A around O in a clockwise or right hand direction represents the high frequency cut-off of the loop, and the curve going anti-clockwise or in a left hand direction from A (only the beginning of this part is shown at Figure 13) represents the low frequency cut-off.

Now consider what the curve has to tell : at the frequency where the feedback is purely negative, OA on the diagram represents AB in equation (4) ; O1 is 1, so A1 represents (1 + AB) in equation (4). At any other frequency, the distance from O to the point on the curve representing that frequency represents the value of AB at that frequency in numerical quantity and phase angle. Take point C as an example : OC represents the numerical value of AB,

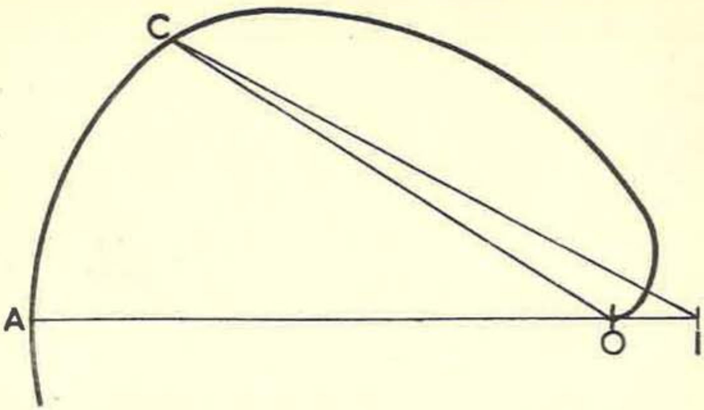

FIG. 13. SIMPLE FORM OF NYQUIST DIAGRAM. (ONLY H.F. CUT-OFF AND PART OF L.F CUT-OFF IS SHOWN)

and the angle AOC represents the total phase shift round the loop, both at the frequency represented by point C. If the product AB were negative, representing pure positive feedback, then it would be measured along the direction O1 on the diagram, and equation (7) would apply. But in the case under discussion, the presence of a phase angle means that neither equation (4) nor equation (7) apply by simple algebra. The diagram helps to show what takes the place of these simple equations when there is a phase shift. The distance OC represents the product AB, and O1 is 1 unit long, so, taking phase into account, the distance 1C takes the place of $(1 + AB)$ in equation (4), or $(1 - AB)$ in equation (7).

In many amplifiers using feedback, there is no phase shift at all in the feedback part of the arrangement ; B is constant and has no phase shift. Using equation (5), then, the gain with feedback is given as $\frac{AB}{(1 + AB)}$ times a constant, $\frac{1}{B}$. This being the case, the quantity $\frac{AB}{(1 + AB)}$, which includes phase shifts and other variations, can be used to find the frequency response of the amplifier with feedback, and the Nyquist diagram helps. Taking the frequency represented by point C, OC represents the product AB, and 1C represents $(1 + AB)$, so the ratio $\frac{OC}{1C}$ represents the

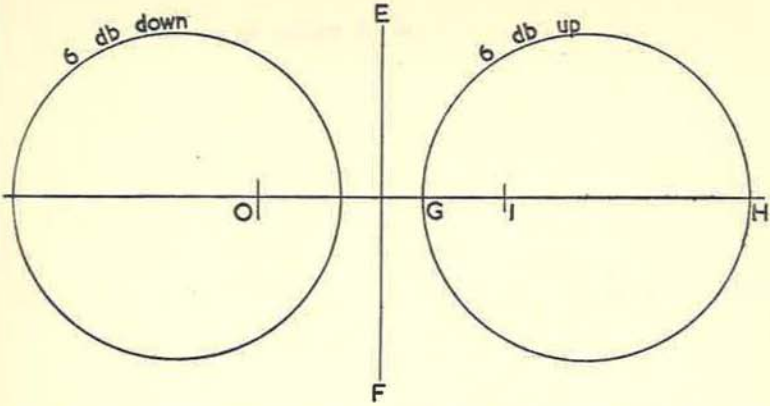

FIG. 14. CONSTRUCTION OF SCALE TO SHOW SHAPE OF RESPONSE FROM NYQUIST DIAGRAM

whole expression $\frac{AB}{(1 + AB)}$, taking the necessary phase shifts into account. To help read off this ratio, a kind of reference scale can be provided for the diagram. Figure 14 illustrates the method.

If $\frac{OC}{1C} = 1$, *i.e.* $\frac{AB}{(1 + AB)} = 1$, then the gain with feedback is $A_f = \frac{1}{B}$, which will be taken as a reference or zero level for the whole response. The gain where the feedback is purely negative will be slightly below this, but only very slightly if the product AB is large at this frequency, as is usually the case. So any point on the diagram such that OC = 1C will result in a response at zero level. All such points lie on a straight line bisecting O1 at right angles, EF in Figure 14. If $\frac{OC}{1C} = 2$, the gain will be doubled, or 6 db up. All points such that $\frac{OC}{1C} = 2$ lie on a circle whose centre is along O1 extended, and passing through points G and H on this line, which satisfy the conditions $\frac{OG}{1G} = 2$ and $\frac{OH}{1H} = 2$. Similarly, if $\frac{OC}{1C} = \frac{1}{2}$, the gain will be halved, or 6db down. The same kind

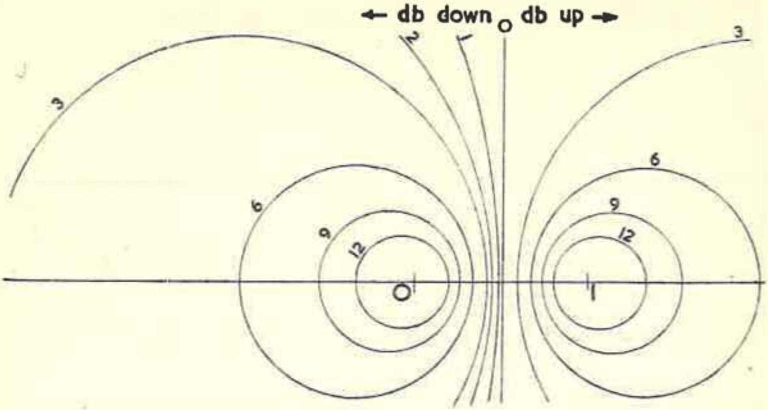

FIG. 15. SCALE PRODUCED BY CONSTRUCTION OF FIG. 14

of construction draws the circle representing 6 db down. In this way a whole family of circles can be drawn to make the required reference scale, as shown at Figure 15.

But although this construction does enable a frequency response to be shown eventually, the method is rather long and roundabout, and if there were no shorter cut, most readers would give up hope of ever being able to design or calculate feedback amplifiers at all, and decide that trial and error is the only way practical for them. However, all this does serve the useful purpose of giving a clear picture as to how Nyquist's criterion of stability works, and how negative feedback can cause peaks in the response. Figure 16 shows some imaginary Nyquist diagrams to illustrate these points. To avoid making the diagrams too confusing, the only part shown is the " inner end " of the high frequency cut-off.

At (*a*) is an example of a curve that will show a peak of about 10 db in the response curve, represented by the point C on the curve. It must be realised that the point 1 on all these diagrams represents infinite gain, or the condition for oscillation, so if the curve passes outside this point, the amplifier will be unstable and burst into oscillation. At (*b*) is shown an example of an unstable amplifier. Reducing the loop gain, as represented at (*c*) by making the curve smaller proportionately, will make this amplifier stable ; but there is still a peak, represented at point C. Further reduction will eliminate the peak, as shown at (*d*).

There is a condition that use of Nyquist's diagram can best explain, known as " conditional stability." It is illustrated at (*e*) and (*f*) of Figure 16. At (*e*) is a rather peculiar form of curve, that goes round the 1 point and comes back again (without going right round, as at (*b*)), finishing up by going into O on the right

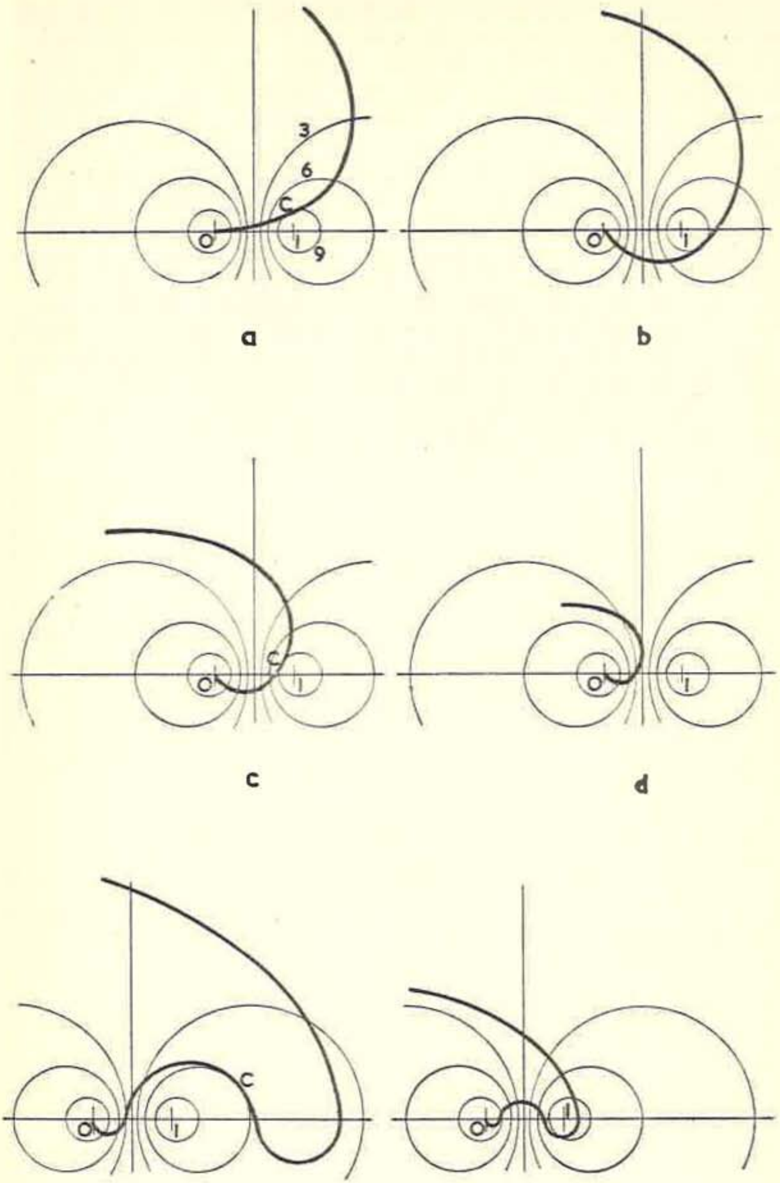

FIG. 16. SAMPLE NYQUIST DIAGRAMS TO ILLUSTRATE THE INFORMATION THEY CAN GIVE

side of point 1. According to the reasoning above, this amplifier will have a peak of about 6 db, at the point C, and will be stable. But if the loop gain drops, the peak at point C will rise, until after a reduction in loop gain of a little over 6 db, as shown at (*f*), it will oscillate. Having started the oscillations, increasing the gain will not stop them, but will cause the waveform to distort so that an extremely distorted oscillation will be maintained. These curves represent the behaviour at a succession of single pure sine waves of increasing frequency, so behaviour leading to oscillation of distorted waveform is not shown in detail by a Nyquist diagram or other response curve. Thus if this amplifier could be started up at full gain in the condition shown at (*e*) no oscillation would occur ; but any shock or variation (such as a transient) that even momentarily dropped the gain by 6 db, would start up a very harsh oscillation. In practice the amplifier would probably start oscillating as it warmed up, but there are some cases where it might not, as such instances of conditional stability do arise in practice.

III

USEFUL METHODS OF CALCULATING

THE understanding given by discussing Nyquist's diagram in the previous chapter paves the way for mathematical short cuts. The actual maths will not be given here, but the results are tabulated in the form of charts or graphs, so that calculations for practical circuits can be made quite easily.

The first thing necessary in working out the performance of a feedback loop, is to find how many of each type of elementary cut-off, represented in the simplest form at Figures 5 and 10, the loop provides. Then the frequency or time constant (explained later in this chapter) of each cut-off is calculated or fixed, according to the amount of feedback required and the frequency range, flatness or margin of stability decided upon. Application of these methods to actual coupling circuits is given in the following chapters. In this chapter, the discussion is arranged in order of the number of cut-offs involved, starting from simple ones, and progressing to the more complicated arrangements. At the end of the chapter a variation called the step circuit is discussed.

Single Cut-off

Feedback over a loop containing only one cut-off will simply extend the frequency range by the same ratio as the feedback applied. Figure 17 illustrates this, for a single high frequency cut-off.

Time Constant

For the presentation of frequency response directly, the 3 db point, used as the centre frequency for Figures 6 and 11, is the most useful reference ; but it entails calculation of the reactance of the capacitance or other reactance component, which, even with the aid of Figure 29, adds an extra step to the calculation for every cut-off considered. It is more direct to work on the basis of the " time constant " of the cut-off.

The time constant of a resistance capacitance combination is simply the product of R and C—resistance in Ohms and capacitance in Farads giving time constant in seconds. Of course, capacitance is never measured in Farads ; but an alternative product also gives a result in seconds ; it is resistance in Megohms and capacitance in microfarads. Taking resistance in Ohms and capacitance in microfarads, or alternatively resistance in Megohms and capacitance in picafarads, gives the result in microseconds. Resistance in

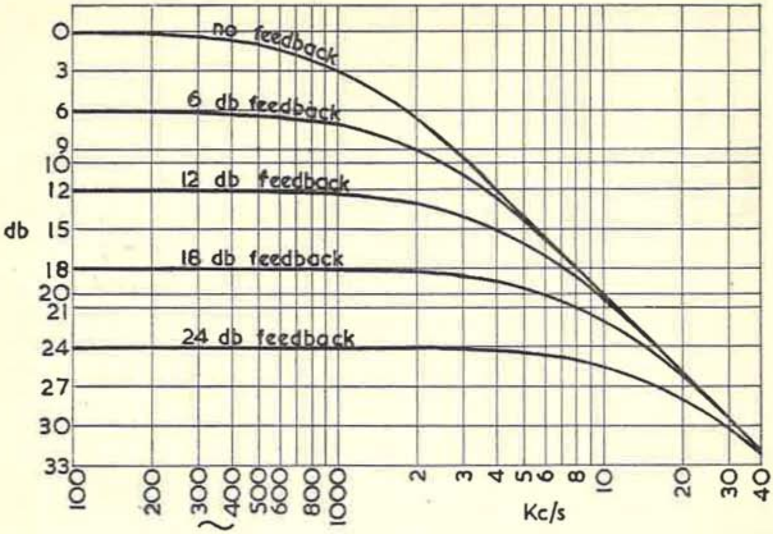

FIG. 17. SHOWING EFFECT OF NEGATIVE FEEDBACK WHEN THERE IS ONLY ONE CUT-OFF EFFECTIVE

Kilohms and capacitance in microfarads gives time constant in milliseconds.

The chart of Figure 30 provides a ready reckoner for calculating time constants. Inductance and resistance combinations also give time constants, calculated by dividing the inductance in Henrys by the associated circuit resistance in Ohms, the result then being in seconds.

Two Cut-offs

Feedback over a loop containing two cut-offs can never become unstable, because complete reversal of phase occurs only at a theoretically infinite frequency (in the case of high frequency cut-offs —zero for low frequency cases) where the numerical value of AB has fallen to zero. But this arrangement can cause peaks in the response. Figure 18 shows what happens when two identical h.f. cut-offs are used with various degrees of negative feedback.

Curve A shows the response of each cut-off by itself, and curve B shows the response of the two together, without any feedback. The straight lines C and D help in showing how the response varies as feedback is applied in varying degree. Line C is at a slope of 6 db per octave, which is the same as the maximum slope reached by a single cut-off ; the response always touches line C at one point. Line D is at the maximum slope of 12 db per octave reached by the two cut-offs, and the other responses always finish

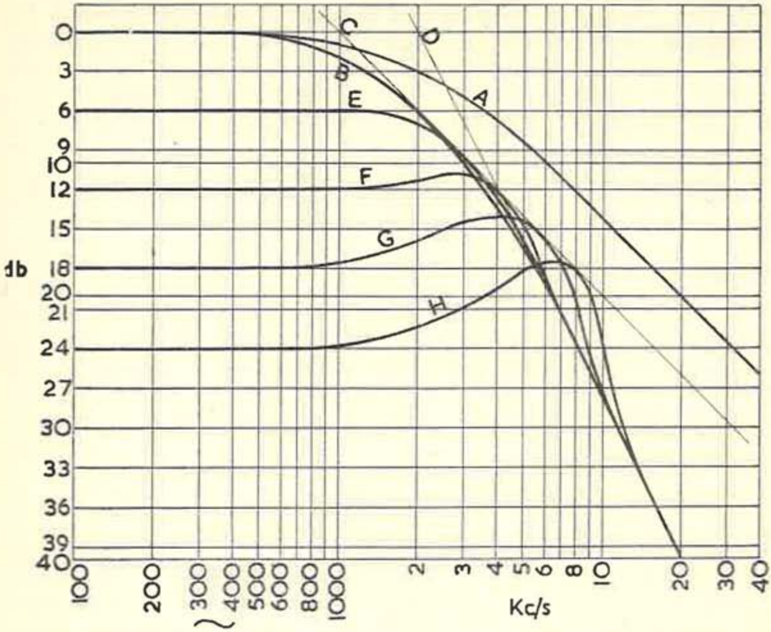

FIG. 18. SHOWING EFFECT OF NEGATIVE FEEDBACK WITH TWO IDENTICAL H.F. CUT-OFFS

up by running into this line. With no feedback the curve touches line C at a point 6 db below line D at the cut-off, or centre, frequency for both cut-offs. (2 Kc in Figure 18).

The maximum that can be applied in this case without starting to cause a peak is 6 db feedback. Curve E shows the response for 6 db feedback. It touches line C at a point 3 db below line D, at a frequency 1.414 times that for the point without feedback, (2830 cycles here).

Increasing to 12 db feedback produces a peak of $1\frac{1}{4}$ db, and the curve, F, touches line C where it crosses line D, at a frequency of 4 Kc, twice that for the point without feedback.

Raising the feedback to 18 db, curve G, gives 3.6 db peak, and the curve touches line C at a point 3 db above line D, at a frequency of 5660 cycles, or 2.83 times the original. Another 6 db, making 24 db feedback, curve H, gives 6.3 db peak, and the curve touches line C at a point 6 db above line D, at a frequency of 8 Kc, or 4 times the original.

Beyond this point, every 6 db more feedback increases the height of the peak by almost exactly 3 db, and the frequency of the peak is raised by the ratio 1.414.

The above applies to two identical cut-offs, *i.e.* two with the same time constant (not necessarily identical values of resistance and capacitance, so long as the product is the same). But by staggering the cut-offs larger feedbacks can be used before peaking starts, or with less peaking. The abac given in Figure 31 gives the amount of feedback that can be used without peaking, or the peaking caused by greater amounts, for different values of n, the ratio between the time constants of the two cut-offs. To illustrate its use, with n = 3, about 8½ db feedback can be used without peaking (using O on the db peak scale), and 20 db feedback with the same n gives about 3½ db peak.

Three Cut-offs

For more than two cut-offs it is not practical to give information in such a way as to foretell the exact performance for every possible combination of cut-offs with different degrees of feedback. To give the most useful idea for practical application, the best plan is to give complete information about the performance with identical cut-offs, and then details within certain limits, for other possible combinations.

With three identical cut-offs, Figure 19 shows the effect of different degrees of feedback on the position and height of the peak, relative to the centre or cut-off frequency, given by the chart of Figure 29 for the capacitance reactance or inductance reactance relationship of one of the cut-offs. Less than 3½ db feedback does not cause a peak at all. Up to 19 db feedback causes various degrees of peak, at which point the arrangement goes into oscillation. It should be noted that increasing feedback by 6 db beyond the 3.5 db where peaking first begins to show up, only causes about 3½ db peak. Extra feedback of 3 db (6.5 total) causes just over 1 db peak. On the other hand, the last 6 db feedback before instability is reached, from 13 to 19, raises the peak height from about 8 db to right off the map. From this it should be realised that a good margin below the oscillation point should be allowed.

The 9.5 db feedback point is a good one to note. The frequency of peak is the same as the cut-off frequency for each circuit. As shown at Figure 7 or 11, the drop for each cut-off at this frequency is 3 db, so the total for 3 cut-offs would be 9 db. Feedback of 9.5 db reduces the gain of the amplifier in the middle of the range by 9.5 db, and causes a response that has a peak of about 3.5 db at this previous cut-off frequency, so the gain has actually risen about 3 db at this frequency, as a result of connecting feedback, instead of falling 9.5 db. Greater degrees of feedback result in the peak being above the former cut-off frequency for high frequency cut-offs, or below it for low frequency cut-offs. Smaller degrees of feedback (between 3.5 and 9.5 db) cause a very small, almost flat, peak nearer the middle of the frequency range.

Figure 19 relates only to three *identical* cut-offs. If different cut-offs are used, they can be distributed in an almost infinite number of ways ; it is not a simple matter of giving one ratio,

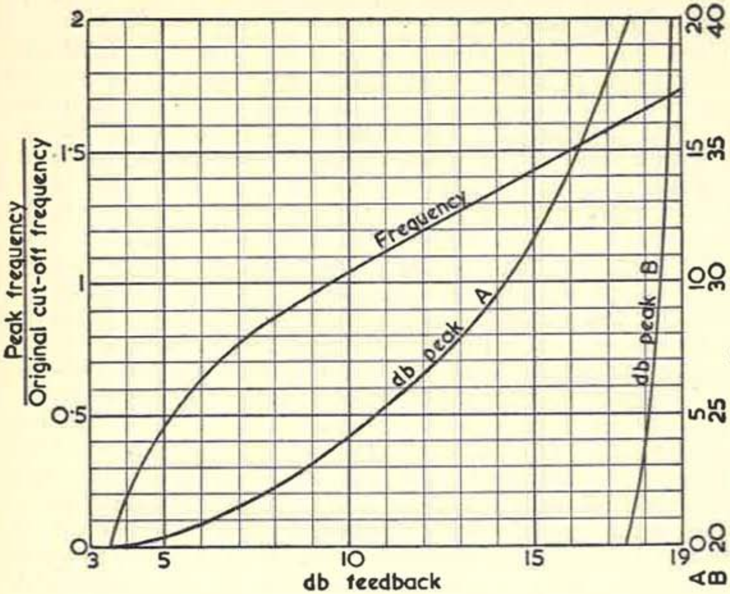

FIG 19. EFFECT OF NEGATIVE FEEDBACK ON THREE IDENTICAL H.F. CUT-OFFS

as for just two cut-offs. But it is possible to show limits beyond which the performance will not go. Figure 32 does this. Here n is the ratio between the largest and smallest time constant of the three used. Having specified this, the third one may be anywhere between these two. The limits are set (*a*) when the third cut-off is identical with one of the others, and (*b*) when it is mid-way between them (on a log scale).

The top two curves in Figure 32 show the feedbacks needed to start oscillation, the upper one with the third cut-off identical with one of the others, and the lower one with it staggered, mid-way. It will be noticed that oscillation starts with the same degree of feedback for a given ratio, whether the third cut-off is the same as the higher or lower of the other cut-offs. This is shown by the fact that the upper curve is symmetrical, as the left hand half represents cases where the third cut-off is the same as the larger time constant, while the right hand half represents cases with the third cut-off at the smaller time constant. This statement only suits high frequency cut-offs. To make it equally suited to both forms of cut-off, the left hand half applies where the two cut-offs act before the single one, while the right hand half applies to cases where the single one acts before the two. The staggered arrangement is obviously symmetrical, because the third one is always midway between the other two.

The lower curves in Figure 32 show the amount of feedback at which peaking begins. The upper right hand curve shows that when the third cut-off is made the same as the one with the smaller time constant (for high frequency cut-offs), or when the single cut-off acts first, considerably more feedback can be used before peaking starts, rising to just over 28 db for n = 100. Two or three db more feedback than that given by the curve will result in only about 1 db peak, as illustrated in the case of identical cut-offs by Figure 19. The lower left hand curve shows that making the third cut-off the same as the other one brings the feedback that can be used without peaking down considerably, although the same amount can be used before instability is reached. When n = .01, which merely means the ratio between the two cut-offs is 100/1, but now the third is like the other one, only 6 db feedback can be used before peaking starts, which is the same as for two identical cut-offs. In other words, the odd cut-off in this case does not materially affect the beginning of peaking, but it may well be the cause of ultimate instability, if too much feedback is used.

The lower curves marked staggered, show the amount of feedback to start peaking when the third cut-off is staggered between the other two. When n = 100, which means that the largest and smallest time constants will be in the ratio of 100/1, and the third will be in between, 10 times one and one tenth of the other, the feedback to start peaking is only about 15 db. The advantage of using two identical cut-offs of smaller time constant (higher frequency) for the high frequency end, in preference to staggered ones, can be seen. On the other hand, comparing the two and one arrangement for n = 10, given by the curve as 11 db, shows that removing the third cut-off by another ratio of ten, raises the allowable feedback before peaking starts to about 15 db. But removing both of them by this ratio raises it to over 28 db.

The extra feedback to cause a certain height of peak beyond the point where peaking starts will vary a little, but not greatly. For example, 6 db more feedback will cause not less than $1\frac{1}{4}$ db peak, and not more than about $3\frac{1}{2}$ db peak.

Four Cut-offs

The case for three cut-offs has shown that the best arrangement for cut-offs within a given n-range is to make one of larger time constant (lower frequency) than the others (for the high frequency end—vice versa for the low frequency end). To avoid complicating matters, the discussion of four or more cut-offs will only consider the case of one cut-off acting first, and all the others acting further from the response range, by some ratio n.

Figure 20 shows the variation of peaking with feedback for

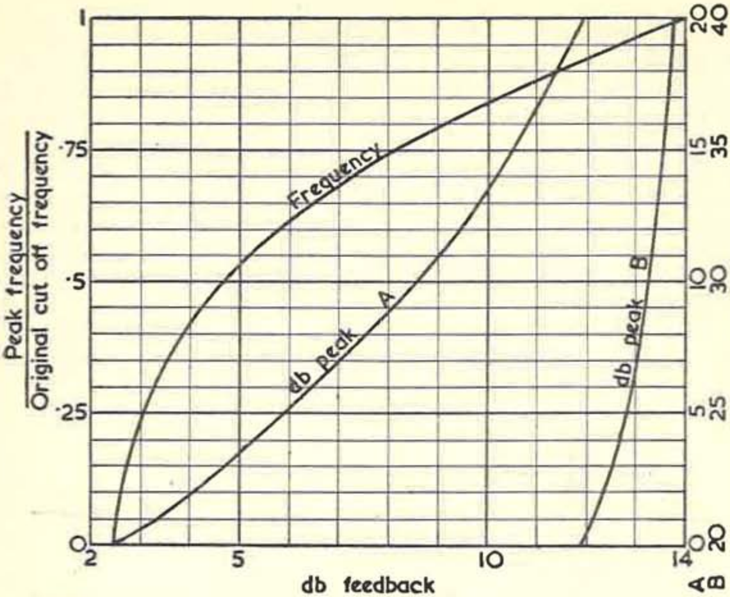

FIG. 20. EFFECT OF NEGATIVE FEEDBACK ON FOUR IDENTICAL H.F. CUT-OFFS

four identical cut-offs, between 2.5 db, where peaking starts, and 14 db, where oscillation sets in.

Figure 33 shows the feedback at which peaking begins, and the highest usable (when oscillation starts), for different values of n. For high frequency cut-offs this means that one cut-off has a time constant n times that of the other three. For low frequency cut-offs, the one cut-off must have a time constant one n-th of the other three.

Fractional values of n are not shown because they only produce a slight rise in possible feedback, reaching ultimately the same figures as given for three cut-offs when $n = 1$, *i.e.* 3.5 db before peaking begins, and 19 db for oscillation.

Five Cut-offs

Figure 34 shows the variation of the feedback for beginning of peaking, and instability, with different values of n, using a one and four combination. When $n = 1$, 2 db is where peaking starts and just under 12 db causes oscillation.

Again fractional values are not shown, because they cause only slight improvement compared with five identical cut-offs, reaching 2.5 db before peaking and 14 db for oscillation.

Step Circuits

All the circuits discussed so far have been of the "cut-off" type. This means that having started to introduce phase shift and voltage reduction towards the end of the frequency range, they go on with increasing phase shift and greater reduction in output, further from the range. Quite an appreciable phase shift can be noticed before voltage reduction is of any consequence. This is the chief reason why using more cut-offs so greatly reduces the amount of feedback that can be used. A step circuit helps to get voltage reduction with less phase shift than the normal cut-off circuit.

The two basic forms of step circuit are shown at Figure 21 with capacitance as the stepping reactance. The arrangement

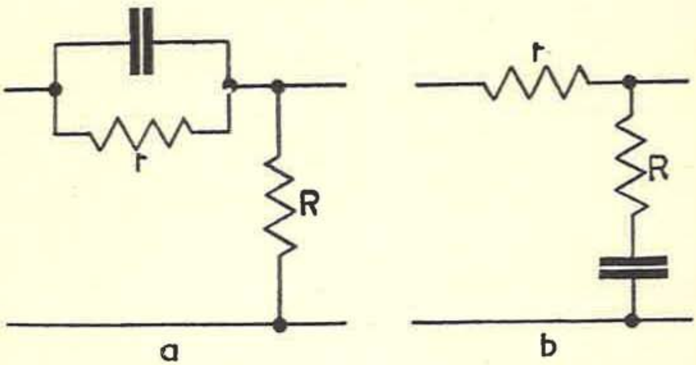

FIG. 21. BASIC FORMS OF STEP CIRCUIT

shown at (*a*) gives a downward step for use at the low frequency end, and that at (*b*) gives a downward step for use at the high frequency end. As all the discussion has been based on cut-off at the high frequency end, details of step circuit performance will be given on the same basis.

Figure 22 shows a presentation for a step circuit similar to the type shown at Figure 8. The step circuit can be regarded as a cut-off circuit at one frequency in conjunction with an "inverted" cut-off circuit at another frequency. This is one way of explaining its usefulness : the inverted cut-off "gets back" some of the phase shift caused by the ordinary cut-off.

There is an easy way of calculating the 3 db, or centre frequency of each part, using the chart of Figure 29 in the same way as for simple cut-off circuits. Using the arrangement of Figure 21 (*b*), the downward cut-off frequency is given by making the reactance of the capacitance equal to the combined series resistance (R + r).

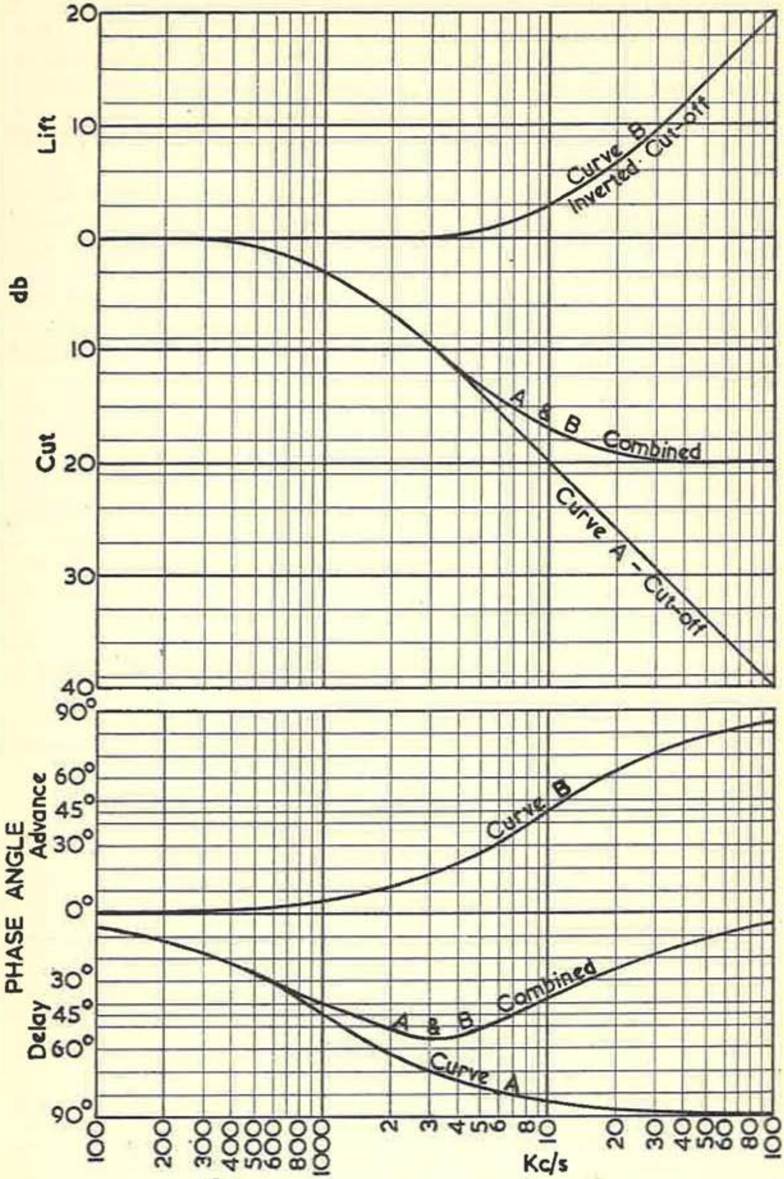

FIG. 22. STEP CIRCUIT CAN BE CONSIDERED AS A COMBINATION OF A CUT-OFF WITH AN INVERTED CUT-OFF

The frequency for the upward, or lift, component, is given by making the reactance of the capacitance equal to just the resistance R. The middle point of the step will be midway between these two frequencies on a logarithmic scale. The ratio between the two frequencies, or between the two time constants, will also be the maximum voltage reduction the step gives. The " slide-rule " scale of Figure 27 can be used to read this off easily in db.

Using the arrangement of Figure 21 (*a*), the downward cut-off frequency is given by making the reactance of the capacitance equal to the parallel combination of resistances r and R. The chart given in Figure 28 will help in calculating parallel combinations of resistances. The frequency of the upward, or lift, part is given by making the reactance of the capacitance equal to just resistance r. The middle point and the db height of the step will be calculated as stated in the previous paragraph.

The biggest reduction in phase shift is at the bottom of the step, where instead of being nearly 90°, as it is in a cut-off circuit, it returns to zero. Even at the point of maximum phase shift, which occurs at the middle of the step, the step circuit shows considerable improvement over a simple cut-off giving the same voltage reduction. Figure 23 shows the difference. Plotted against a double scale of db at middle of step for the step circuit, and also total step depth, the thick curve shows the phase shift at middle of step. Using the upper db scale, the thin curve shows the phase shift for corresponding db reduction by cut-off circuit. Near the bottom of the step, the phase shift improvement is still greater, because the step circuit is returning to zero phase shift.

The exact method of calculating for step circuits will vary somewhat according to the degree of feedback used, and the number of cut-offs combined with it.

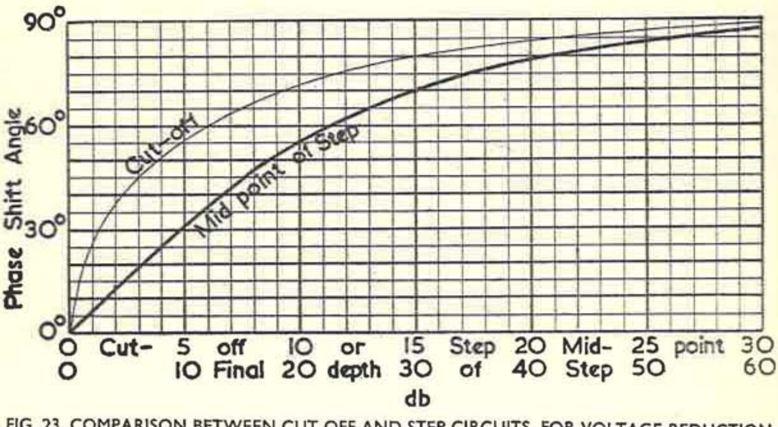

FIG. 23. COMPARISON BETWEEN CUT-OFF AND STEP CIRCUITS, FOR VOLTAGE REDUCTION AND PHASE SHIFT

IV

APPLYING CALCULATIONS TO PRACTICAL CIRCUITS

THE previous chapters have given information that will be useful in calculating the behaviour of amplifiers with feedback applied. But practical circuits are not quite so simple as those shown at Figures 5, 10 and 21. This chapter shows how to reduce practical circuits to " equivalent circuits " of these simple types, for the purposes of calculation.

Three sets of calculations must be made, and each coupling circuit for the whole amplifier will have a different " equivalent circuit " for each set of calculations. The loop gain where there is no phase shift is calculated by means of a set of equivalent circuits in which reactance is entirely absent. The low frequency cut-off requires a set of equivalent circuits bringing in those reactances that become effective at the l.f. end. The high frequency cut-off calculations require a set of equivalent circuits quite different again.

Resistance Capacitance Coupling

Figure 24 shows a typical coupling circuit of this type, together with the necessary equivalent circuits. At (*b*) is the equivalent for calculating gain. The amplification factor of the valve (given as μ in valve tables) for the operating conditions used, is multiplied by the fraction $\frac{R}{R + r_A}$, where R is the combined resistance of R_1 and R_2 in parallel, and r_A is the a.c. resistance of the valve V_1.

The equivalent circuit for low frequency cut-offs is shown at (*c*). The reactance of the coupling capacitor must be compared with the combined resistance r and R_2 in series. Here r is the parallel combination of r_A and R_1. When the reactance of the coupling capacitor is equal to $r + R_2$, the response is 3 db down and shows a phase shift of 45° advance. From the viewpoint of using the curves given for various arrangements to foretell the overall behaviour, use of the time constant $(r + R_2)\ C_c$, calculated with the aid of Figure 30, gives a quicker method.

The equivalent circuit for high frequency cut-off is shown at (*d*). The shunt capacitance C_p is made up of many parts. The effective anode to earth capacitance of V_1, the effective grid to earth capacitance of V_2, and all stray capacitances in the coupling circuit itself, need to be added together. If the wiring follows the good rule of keeping " hot " wiring to a minimum, by making the

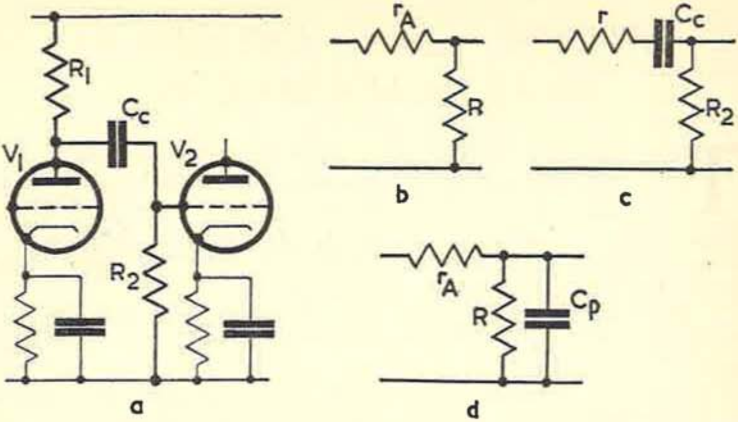

FIG. 24. CONVENTIONAL RESISTANCE CAPACITANCE COUPLING CIRCUIT, SHOWING EQUIVALENTS FOR CALCULATING GAIN, L.F. AND H.F. CUT-OFFS

connection between anode of V_1 and grid of V_2, through C_c, a short one, and having resistors R_1 and R_2 close to the valveholder pins, the stray capacitance due to the wiring itself will be small—only a few pF. Small coupling capacitors will not add much stray capacitance to earth, but larger ones will, because of their greater size. The anode to earth capacitance of V_1 will be practically the same as the combined capacitance from anode to all other electrodes, given in the valve data. The grid to earth capacitance of V_2 is partly dependent on the gain at which V_2 is working, as well as on the actual capacitances. This will be discussed briefly under the heading " Miller Effect."

Having found a figure for the total shunt capacitance between V_1 and V_2, the important comparison is between its reactance and the combined resistance of r_A, R_1 and R_2 in parallel. When this reactance and resistance are equal, the response is 3 db down, and the phase shift shows a delay of 45°. As for the l.f. cut-off, the chart given in Figure 30 can be used to find the time constant.

Choke or Transformer Coupling

There are so many " unseen " quantities when these forms of coupling are used, that calculation from scratch is very complicated. The simplest method is to make certain measurements using the choke or transformer in the same circuit in which it will be used. Probably the best method is to connect the second valve as a cathode follower, and then add a small pF capacitance equal to the grid to earth capacitance of V_2 as it will actually be used. Then the gain from the grid of V_1 to the output from V_2 as cathode follower is measured (at the same level as will normally be used), and a frequency response plotted by measuring the gain at different

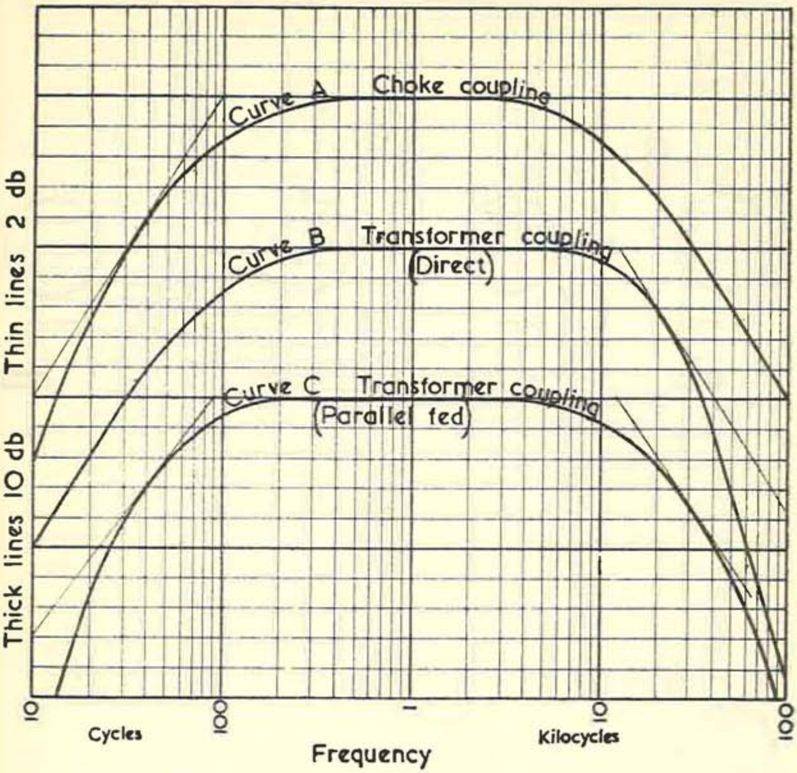

FIG. 25. TYPICAL RESPONSE CURVES FOR CHOKE COUPLING, DIRECT AND PARALLEL FED TRANSFORMER COUPLING, SHOWING CONSTRUCTION FOR APPLYING INFORMATION FROM THE ABAC OF FIG. 31

frequencies. The gain can be calculated for each frequency from the ratio of voltage output to voltage input. Each ratio is converted into db, and the response curve plotted on a logarithmic frequency scale. Figure 25 shows typical responses : at A for choke coupling, at B for direct transformer coupling, and at C for parallel fed transformer coupling. The circuits are shown at Figure 26, A, B and C respectively.

The points to notice are : Using choke coupling, the l.f. end is equivalent to two simple cut-offs, and the h.f. end to one ; using direct transformer coupling, the l.f. end is equivalent to a single simple cut-off, and the h.f. end to two ; using parallel-fed transformer coupling, both ends are equivalent to two simple cut-offs.

Where the cut-off is equivalent to a single simple one, the point where the response is 3 db down gives the cut-off frequency. If the time constant is required, any combination of resistance and

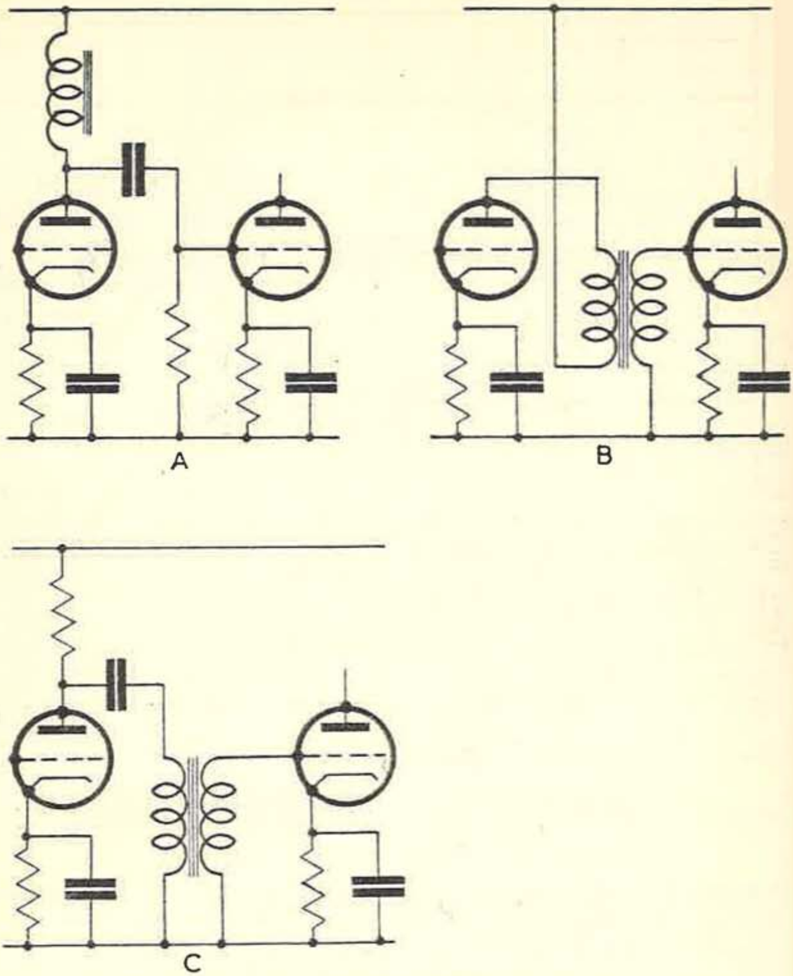

FIG. 26. THE CIRCUITS FOR WHICH RESPONSE CURVES ARE SHOWN IN FIG. 25

inductance or capacitance on the chart of Figure 29 giving that cut-off frequency can be referred to the abac of Figure 30 to find the time constant.

Where the cut-off is equivalent to two simple ones, the easiest way to foretell performance when there are not other cut-offs at the same end of the frequency range in the feedback loop, is by drawing a tangent to the response curve at a slope of 6 db per octave. The angle can be found by joining any two points on the paper representing a frequency ratio of 2/1 and a difference in level of 6 db. A line parallel to this, just touching the plotted curve, is then needed. Note how far below the "zero" level (where the response is flat) is the point where the tangent touches the curve. Subtract 3 db from this figure, and double the remainder. This gives the amount of feedback that can be used before the circuit commences to peak.

For example, at the l.f. end of curve A in Figure 25, the tangent touches at about 9 db down. This means that $2(9 - 3) = 12$ db feedback could be used before peaking starts. At the h.f. end of curve B, the tangent touches at about $4\frac{1}{2}$ db down, so $2(4\frac{1}{2} - 3) = 3$ db feedback only can be used before peaking starts. At curve C, the tangent for the l.f. end touches at 5 db down, and that for the h.f. end at about $7\frac{1}{2}$ db down. So $2(5 - 3) = 4$ db feedback can be used before peaking starts at the l.f. end, and $2(7\frac{1}{2} - 3) = 9$ db before it starts at the h.f. end.

The abac of Figure 31 can be used for these calculations as well as to see how much peaking further feedback produces. The scale on the left hand side of the centre line is aligned with O db feedback according to the position of the tangent. This finds a reference point on the right hand line from which alignment with db feedback gives db peak. For example, 20 db feedback to the response of Figure 25, curve C, causes about $5\frac{1}{2}$ db peak at the l.f. end and just over 3 db peak at the h.f. end.

It is not generally practical to use large degrees of feedback over big loops including intervalve or input transformers, although a small degree, calculated as shown, can be used over just one stage of amplification including one of these transformers or chokes.

Output Transformers

Usually there is not much of a problem here. Feedback may be taken from primary, secondary, or a special "tertiary" winding. The chief difference in calculation is that the turns ratio of the transformer affects the voltage available for feedback. An additional point is that feedback from the primary needs a d.c. blocking capacitor, causing an extra l.f. cut-off in the loop, and probably adding to the self capacitance for the h.f. cut-off.

There is an l.f. cut-off due to the inductance of the transformer, which can be calculated from the primary inductance and the a.c. resistance of the valve. The h.f. cut-off is usually due to primary capacitance, acting with the a.c. resistance of the valve. When a load impedance, such as a loudspeaker, is connected, it will modify the feedback, and the cut-off will be fixed by the whole arrangement. It is essential that the arrangement should be stable whether or not the load is connected, and also whatever frequency response the load impedance itself should take. Usually, whether the a.c. resistance is low or high compared to the load resistance, the effect of connecting it upon loop gain will be greater than its effect on overall frequency response, so the safest plan is to calculate the performance of the amplifier with no load connected, when it will usually be at least as good when the load is connected.

Most output transformers have a large step down ratio, so capacitances on the secondary side have little if any effect. Loudspeakers provide a load with a component of inductance that becomes effective at the higher frequencies in the range, and the inductance of the loudspeaker speech coil usually swamps any leakage inductance there may be in the output transformer. With a good output transformer there is little difference, beyond the effect of its ratio, whether the feedback is connected to its primary or secondary.

Step Circuits : Decoupling and Other Types

Step circuits affecting l.f. cut-off are usually part of the ordinary circuit of the amplifier, such as decoupling. Anode decoupling makes a step the wrong way, but its initial phase change would help, rather than hinder, since it causes phase delay while all other l.f. circuits cause phase advance. However, the time constant of anode decoupling circuits is usually so much larger than the coupling time constants, to satisfy its purpose of decoupling, that such help as it might give in phase delay occurs at too low a frequency to be of real help.

Cathode decoupling gives a step in the right way (so can screen decoupling) in the case of tetrodes or pentodes. Both these work this way because they are really a form of local feedback loop on their own.

Cathode decoupling is simple to calculate. The effect of the cathode resistor in producing negative feedback can be calculated by finding how much gain there is at the cathode (in the absence of a decoupling capacitor) from an input applied to the grid. The degree of negative feedback so provided tells the height of the step that the decoupling capacitor will produce. The frequency at which the reactance of the decoupling capacitor is equal to the cathode resistor gives the frequency of the inverted cut-off, or lift, produced by it. The direct cut-off component will have a frequency greater by a ratio corresponding to the db negative feedback caused by the cathode resistor. The time constant of the lift component can be calculated from the abac of Figure 30, using the actual values of cathode capacitor and resistor, and the time constant of the cut-off will be this value divided by the step ratio.

To find the height of step caused by screen decoupling, unhook the screen capacitor while the valve is handling a signal which is being properly decoupled (say 1000 cycles), and note the drop in gain so caused. This drop is the height of step caused by screen decoupling. The frequencies related to the step are then calculated in exactly the same way as for cathode decoupling, as outlined above.

Steps to help the h.f. end can be easily applied by shunting the anode resistor with a resistor and capacitor in series. Referring to the equivalent circuit of Figure 21 (*b*), r is made up of all the impedances at the anode, before the shunt is applied, in parallel. This will usually be the anode coupling resistor, the grid resistor, and the a.c. resistance of the valve. R is the resistor in series with the capacitor making up the shunt. In the example shown at Figure 39, the step is produced by connecting R_6 and C_3 across R_5. (See next chapter).

Miller Effect

Anything connected between the anode and grid of the same valve is part of a feedback loop, using that valve as the amplifier. Generally there is no deliberate connection of this kind. But there is inevitably a small capacitance between these two electrodes, usually much larger in triodes than in tetrodes or pentodes. The effect of the feedback through this capacitance is to modify the input impedance of the valve, and this effect is usually known as the " Miller Effect."

The most important aspect of Miller effect in relation to amplifier and feedback calculations, is that the anode load resistance causes an effective increase in the grid to earth capacitance. If the anode has no signal voltage on it—*e.g.* when the valve is used as a cathode follower—then the grid to anode capacitance is simply in parallel with grid to earth capacitance effectively. But, if the valve gives amplification at the anode, the effective capacitance from grid to earth, due to this grid to anode capacitance, is multiplied by the gain at which the valve is working. So, if the anode is the only electrode besides the grid that has signal voltage on it, the total grid input capacitance is the total capacitance from grid to all other electrodes, plus that from grid to anode multiplied by the gain at which the valve is working.

V

CHARTS AND EXAMPLES

THE " slide-rule " scale giving db and both fractional ratio and ratios larger than one, needs no explanation (Figure 27). For example, 14 db corresponds to a voltage ratio of 5 or .2.

The abac for calculating parallel resistances given in Figure 28 requires a little care. The scales on *either* the left hand side, marked A, *or* on the right hand side, marked B, may be used for any individual calculation. *Those on the same side of each scale must be used*, according to which set is best suited for the particular values. The figures may be taken to represent Ohms, tens of Ohms, hundreds of Ohms, thousands of Ohms, or whatever best suits the calculation ; but again *the figures must be taken to represent the same units on each scale*. For example : It is required to find the parallel combination of 1 Megohm, 1.5 Megohms and 220 Kilohms. (This is required later for Figure 35). Taking first the 1 and 1.5 Megohms, using the A scales to represent units of .1 Megohm (or 100 K), gives the combined resistance on the A centre scale as .6 Megohm or 600 K. Again using the A scales, but this time to represent units of 10 K, 60 and 22 combined, give a reading on the centre A scale of about 16, representing 160 K.

Another example : To find the parallel resistance of 330 K and 500 K. (This is required for the anode load of the 6F5 in Figure 38). Using the B scales to represent units of 10 K, 33 and 50 combined, give a reading of 20 on the B centre scale, representing 200 K.

The reactance chart of Figure 29 is direct reading, and requires no further explanation. For example : .005 μF has a reactance of 80 K at a frequency of about 410 cycles ; and 10 H has a reactance of 30 K at about 490 cycles. Any two quantities can be used to find a third. The chart can also be used to find inductances and capacitances that resonate at any frequency, but this use is not of importance in this book.

The time constant abac, Figure 30, is straightforward to use,

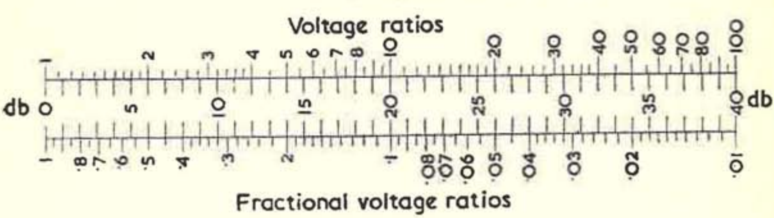

FIG. 27. QUICK REFERENCE " SLIDE RULE " FOR READING OFF db AND VOLTAGE RATIO

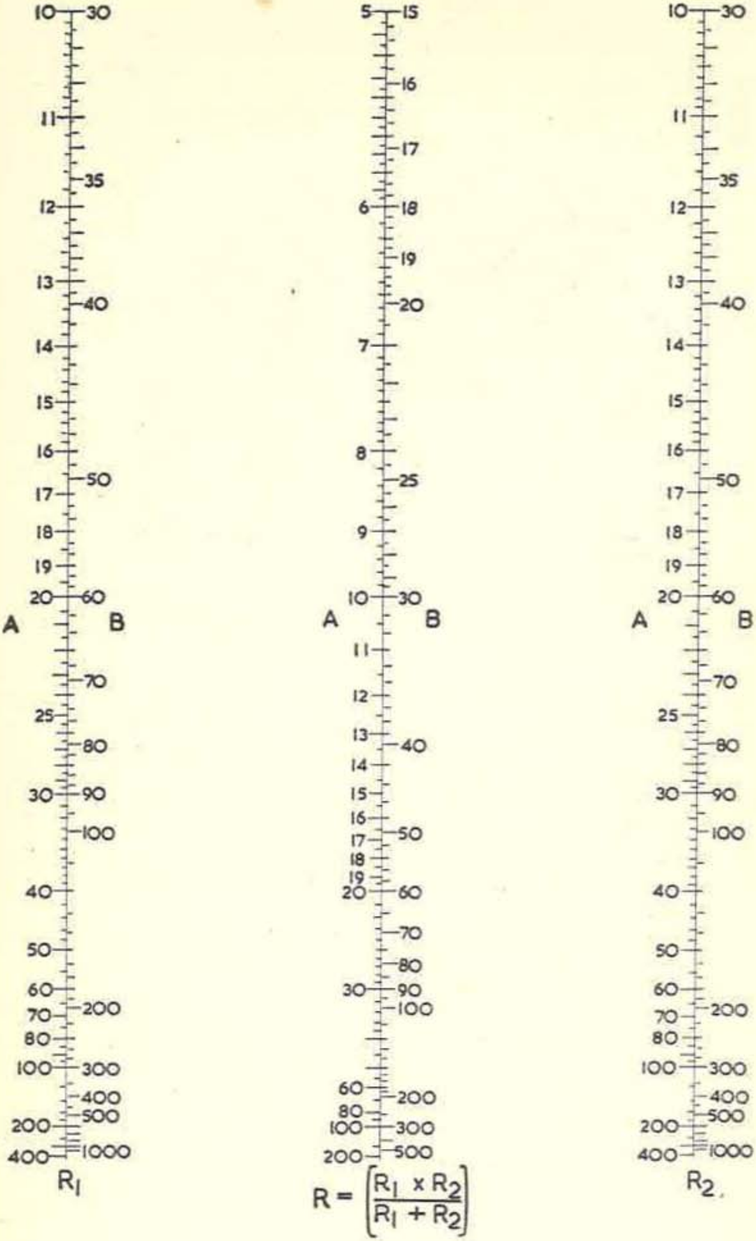

FIG. 28. ABAC FOR CALCULATING COMBINED RESISTANCE OF RESISTANCES IN PARALLEL

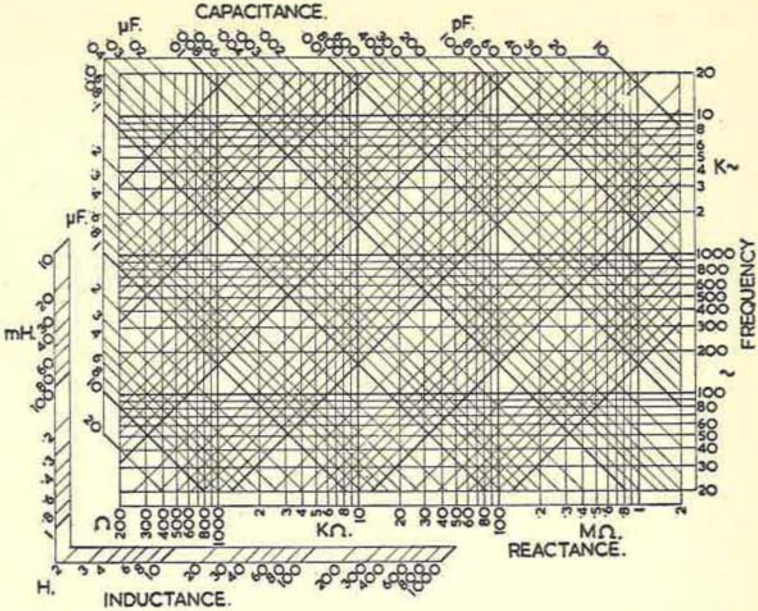

FIG. 29. CHART FOR CALCULATING REACTANCE OF CAPACITANCE OR INDUCTANCE AT ANY AUDIO FREQUENCY

but here again care is needed to use consistent scales. The centre time constant scale is the same whichever of the other scales are used. There is a choice of three for use in the other scales : C_1 with R_1 ; C_2 with R_2 ; and L with R. The scales are clearly marked with the units used, *e.g.* using scales C_1 and R_1, 50 pF with 1.6 Meg gives a time constant of 80 µ-sec ; using scales C_2 and R_2, .1 µF and 33 K gives a time constant of 3.3 m-sec ; using L and R, 4 H and 22.5 K gives a time constant of 180 µ-sec (or .18 m-sec).

The abac of Figure 31 has been explained in chapters 3 and 4, where examples were also given (see headings *Two Cut-offs*, and *Choke or Transformer Coupling*).

The limit charts given in Figures 32 to 34 have been explained in Chapter 3, under the headings *Three Cut-offs*, *Four Cut-offs* and *Five Cut-offs*.

Single Stage Feedback

A variety of circuits can be provided giving feedback over just one stage. A simple one is shown at Figure 35, using feedback over the 6J5. Taking the a.c. resistance of the 6J7 as 1.5 M., the bottom end of the negative feedback pot. is R_1, R_2 and the a.c. resistance of the 6J7, all in parallel. Using the chart of Figure

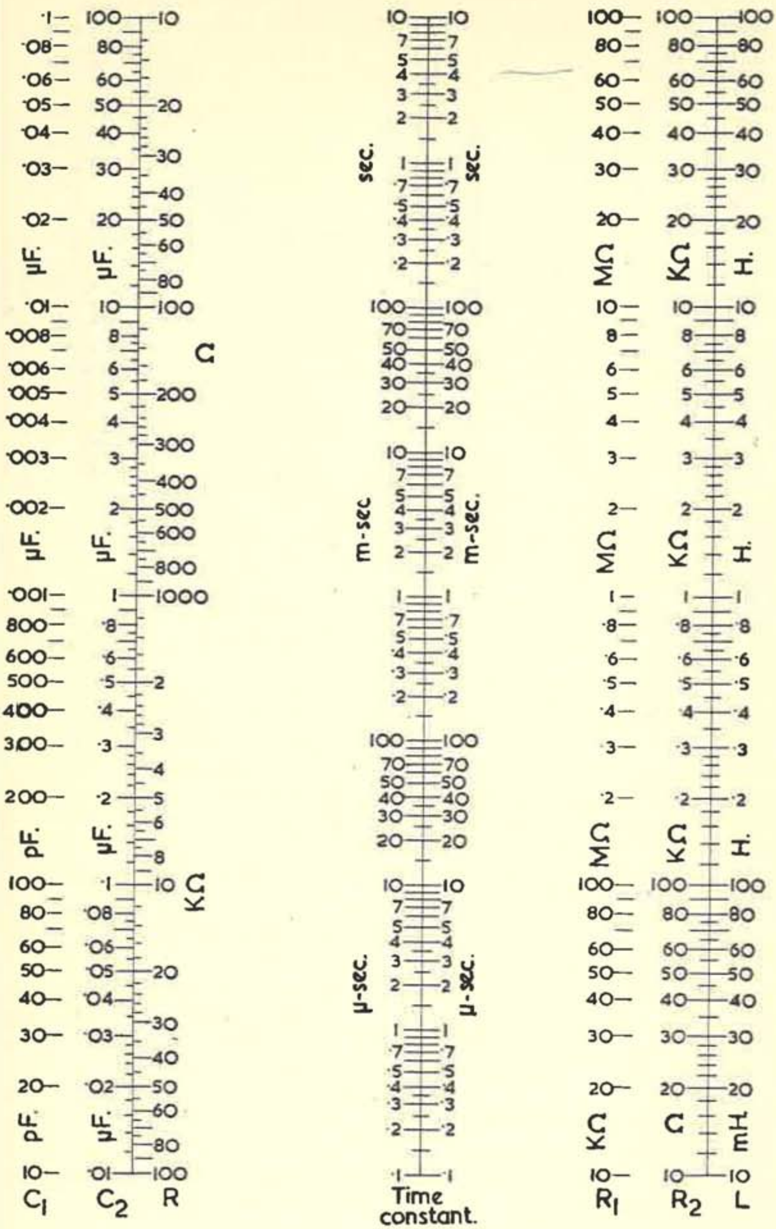

FIG. 30. CHART FOR CALCULATING TIME CONSTANTS

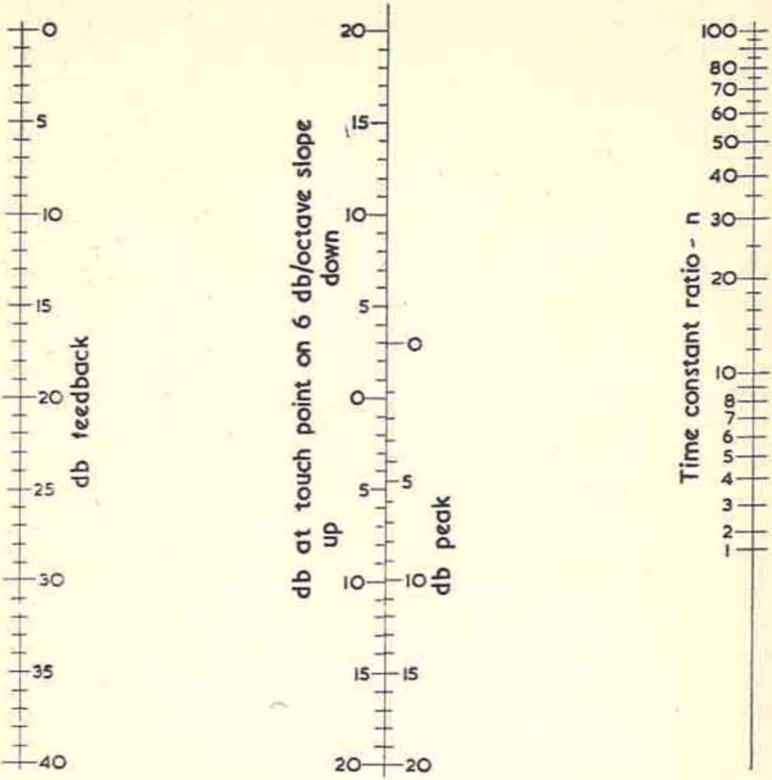

FIG. 31. CHART FOR CALCULATION OF RESPONSE OF ANY COMBINATION OF TWO CUT-OFFS, WITH NEGATIVE FEEDBACK

28, in stages, this gives 160 K. The gain of the 6J5 with 47 K. load is A = 17. B is $\frac{.16}{1.16} = .138$. So AB = 17 × .138 = 2.35. 1 + AB = 3.35, corresponding to 10½ db feedback. The l.f. end consists of a cut-off caused by C_1 with a little over 1 M., and a step caused by the 6J5 cathode decoupling, which will have negligible effect. Feedback will extend the range at the l.f. end in a manner similar to that shown at Figure 17 for the h.f. end. At the h.f. end, assuming that the total capacitance to earth, as from the 6J7 anode with 6J5 grid, and from the 6J5 anode, is of the same order, their time constants will be in similar ratio to their anode loads, between 4/1 and 5/1. Reference to the abac of Figure 31 shows that 10½ db feedback will not cause peaking.

The feedback could be made variable, within limits, by a circuit of the type shown at Figure 36.

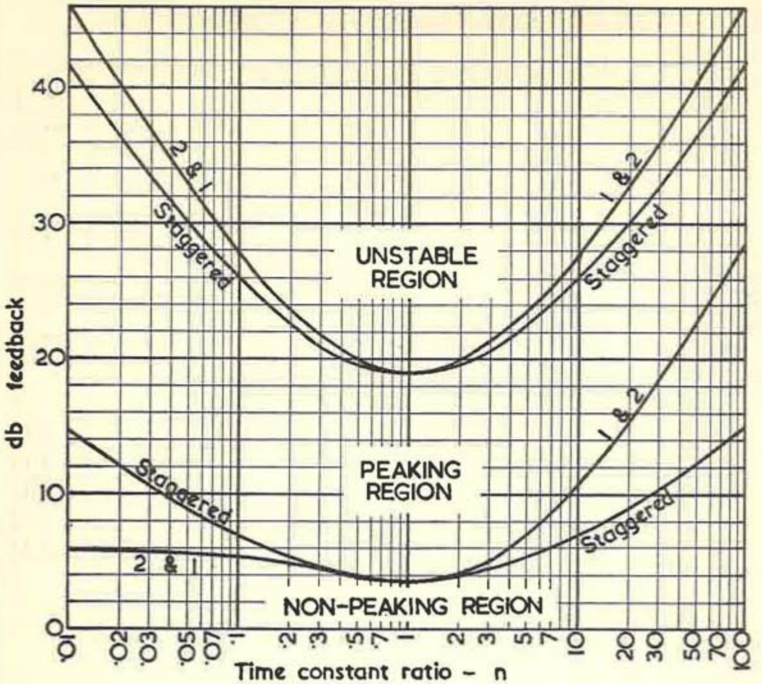

FIG. 32. LIMIT CHART TO HELP TO FIND SUITABLE VALUES USING THREE CUT-OFFS

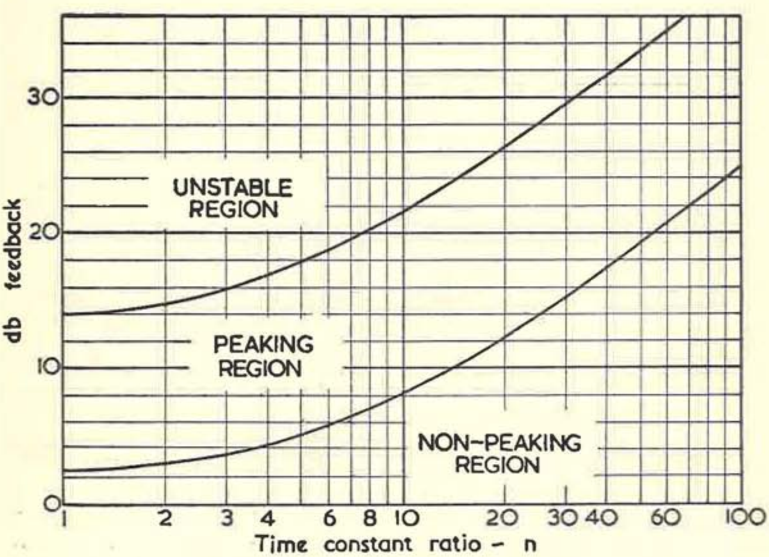

FIG. 33. LIMIT CHART FOR FOUR CUT-OFFS

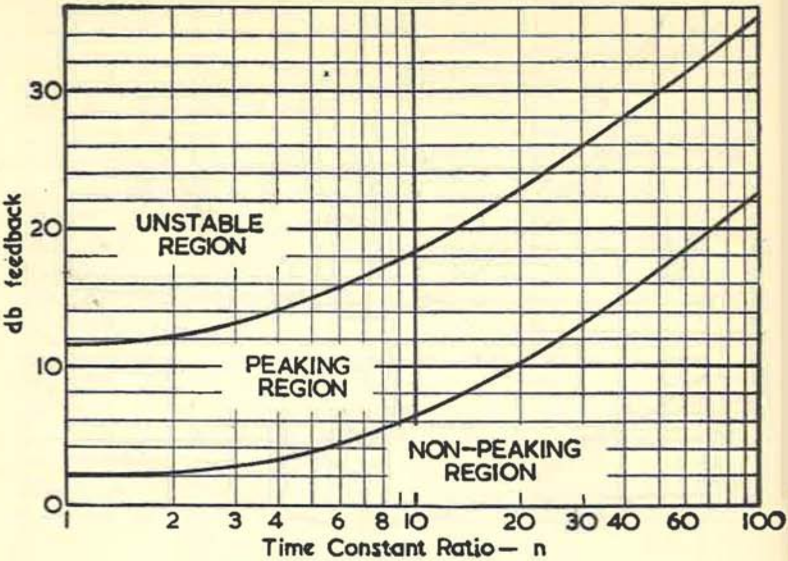

FIG. 34. LIMIT CHART FOR FIVE CUT-OFFS

Feedback and Volume Controls

An important feature to remember in working out feedback circuits is that *a volume control should never be included in the amplifying section over which negative feedback is applied.*

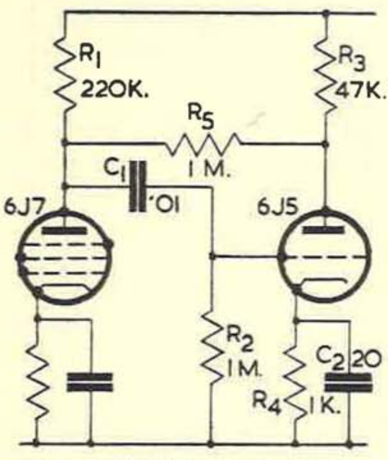

FIG. 35. THE RESISTOR FROM ANODE TO ANODE, R_5, PROVIDES NEGATIVE FEEDBACK FOR THE 6J5

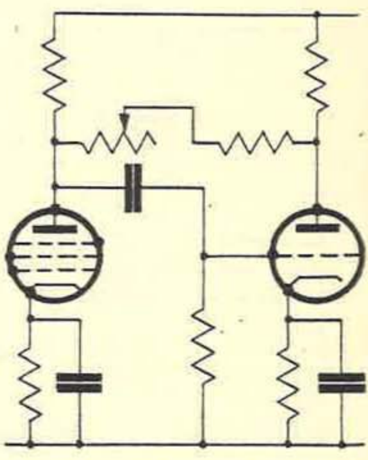

FIG. 36. METHOD OF MAKING THE FEEDBACK OF FIG. 35 VARIABLE

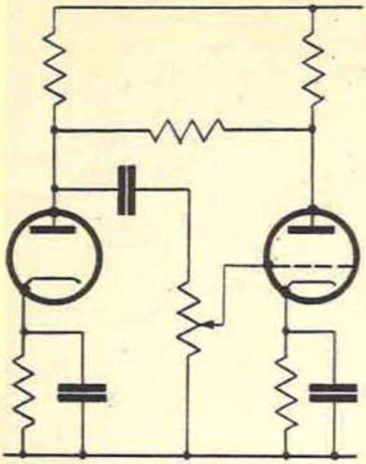

FIG. 37. A VOLUME CONTROL SHOULD NOT BE INCLUDED IN THE FORWARD PART OF AN AMPLIFIER WHERE NEGATIVE FEEDBACK IS USED

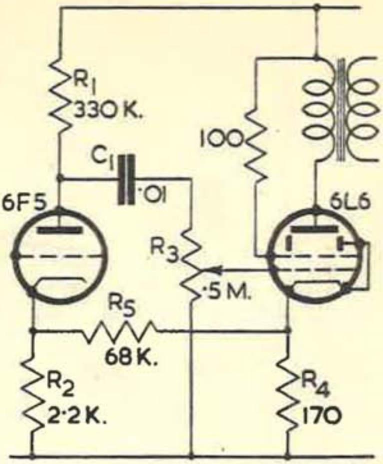

FIG. 38. COMBINED POSITIVE AND NEGATIVE FEEDBACK ARE USED TO INCREASE THE RANGE OF CONTROL PROVIDED BY R_3

Figure 37 shows a modified version of the circuit of Figure 35 to illustrate this. When the volume control is full on, there is plenty of negative feedback. As the control is turned down, feedback is reduced. When the fed-back signal at the anode of the first stage is equal to the original signal, there is 6 db negative feedback. Turning the volume control down further, a stage is reached where the signal due to the two valves is equal at the anode of the second valve. This will mean that the output cancels ; but any distortion caused in the second stage will not cancel. Below a certain point, therefore, use of feedback in this way increases distortion, instead of reducing it. This is true of any circuit where a volume control is used in an amplifier over which negative feedback is applied. The volume control should be put in a place where it is not included in a feedback loop.

Combining Positive and Negative Feedback

Sometimes it is advantageous to use positive feedback as well as negative feedback. Two examples of this type are included, complete with values, as they provide good application for the principles of calculation given in this book.

Figure 38 shows a circuit where combined feedback is used to increase the effective " range " of the volume control R_3. This usage does not violate the rule given in the previous section, because the volume control is in the part of the amplifier using positive feedback. Negative feedback is caused by the cathode resistor of the 6F5, to the extent of AB = 2.5 or 11 db. The gain of the

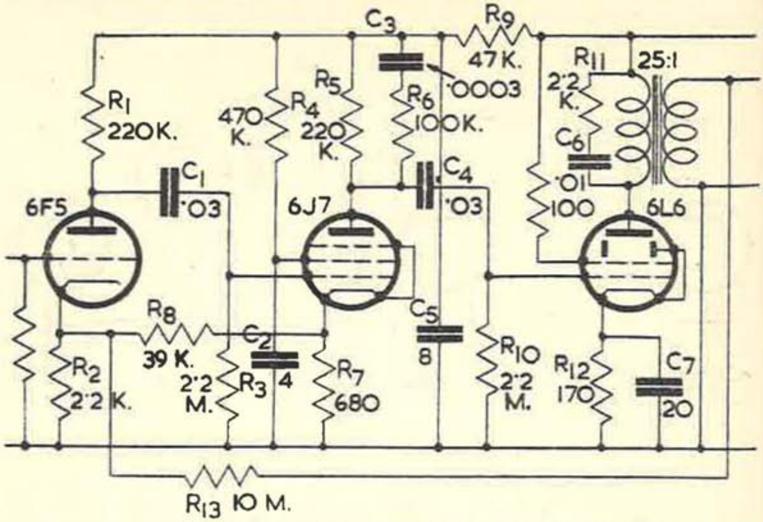

FIG. 39. POSITIVE FEEDBACK BY R_8 INCREASES GAIN FOR THE AMPLIFIER USING R_{13} FOR NEGATIVE FEEDBACK

6F5 without feedback is 75, and the gain of the 6L6, from grid to cathode, across resistor R_4 is about unity, assuming the anode load to be not greater than the optimum load of 2500 Ohms. When the volume control is full up, positive feedback is made about equal to the negative feedback so as to neutralise it. A is 75, and B is

$$\frac{2.2}{70.2} = .0313, \text{ so } AB = 75 \times .0313 = 2.35,$$ which is as near as

standard tolerance components will allow. When the volume control is nearly off, there is no positive feedback, so the 11 db negative feedback increases the contrast given by the volume control by about that much. The l.f. and h.f. response in the negative feedback loop contains only one stage, the 6J5, in which there is no coupling capacitor, and shunt capacitance has negligible effect. The effect of positive feedback is the reverse of negative feedback, so the two l.f. cut-offs, due to C_1 and the transformer primary inductance, will combine to produce a cut-off that starts earlier and more gradually when feedback is applied, *i.e.* as the volume control is turned up. Similarly with the h.f. cut-offs, due to capacitances to earth. It is seen that turning up the volume control increases the gain and slightly reduces the frequency range.

Figure 39 gives a complete circuit, where positive feedback is used to increase the gain of the amplifier before negative feedback is applied. This helps to offset the loss of gain caused by negative feedback. Positive feedback is used where distortion is low,

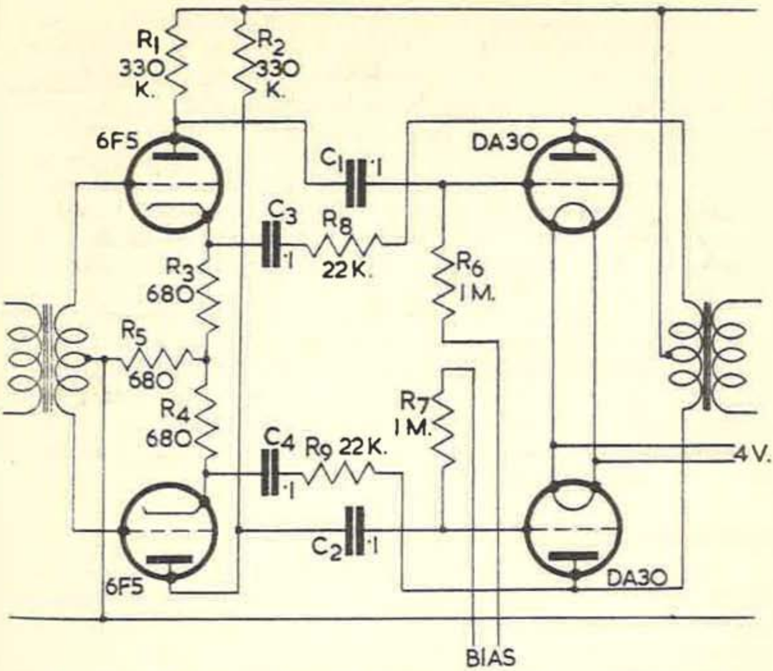

FIG. 40. IN THIS CIRCUIT, NEGATIVE FEEDBACK PROVIDES " SELF-BALANCING "
(see *Self-balancing Push-Pull Stages*, page 62)

due to signal being small. The primary object of negative feedback is to reduce the effective a.c. resistance of the 6L6 to about 500 Ohms. The gain of the 6F5 without feedback is 75. It has negative feedback of its own, of AB = 2.5. The gain to the 6J7 cathode (without feedback) is 56 (the gain of the 6J7 being .75), so R_8 gives positive feedback of AB = 3. Subtracting the negative of 2.5 leaves a net positive AB of .5, which is equivalent to 6 db positive feedback, so the effective gain of the 6F5 is raised to 150. Its a.c. resistance will be raised by 2/1 ratio as well. This must be remembered in calculating the overall cut-off performance. The gain of the 6J7 to its anode is 250, and the theoretical gain of the 6L6 with open circuit load is 135, so the total voltage gain, over the output transformer is, $150 \times 250 \times 135 \div 25 = 202{,}500$. To meet the required a.c. resistance reduction, the overall AB product must be 50, so B needs to be about $\frac{1}{4000}$. R_{13} of 10 Meg gives about $\frac{1}{4500}$, which is nearest to what standard values can give. The overall gain is approximately 4500, so that to give

5 volts at the loudspeaker, a little over 1 millivolt input is required.

Factors contributing to the l.f. response are : Coupling capacitors C_1 and C_4, primary inductance of the output transformer, 6J7 screen decoupling, 6L6 cathode decoupling, and the h.t. supply decoupling, C_5, R_9. The basic arrangement is the combination of the coupling capacitors C_1 and C_4, each giving a time constant greater than 70 m-sec, with the transformer primary inductance of 4 Henrys giving a time constant of 180 μ-sec. Any slight tendency to peak is eliminated by the step (6 db) introduced by R_{12}, C_7. The decoupling components, C_2, R_4 and C_5, R_9 each give time constants well above those of C_1 and C_4.

The h.f. response is taken care of by the two inter-stage capacitances to earth, and the impedance correction components C_6, R_{11} used to ensure correct loading of the 6L6 with loudspeaker load, so making inherent distortion lower. This arrangement causes considerable peak, and, under some conditions, (due to different loudspeakers) instability. The step circuit, consisting of C_3, R_6 overcomes this.

Self-balancing Push-Pull Stages

Figure 40 shows a circuit where feedback is used to give about 17 db straight feedback, and the common resistor, R_5, raises the feedback for balancing to nearly 26 db. This means that feedback will reduce a 20% out-of-balance to about 1%. An important feature of this circuit is the accuracy of balance in components R_3, R_4, R_8 and R_9, as balance depends upon these values rather than the " matching " of the valves used.

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