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"To promote the general advancement of and to facilitate the exchange of information and ideas on Radio Science."

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FESTIVAL OF BRITAIN 1951

As part of the Institution's Silver Jubilee Celebrations, it was intended to hold a Convention this year on similar lines to that which took place in Bournemouth in 1947. Eighteen months ago, however, the Government announced plans for holding a Festival similar to the Great Exhibition of 1851. Immediately after this statement, the Institution suggested to the Festival organizers that it would be in the national interest if such Conventions, as that held by the Institution, coincided with the Festival of Britain—a suggestion which has recently received a great deal of support.

The Institution was invited to a science conference held by the Festival Committee on January 26th, when Sir Edward Appleton, outlined the part which science and technology will play in the Festival. It was then announced that there will be three Exhibitions :—

(1) On the South Bank of the Thames, London, where the practical consequences of pure science will be illustrated.

(2) At the Science Exhibition in South Kensington, where the emphasis will be on the revolutions in human thought which have resulted from man's scientific curiosity concerning the ultimate nature of matter.

(3) At the Exhibition of Industrial Power in the Kelvin Hall, Glasgow, where scientific knowledge and techniques are illustrated in their promotion of industrial and engineering progress.

The Exhibition of the South Bank, London, will include in the Transport Section a prominent display emphazising British advances in telecommunications (radio, radar, etc.) which have important applications for transport control. British contributions to broadcasting, television and cables will be strongly represented. A small pavilion dealing with cinematography and television will introduce the Telecinema itself, where both television and films will be shown. The Telecinema, which will seat about 400 people, has been designed on entirely new principles incorporating the latest techniques in projection and acoustics.

Radio engineers will also have particular interest in the Dome of Discovery, where the achievements of such men as Faraday, Thomson and Rutherford will be displayed. The basic theme is that British initiative in discovery and exploration continues with all its old force but nowadays with far more effective tools and methods. Astronomy, for example, is now aided by the use of radio techniques, as will be demonstrated by the "radar telescope."

Sir Edward Appleton stated :

"Now there is another and altogether different aspect of the Festival, but one that is by no means less important than the officially arranged programme. I am thinking of the part that can be played by the learned societies and scientific institutions who can very appropriately make 1951 a special occasion for the display and assessment of British achievement in their own chosen fields.

"In this matter, a number of learned bodies have already taken action. I know that the Council of the British Association, for example, intends that its 1951 meeting should do all it can to further the objects of the Festival. It is just such activities as these that are needed to complete the scientific contribution to the Festival, and I sincerely hope that before long we shall be hearing of many more additions to this programme of special sessions arranged by the learned bodies which for so long have formed an essential part of our way of life in Britain."

The Council of the Institution has already made known its intention of assisting by holding the next Convention in 1951, in London. As far as possible, co-operation will also be given to other Institutions whose meetings may be of interest to the radio engineer, by agreeing on dates not likely to conflict.

EDUCATION COMMITTEE REPRESENTATIVES

George Austin Taylor was born in Wallasey, Cheshire, in October, 1903. He received his early technical education at the Croydon Polytechnic and his radio training with the Radio Communication Company. Towards the end of the first world war he saw service in the Royal Engineers (Signals) and subsequently with the Indian Posts and Telegraphs Department.

Returning to this country in 1929, Mr. Taylor was for a time in the test laboratory of Messrs. Ericsson Telephones of Beeston, Nottingham. In 1935 he joined the laboratory staff of Philips Electrical Ltd., at Mitcham, and in 1943 he was appointed Training Officer, responsible for staff and works training. He is now Group Education and Training Officer for the Philips Group of Companies in this country, covering training for all grades of staff.

Mr. Taylor was elected an Associate Member of the Institution in 1944 and was appointed a member of the Education Committee in 1945, of which he was the chairman during 1946. He was examiner in radio technology for the Institution during 1945 and 1946 and was elected a member of the General Council for the period 1946 to 1948.

During that time he was also serving on a number of external Committees concerned with education and was a member of the Surrey Mid-Eastern Divisional Executive which was set up to implement the schemes of education under the Education Act, 1944.

Transferred to full Membership of the Institution in March, 1948, Mr. Taylor is now a member of the Professional Purposes and Finance Committees, as well as representing the Institution on the Radio Trades Examination Board and on the Radio Service Work Advisory Committee of the City and Guilds of London Institute.

Erich Ernst Zepler was born in Herford, Germany, in January 1898. He studied at the Universities of Bonn and Berlin and obtained his D.Phil. at the University of Wurzburg in 1923. Subsequently, he continued his research work at the Physical Institute, Wurzburg.

In 1925, Dr. Zepler joined the staff of Telefunken (Berlin) and was later appointed head of the receiver development laboratories. Eventually, he held the position of head of the firm's research laboratories until, in 1935, he left to join the Marconi's Wireless Telegraph Company at Chelmsford. He remained with the Marconi Company until 1940, when he concentrated on the preparation of his book "The Technique of Radio Receiver Design." This is considered a standard work of reference and has now reached the fourth impression.

From 1941 to 1943, Dr. Zepler lectured at University College, Southampton. After a further three years as a lecturer at the Cavendish Laboratories, Cambridge University, he returned to University College, Southampton, and a year later was appointed head of the Department of Electronics, Telecommunications and Radio Engineering. A Chair in Electronics has recently been created at the University and Dr. Zepler has been appointed the first Professor.

Prof. Zepler is an outstanding chess player, having represented Essex and Hants in tournaments. He is also internationally known as a composer of chess problems.

Elected a full Member of the Institution in 1944, Prof. Zepler has been the examiner for the Radio Reception paper of the Graduateship Examination since 1945. He has served on the Education and Examinations Committee since 1945 and was elected its Chairman in 1947. In addition he is the Institution's representative on the Telecommunication Advisory Committee of the City and Guilds of London Institute.

AUDIO FREQUENCY SELECTIVE AMPLIFIERS*

by

S. W. Punnett, B.Sc.†

SUMMARY

Various types of selective audio frequency amplifiers are considered. Particular attention is paid to the problem of varying the frequency continuously. This leads to a consideration of the effects produced by the fact that two or more components in the frequency selective network must be varied simultaneously. Since the components will not stay exactly in step as they are varied, the gain, Q-factor and bandwidth of the amplifier will change with tuning.

1.0. Introduction

The use of feedback to produce highly selective audio frequency amplifiers is now well known, but little information exists as to the relative merits of the various circuits employed particularly with reference to constancy of the amplifier gain and band-width. This paper will consider various types of circuit which may be employed to produce frequency selectivity, comparing their relative merits with respect to simplicity, ease of adjustment, constancy of gain, Q-factor and band-width. The effect of errors in the components of the selective network will receive particular attention, as these errors are the limiting factors controlling the maximum value of the Q-factor that may be obtained if reasonable constancy is also of importance. In fact if these errors are too large they can cause self oscillation in any of the circuits. The first network dealt with will be that employed in R-C audio frequency oscillators, and various circuits using this network will be considered. The twin-T network will then be dealt with and finally the Wien bridge.

2.0. Simple Selective Network

The network employed in R-C oscillators for audio frequencies suggests one network which might be used to produce a selective amplifier. It will be necessary to make some changes. First, the gain of the amplifier used with the network must be reduced to a point below that required for oscillation. Secondly, arrangements must be made to apply the input signals to the circuit. The action of the selective network shown in Fig. 1 is considered first. Initially equality of resistors and of capacitors will be assumed and the general case will be dealt with later. If a voltage V_o is applied to the terminals 1 and 2 then the voltage appearing between terminals 3 and 4 is given by the following expression :—

$$v = V_o \cdot \frac{\dfrac{\dfrac{R}{j\omega C}}{R + \dfrac{1}{j\omega C}}}{R + \dfrac{1}{j\omega C} + \dfrac{\dfrac{R}{j\omega C}}{R + \dfrac{1}{j\omega C}}}$$

$$= V_o \cdot \frac{1}{1 + \left(1 + \dfrac{1}{j\omega CR}\right)(1 + j\omega CR)}$$

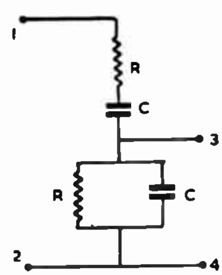

Fig. 1.—Simple selective network such as is used in certain R-C oscillators.

To obtain expressions which can be handled more easily, the following substitutions will be employed throughout this paper :—

$$\omega_o = \frac{1}{CR}, \ \alpha = \frac{\omega}{\omega_o}, \ y = \left(\alpha - \frac{1}{\alpha}\right)$$

Inserting these in the above expression for the output voltage v results in the following

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$$v = V_o \cdot \frac{1}{3 + j\left(\alpha - \dfrac{1}{\alpha}\right)} = V_o \cdot \frac{1}{3 + jy}$$

Thus if V_o is the output voltage of an amplifier and if v is used as the feedback voltage, the value of the quantity usually designated by β will be $1/(3 + jy)$. The manner in which the magnitude and phase of $1/(3 + jy)$ varies with y is shown in Figs. 2 and 3 respectively. From

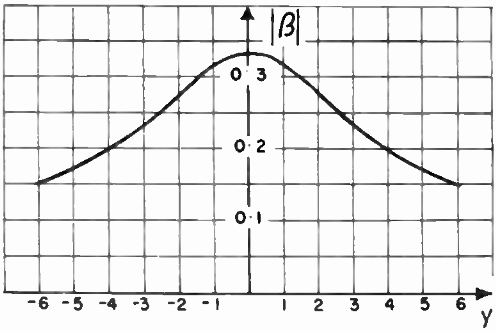


Fig. 2.—Variation of the magnitude of the feedback with y.

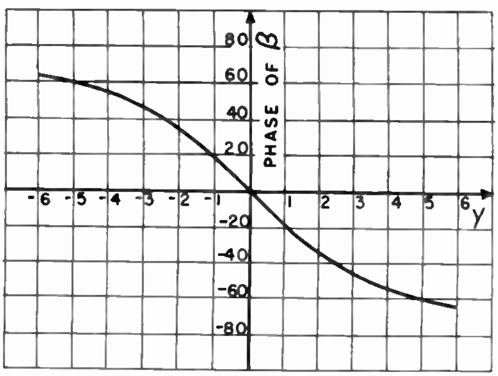


Fig. 3.—Variation of the phase of the feedback with y.

the form of the curve shown in Fig. 2 it is obvious that, if this network is to be employed, to obtain a selective amplifier the feedback voltage must be applied to the input in phase with the voltage already existing at this point when the frequency of the applied signal has a value given by $f_o = \omega_o/2\pi$. This frequency f_o will be referred to as the resonant frequency of the circuit as a whole. Thus, on resonance, the feedback is positive. Since on resonance the magnitude of v is $V_o/3$, the gain of the amplifier must be less than 3 if oscillation is not to take place. A block diagram of the suggested amplifier is shown in Fig. 4, the amplification of the amplifier proper is reduced by means of negative feedback β^- from M to a value A which is little less than 3. Since the value of the M may be very large, A will depend only on the value of β^- and changes in the valves and voltages will have little or no effect. Furthermore the phase shift occurring in the amplifier will be very small and the output may be assumed in phase with the input so that A is real and positive. Hence the gain G of the circuit as a whole is given by

$$G = \frac{A}{1 - \dfrac{A}{3 + jy}}$$

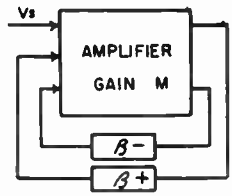


Fig. 4.—Block diagram of a selective amplifier using the network of Fig. 1.

The variation of the magnitude of G with y, and hence indirectly with frequency, is shown in Fig. 5. The response curve of the amplifier is somewhat similar to that of a tuned circuit, but there are some marked differences. What might be termed the Q-factor of the circuit depends on the manner in which it is derived. In this paper the Q-factor is derived from the variation of the response near resonance, although another value could be obtained from a consideration of the protection afforded by the circuit against harmonic and sub-harmonic frequencies. It is to be noted that on resonance the gain $G_o = A/(1 - A/3)$, and hence the effect of the negative feedback is largely counteracted by the positive feedback due to the selective network. The magnitude of the resultant negative feedback is $(\beta^- - 1/3)$, and hence the

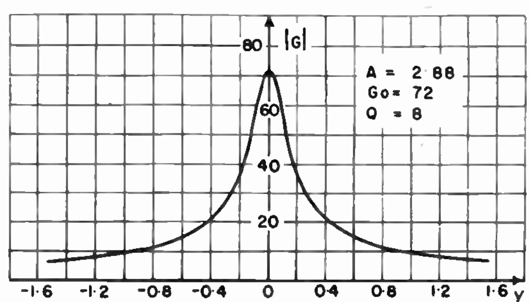


Fig. 5.—Response curve of the amplifier of Fig. 4.

stability of the circuit as a whole is less than that indicated by the magnitude of the negative feedback β^-. In fact, if M is less than G_o, the stability may actually be degraded.

In terms of the negative feedback β^- and the amplification M of the amplifier the value of A is given by :—

$$A = \frac{M}{1 + \beta^- . M}$$

If this value of A is substituted in the expression for G_o, the following expression is obtained :—

$$G_o = \frac{M}{1 + M\left(\beta^- - \frac{1}{3}\right)}$$

If this expression is differentiated with respect to M and the result rearranged, the following equation giving the fractional variation of G_o in terms of the fractional variation of M is obtained :—

$$\frac{dG_o}{G_o} = \frac{dM}{M} \cdot \frac{1}{1 + M\left(\beta^- - \frac{1}{3}\right)}$$

From this equation it can be seen that if the percentage variation in G_o is to be kept to ± 1 per cent. and if the variation in the amplifier gain M due to variations in valves, voltages and so on is ± 20 per cent., then the value of $\left\{1 + M\left(\beta^- - \frac{1}{3}\right)\right\}$ must be of the order of 20. Now it is shown below that to obtain a Q-factor of the order of 20 the value of G_o, which must be employed, is 180. Thus :—

$$180 = \frac{M}{1 + M\left(\beta^- - \frac{1}{3}\right)}$$

The denominator of the last expression must equal 20 from the above consideration of the stability of the resultant selective amplifier. Thus the value of M required must be of the order of 3,600 and the value of β^- M must be approximately 1,200, since the gain A must be less than 3 ; A will be very nearly equal to $1/\beta^-$ with such heavy negative feedback. Since at least two stages will be required to give the necessary value of M, and since the value of the feedback factor is so large, the amplifier will need careful design to avoid self-oscillation at both very low and very high frequencies where the phase shift in the amplifier has altered by 180 degrees. The magnitude of β^- M must fall below unity before these frequencies are reached. This is not too difficult with two stages but in the case of three stages the problem becomes very complicated. This constitutes one of the main disadvantages of this type of circuit.

3.0. Response Near Resonance

For a simple tuned circuit the response near resonance in terms of the maximum response S_o and the Q-factor of the circuit is :—

$$S = \frac{S_o}{1 + jyQ}$$

This expression only holds for small values of y where the lack of symmetry may be ignored. In the case of the selective amplifier the gain G may be written :—

$$G = \frac{A(9 + y^2)}{(9 + y^2 - 3A) + jyA}$$

$$= \frac{A}{\left(1 - \frac{3A}{9 + y^2}\right) + jy \cdot \frac{A}{9 + y^2}}$$

Hence, close to resonance, where the value of y is small this approximates to :—

$$G = \frac{A}{\left(1 - \frac{A}{3}\right) + jy \cdot \frac{A}{9}} = \frac{G_o}{1 + jy \cdot \frac{G_o}{9}}$$

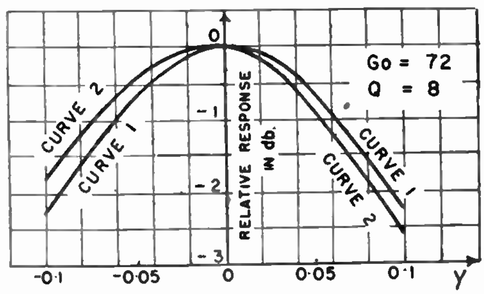


Fig. 6.—Curve 1 shows the relative response of the amplifier near resonance while curve 2 shows that of a series tuned circuit of the same Q as the amplifier.

In place of $A/(1 - A/3)$ the value of the gain G_o on resonance has been substituted. Comparing this equation with that for the simple tuned circuit it is seen that the effective Q-factor is $G_o/9$. Curves 1 and 2 of Fig. 6 show respectively the response of the selective amplifier and of a simple tuned circuit having the same effective Q. These are plotted for small values of y. Curves 1 and 2 of Fig. 7 show respectively the responses in each case for large values of y; it is to be noted that, while the response of the simple tuned circuit is very asymmetric, that of the selective amplifier about the resonance point is symmetrical. This is seen analytically in the next section.

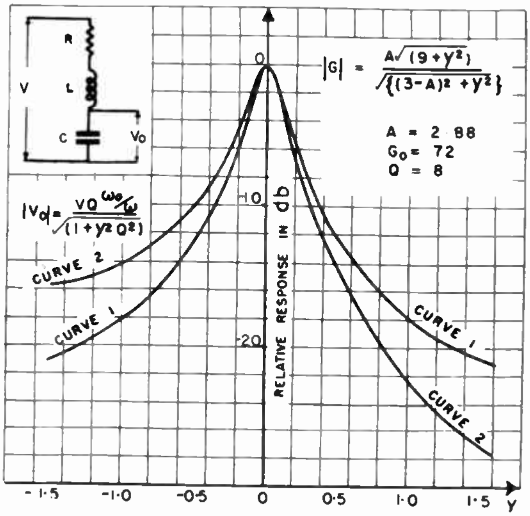


Fig. 7.—The curves are the same as in Fig. 7 but for large values of y.

4.0. Response For Large Mistunings. (Harmonic Protection)

At resonance the gain of the circuit is given by $G_o = 3A/(3 - A)$. At $2f_o$ and at $f_o/2$ the gains are respectively

$$\frac{A\,(3 + j\,1{\cdot}5)}{(A - 3) + j\,1{\cdot}5} \text{ and } \frac{A\,(3 - j\,1{\cdot}5)}{(A - 3) - j\,1{\cdot}5}$$

In either case the magnitude of the gain is,

$$\frac{A\sqrt{11{\cdot}25}}{\sqrt{(A - 3)^2 + 1{\cdot}5^2}}$$

To secure a reasonable harmonic protection, that is a large effective Q, the value of A must be nearly equal to 3, the problem of stability being the factor which limits the closeness of approach to this figure. As will be seen later, the more closely the value of A approaches the limiting value 3, the smaller are the permissible tolerances on the components used in the selective network. Hence the factor $(A - 3)$ can be ignored in the above expression for the gain, and at either $2f_o$ or $f_o/2$ the gain tends to the value $A\sqrt{5}$. Thus the ratio of the gain at the fundamental frequency to that at $2f_o$ or $f_o/2$ is $G_o/A\sqrt{5}$; since A is approximately 3 this has the value $G_o/3\sqrt{5}$. It can easily be shown that, in the case of a simple tuned circuit, this ratio is 3Q in the case of $2f_o$, and 3Q/4 for $f_o/2$. These are more easily compared with the value for the selective amplifier if G_o is replaced by 9Q, where Q is the effective Q-factor found for small values of y. Thus the ratio of gain at the fundamental frequency f_o to that at $2f_o$ or $f_o/2$ is $3Q/\sqrt{5}$. This is smaller than for the simple tuned circuit in the case of $2f_o$ but is greater in the case of $f_o/2$.

5.0. Input Circuit Arrangements

The method by which the various voltages are applied to the input of the amplifier is of importance; not only may it seriously affect the stability but it also alters the exact form of the response curve obtained. Fig. 8 shows a simple method of arranging the various inputs; with the circuit shown it would not be possible to make M very large, and hence the circuit cannot be expected to give a very large Q-factor if the stability is to be reasonable. The resistors R_1 and R_2 produce the necessary negative feedback.

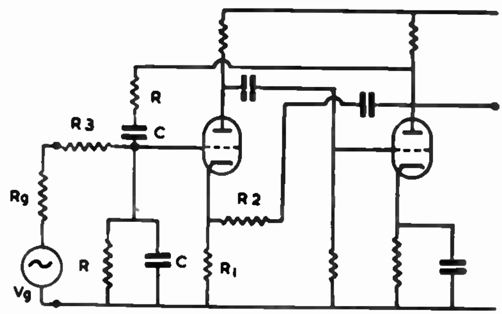


Fig. 8.—Input circuit of the selective amplifier when signal source is of low impedance.

The resistor R_3 must be large in comparison with R, since, in series with the internal resistance of the signal source, it is shunted across one of the selective network resistors R. The effect of a relatively small value of R_3, say R_3 equal to 10R, on the tuning can easily be compensated by slightly increasing the parallel resistor R. This compensation cannot be perfect at all values when resistive tuning is employed. Furthermore this is not the only effect of the resistor R_3; signals whose frequencies are lower than f_o are unduly emphasized; the response curve is no longer symmetrical. At frequencies well below f_o the reactances of the capacitors C are very large and their effect can be ignored. Thus the network at the input reduces to a simple resistive potential divider network formed by R_3 and R. The gain at very low frequencies thus approximates to the constant value $AR/(R + R_3)$, and at very high frequencies it approaches zero.

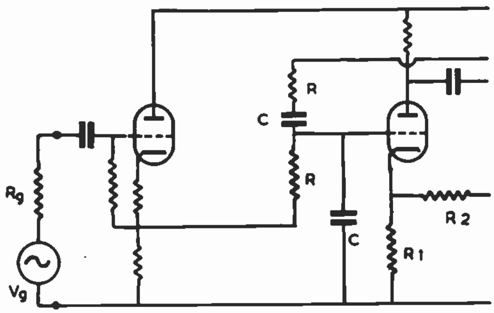


Fig. 9.—Cathode follower input for use with a signal source of high impedance.

If the source of the signal voltage has a low output impedance then it may be connected between the lower end of the parallel resistor R and earth. As this is not the case the circuit arrangement of Fig. 9 can be employed. The cathode follower has a very low output impedance. In consequence, the effect on R is negligible, and the response curve is now almost exactly symmetrical. Furthermore, by proper choice of the load resistance for the cathode follower, the D.C. conditions of the first valve in the amplifier proper can be adjusted at will.

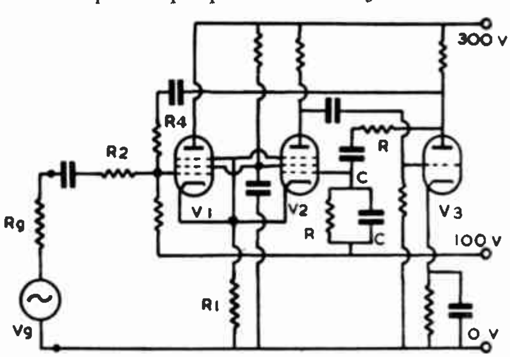


Fig. 10.—Selective amplifier embodying a cathode-coupled pair so that the selective network is isolated from the input circuit. The grid resistor of V_1 is R_3.

Thus the size of R_1 can be increased without the first valve working with too small a current. The shunting effect of R_1 and R_2 on the second valve is decreased and the gain M can be made larger with a consequent increase in the stability of the circuit as a whole. The input impedance of the circuit is also very large, which is an advantage.

The circuit of Fig. 9 suggests a modification which would appear to confer several advantages; this modification is shown in Fig. 10. The first two valves are now arranged as a cathode coupled amplifier; the resistor R_1 being large, signals applied to one grid have almost equal but opposite effects on the two valves. If required the resistor R_1 may be replaced by a pentode so that the A.C. resistance in the common cathode lead is of the order of a megohm. If now a signal is applied to one grid while the other is earthed the anode voltages are almost exactly equal and opposite in phase. The negative feedback is controlled by the resistors R_2, R_3 and R_4 (the value of R_2 must be such that it masks the impedance of the signal source) or alternatively the input may be fed into the circuit via another valve so that the impedance presented to the circuit is constant.

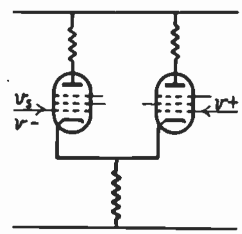


Fig. 11.—Skeleton circuit of first two valves of Fig. 10 showing the signal voltage v_s, the negative feedback v-, and the positive feedback v+.

The positive feedback, which comes from the same point as the negative feedback, is applied to the grid of the second valve. This would appear to be a very satisfatory arrangement since the gain of each valve of the coupled pair is now half the value for each working separately into the given anode load. This is very much higher than in the circuits shown in Figs. 8 and 9, because the presence of the large resistor in the cathode lead of the first valve of the amplifier proper reduces the amplification of this valve very considerably in these circuits. Thus in this new circuit the value of M can be much larger with a resultant increase in stability. The increase in stability is illusory since the first stage is itself very unstable when used in this fashion. The skeleton circuit of the cathode coupled pair shown in Fig. 11 is considered. The signal voltage v_s and the negative feedback voltage v- are applied to the grid of the first valve while the positive feedback voltage $v+$ is applied to the grid of the second valve ; note that it is from the anode of the second valve that the signal is passed on to the next stage of the amplifier. If the mutual conductances of the two valves are g_{m1} and g_{m2} respectively and if the anode resistances of the pentodes are assumed infinite, it is easy to show that the value of the anode current i_{a2} of the second valve is :—

$$i_{a2} = \frac{-g_{m1}g_{m2}v_s}{\frac{1}{R_1} + g_{m1} + g_{m2}} + \frac{g_{m2}\left\{g_{m1}v- - v+\left(g_{m1} + \frac{1}{R_1}\right)\right\}}{\frac{1}{R_1} + g_{m1} + g_{m2}}$$

The second term is proportional to the resultant feedback due to the difference between the positive and negative feedbacks. Now suppose that the two valves are interchanged (this at least leaves the D.C. conditions unaltered, which might not be the case if one of the valves were replaced by a new valve). The first term in the above expression is not changed, but the second becomes :—

$$\frac{g_{m1}\left\{g_{m2}v- - v+\left(g_{m2} + \frac{1}{R_1}\right)\right\}}{\frac{1}{R_1} + g_{m1} + g_{m2}}$$

As will be seen in the numerical example below, this alteration can be quite considerable even when the resistor R_1 is very large.

It is assumed that the circuit has been adjusted to a Q value of about 11 so that the gain G_o on resonance is 100 ; then the value of the amplification A is 300/103. Now if the amplification M is large, the factor β^- is almost exactly equal to 1/A, and hence $\beta^- = 1{\cdot}03/3$. Thus, when the circuit is tuned to resonance, v^+ is equal to $V_o/3$ and v^- to $1{\cdot}03V_o/3$, where V_o is the output voltage of the amplifier. Finally let the mutual conductances of the two valves be 1.10^{-3} and 2.10^{-3} mhos respectively, and the value of R_1 be 50.10^3 ohms.

Valves in normal order.

Term representing feedback is,

$$V_o . \frac{2 . 10^{-5}}{9{\cdot}06}$$

Valves reversed.

Term representing feedback is,

$$V_o . \frac{4 . 10^{-5}}{9}$$

Thus in this case there is over 100 per cent. change in the term representing the feedback, and hence a very considerable change in the value of the gain on resonance and in the Q-factor and band-width of the circuit. Thus changes in the mutual conductances of the valves due to age or to replacement will cause very undesirable changes in the characteristics of the selective amplifier. With a higher value of the Q-factor and hence of G_o it may be sufficient to cause the circuit to go into oscillation.

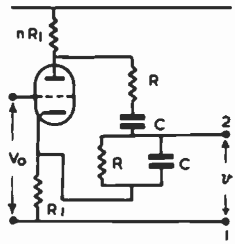


Fig. 12.—Simple selective network connected between anode and cathode of a phase-splitting valve.

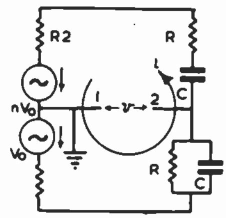


Fig. 13. — Equivalent electrical circuit of Fig. 12.

6.0. Simple Selective Network Plus Phase Splitter Valve

The circuit shown in Fig. 12 is considered. The output from the amplifier is applied to the grid of a triode phase splitter valve and the selective network is connected between anode and cathode. The resistors in the selective network will be very much larger than the input impedance at the cathode of the valve so that, as a first approximation, the feedback from the anode to the cathode via this network may be neglected. Thus the approximate equivalent circuit is as shown in Fig. 13.

The resistors R_2 and R_3 are the output impedances at the anode and cathode respectively, but that at the cathode is so small that its effect can be ignored ; R_2, which is of the order of nR_1, can be allowed for by decreasing the size of the series resistor in the selective network by a suitable amount. The current through the anode and cathode resistors is the same, and hence the voltage across the cathode resistor is almost exactly equal in magnitude and phase to

V_o, and the voltage across the anode resistor is exactly n times that at the cathode but opposite in phase. From the equivalent circuit, the current i flowing through the selective network is given by :—

$$i = V_o \cdot \frac{(1+n)(1+j\alpha)}{R(3+jy)}$$

Where the notation is that used in previous formulæ.

Hence the voltage v between terminals 1 and 2 is as follows :—

$$v = V_o - i \cdot \frac{R}{1+j\alpha} = V_o \cdot \frac{(2-n)+jy}{3+jy}$$

If n is made equal to 2, the value of v is $jyV_o/(3+jy)$. If the voltage v is used as feedback the value of β will be $1/(1+3/jy)$. The form of this factor is of a very different nature from that of the previous arrangement where the feedback was equal to $1/(3+jy)$ and therefore it will be examined more closely before considering the exact circuit arrangements.

When the frequency of the applied signal is $f_o = 1/2\pi CR$, the value of y is zero, and thus the value of β is also zero. When the frequency is very high or very low, y is very large indeed and therefore the magnitude of β approaches unity. The manner in which the magnitude of β varies with y is shown in Fig. 14, and the variation of phase is as shown in Fig. 15.

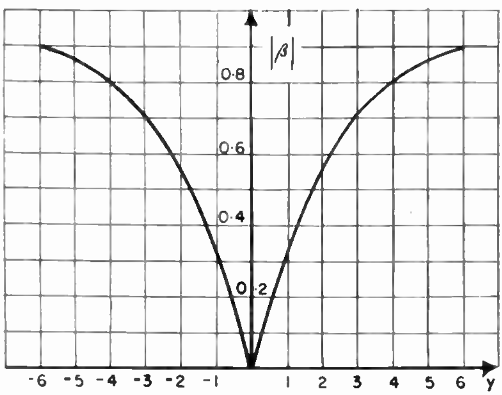


Fig. 14.—Variation with y of the magnitude of the feedback produced by the circuit of Fig. 12 when n = 2.

It is clear that this circuit cannot be used in the same manner as before ; the connections must be so altered that, on resonance, the selective network having zero output, the gain of the amplifier is high, whereas off tune the network applies a large inverse feedback so that the gain of the amplifier falls sharply. Thus if this circuit is employed with an amplifier of gain G_o, the gain of the circuit as a whole will become :—

$$G = \frac{G_o}{1 + \dfrac{G_o}{1 + \dfrac{3}{jy}}} = \frac{G_o\left(1 + j\dfrac{y}{3}\right)}{1 + jy\left(\dfrac{G_o + 1}{3}\right)}$$

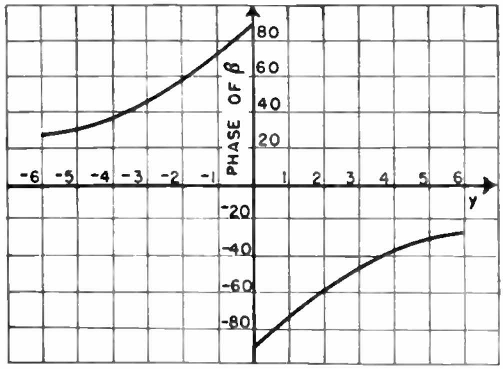


Fig. 15.—Variation with y of the phase of the feedback produced by the circuit of Fig. 12.

The manner in which the magnitude of the gain varies with frequency is shown in Fig. 16 ; it is of the same form as in the previous case. It is noted, however, that the shape of the β curve is not merely that of the previous case inverted.

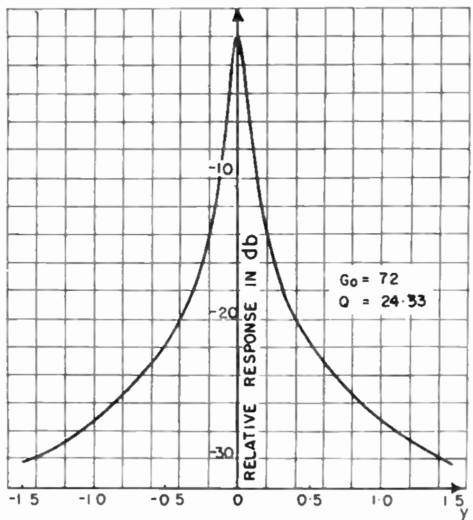


Fig. 16.—Response curve of an amplifier employing the circuit of Fig. 12.

This is seen most clearly from a consideration of the slope of the $|\beta|$ with y curve. If the magnitude of β is differentiated with respect to y the following is obtained :—

$$\frac{d|\beta|}{dy} = \pm \frac{9}{(9 + y^2)^{\frac{3}{2}}},$$

where the positive sign is taken when y is positive and the negative when y is negative. This equation shows that, at the resonant frequency f_o, when y is zero, the slope of the curve is $\pm 1/3$, whereas in the previous case it is easily shown that when y is zero the rate of change of the magnitude of the feedback with y is zero. The variation of the gain with y, and thus with frequency, is of the same form in both cases and it can be shown that the value of $d|G|/dy$ is zero when y is zero. This can also be seen in the following way. Previously the gain G was equal to $A(3 + jy)/(3 - A + jy)$, and since $G_o = 3A/(3 - A)$ this may be rearranged to give :—

$$G = \frac{G_o\left(1 + j\frac{y}{3}\right)}{1 + jy \cdot \frac{G_o}{3A}} \triangleq \frac{G_o\left(1 + j\frac{y}{3}\right)}{1 + jy \cdot \frac{G_o}{9}}$$

The last approximation is justified since A is approximately equal to 3 when the circuit is adjusted to give a reasonable Q. With this new arrangement the value of the gain is given by :—

$$G = \frac{G_o\left(1 + j\frac{y}{3}\right)}{1 + jy\left(\frac{G_o + 1}{3}\right)}$$

From these two expressions the new value of the Q-factor can be immediately obtained. In the first case the value of the effective Q-factor was $G_o/9$, so that in the second it will be $(G_o + 1)/3$. Thus, with the same gain on resonance, the effective Q-factor near resonance has been increased approximately three times ; alternatively, with the same amplifier but with increased negative feedback to reduce the gain on resonance the same Q-factor can be obtained with much improved stability.

A suitable circuit employing the phase-splitting arrangement is shown in Fig. 17. A two-stage amplifier is used, but it is important to note that, while at low frequencies there are only two phase-turning networks, at high frequencies there are four apart from the selective network. Thus there is no danger of self-oscillation at low frequencies, but it may occur at high frequencies, and so there is a limit to the gain that may be employed. It is necessary to ensure that the magnitude of the feedback factor, that is the product of the gain and the feedback, falls to less than unity before the frequency is reached at which the phase shift in the amplifier has changed by 180° from its value over the middle range of frequencies. Fortunately the phase-shifts which occur at the cathodes of the phase-splitting valve and at the cathode coupled pair do not become serious until very high frequencies are reached when the gains of the two amplifier valves have decreased very considerably. It is to be noted that in this circuit no special arrangement has been made to provide negative feedback by means of a separate circuit. This is allowed for by making the resistor in the anode lead of the phase-splitter valve have a value other than twice that of the cathode resistor. The effect of this alteration can be seen by re-examining the formula for the voltage between the terminals 1 and 2 of the circuit shown in Fig. 12. This was $v = V_o \cdot \frac{(2 - n) + jy}{3 + jy}$

If the anode resistor of the phase-splitter valve is made a little less than twice the size of the cathode resistor, then when y is equal to zero, that is when the circuit is on tune, there will be a small voltage fed back to the first pair of valves. This, with the connections shown, will be negative feedback and will result in a reduction in the gain. Thus the amplifier is designed to have a gain much greater than G_o and the value

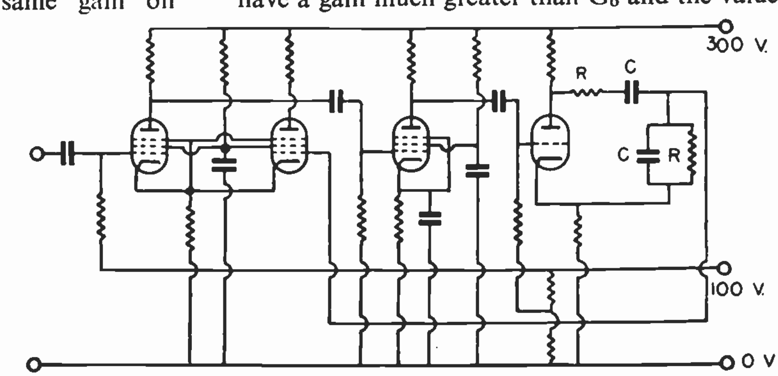


Fig. 17.—Circuit of a selective amplifier using the circuit of Fig. 12 to provide the frequency dependent feedback.

of n is adjusted so that on resonance the gain is reduced to G_0. Therefore the gain of the circuit as a whole is given by :—

$$G = \frac{M}{1 + \dfrac{M\{(2-n)+jy\}}{3+jy}}$$

$$= \frac{M(3+jy)}{\{3+M(2-n)\}+jy(M+1)}$$

M is the amplification of the amplifier in the absence of the selective network. When on tune, so that y is zero, $G_0 = 3M/(3+2M-nM)$, and so if the values of M and G_0 are known the value of n may be calculated. Alternatively, the anode resistor may be made partially adjustable and the value set by experiment so that the required Q-factor is obtained. The effect on the overall characteristic of n being other than 2 is very slight as may be seen by considering the formula for G, which can be arranged in the form :—

$$G = \frac{\left(\dfrac{3M}{3+2M-nM}\right)\left(1+j\dfrac{y}{3}\right)}{1+jy\,.\,\dfrac{(M+1)}{(3+2M-nM)}}$$

$$= \frac{G_0\left(1+j\dfrac{y}{3}\right)}{1+jy\left(\dfrac{G_0+\dfrac{G_0}{M}}{3}\right)}.$$

If this is compared with the previous formula for G it will be noted that the factor G_0/M occurs in the denominator instead of unity, which is a very slight change. The value of n is very close to 2, so if M equals 1,000 and the value of G_0 required is 100 then the value of n is 1·973. The adjustable portion of the anode resistor can be very small, and a few hundred ohms is quite sufficient to provide any reasonable variation in n. Thus it is seen that this circuit is a distinct improvement over the circuits dealt with first ; further the input and output networks are easy to arrange and the danger of self-oscillation is not too great provided too high a value of the amplification M is not aimed at.

7.0. Simple Selective Network with Unequal Components

Up to this point the selective network has been assumed to have equality of resistance and of capacitance since this is easiest to arrange when variable components are employed in order that the selective amplifier can be tuned over a desired range. This is not essential ; it is now possible to obtain very accurate variable resistors which can be ganged so that unequal components can be employed. With variable capacitors this is hardly possible, a ratio of 2 or 3 to 1 being the most that can be hoped for in this case. Further, variable capacitors cannot be as accurately aligned as variable resistors and this is important as will be seen in a later section.

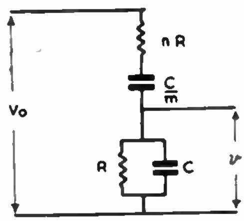


Fig. 18.—General case of the simple selective network.

8.0. Simple Selective Network

The modified form of the circuit is shown in Fig. 18. The series resistor has been replaced by nR and the series capacitor by C/m. If the same method and notation used before are employed again the value of the output voltage v is given by :—

$$v = V_0\,.\,\frac{1}{\left(1+n+\dfrac{m}{n}\right)+j\left(n\alpha-\dfrac{m}{\alpha}\right)}$$

The maximum value of the magnitude of v occurs when $n\alpha_r - m/\alpha_r = 0$, that is $\alpha_r^2 = m/n$. Thus the resonant frequency is no longer f_0 but $f_r = \sqrt{m}/2\pi CR\sqrt{n}$, or $f_r = f_0\sqrt{m}/\sqrt{n}$. At this frequency v is in phase with V_0. The variation with y of the magnitude and phase of v and hence of the feedback β are of exactly the same form as those shown in Figs. 2 and 3. In this case y must be obtained by reference to the resonant frequency f_r and not to f_0. The quantities $n\alpha$ and m/α in the above formula can be rearranged in the following way :—

$$n\alpha = \frac{nf}{f_0} = \frac{f}{f_r}\,.\,\sqrt{mn}, \qquad \frac{m}{\alpha} = \frac{mf_0}{f} = \frac{f_r}{f}\,.\,\sqrt{mn}$$

Therefore the values of v and the feedback become :—

$$v = V_0\,.\,\frac{1}{\left(1+n+\dfrac{m}{n}\right)+jy\sqrt{mn}}$$

and $$\beta = \frac{1}{\left(1 + n + \frac{m}{n}\right) + jy\sqrt{mn}}$$

The essential difference being that, when y is zero, the value of the feedback is $1/(1+n+m/n)$ and not $1/3$. Thus with this circuit the value of A must be a little less than $(1 + n + m/n)$. Therefore the gain of the complete circuit will be :—

$$G = \frac{A}{1 - A\beta}$$

The gain of the amplifier on resonance is thus,

$$G_o = \frac{\left(1 + n + \frac{m}{n}\right)A}{\left(1 + n + \frac{m}{n}\right) - A}$$

Insertion of this value of G_o in the formula for G leads to the expression,

$$G = \frac{G_o\left(1 + j \cdot \frac{\sqrt{mn}}{1 + n + \frac{m}{n}}\right)}{1 + j\frac{yG_o}{A}\left(\frac{\sqrt{mn}}{1 + n + \frac{m}{n}}\right)}$$

From the previous formulæ of this kind, it can be seen at once that the effective Q-factor near resonance will be :—

$$Q = \frac{G_o\sqrt{mn}}{\left(1 + n + \frac{m}{n}\right)^2}$$

Here A has been replaced by $(1 + n + m/n)$ since to obtain a reasonable value of Q the value of A must be nearly equal to this quantity. If the expression for Q is differentiated first with respect to m and then n and the resulting differentials placed equal to zero, two conditions are obtained which must be satisfied in order that the value of Q shall be a maximum. These two conditions are $n^2 + n - 3m = 0$ and $3n^2 - n - 5m = 0$; solving these it is found that $n = 2$ and $m = 2$ for a maximum in Q. If these values of m and n are substituted in the formulæ above, the result is that $Q = G_o/8$ and the feedback $\beta = 1/(4 + j2y)$. The value of A is thus a little less than 4. In the case of a tunable selective amplifier the requirement of m equal to 2 is easily satisfied by employing a three-gang variable condenser.

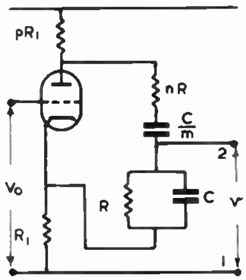


Fig. 19.—Phase-splitting valve plus the general case of the simple selective network.

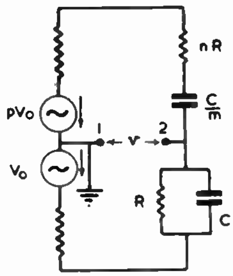


Fig. 20.—Equivalent electrical circuit of Fig. 19.

9.0. Simple Selective Network and the Phase Splitter Valve

The circuit diagram and the equivalent electrical circuit for this case are shown in Figs. 19 and 20. The anode load resistor is now p times that in the cathode lead. In the same manner as before, the voltage v between the terminals 1 and 2 is found to be :—

$$v = V_o \cdot \frac{\left(n + \frac{m}{n} - p\right) + jy\sqrt{mn}}{\left(1 + n + \frac{m}{n}\right) + jy\sqrt{mn}}$$

If now $p = (n + m/n)$, then v is zero when y is zero ; that is, at the resonant frequency $f_r = f_o\sqrt{m}/\sqrt{n}$, the feedback is zero. The effect of making p a little less than $(n + m/n)$ is ignored in this section ; the results of doing this are the same as before. With p equal to $(n + m/n)$ the value of the feedback β becomes :—

$$\beta = \frac{jy\sqrt{mn}}{\left(1 + n + \frac{m}{n}\right) + jy\sqrt{mn}}$$

If this circuit is now employed with an amplifier of gain G_o the gain of the resultant amplifier is given by :—

$$G = \frac{G_o\left\{1 + j\frac{y\sqrt{mn}}{\left(1 + n + \frac{m}{n}\right)}\right\}}{1 + j \cdot \frac{y(G_o + 1)\sqrt{mn}}{\left(1 + n + \frac{m}{n}\right)}}$$

The effective Q-factor for small values of y is thus :—

$$Q = \frac{(G_o + 1)\sqrt{mn}}{1 + n + \frac{m}{n}}$$

$$= \frac{(G_o + 1)}{\frac{1}{\sqrt{mn}} + \sqrt{\frac{m}{n}} + \frac{1}{n}\sqrt{\frac{m}{n}}}$$

Examination of this expression indicates that m and n should be as large as possible. In the limit if m and n are equal and infinitely large $Q = (G_o + 1)$. This is confirmed by differentiation with respect to m and n, that with respect to n yielding n equal to infinity, that with respect to m yielding, however, m equal to $n(n + 1)$. If this latter condition is substituted in the formula for the Q-factor, the result is :—

$$Q = (G_o + 1) \cdot \frac{n}{2\sqrt{n+1}}$$

Thus it would appear possible to obtain a Q-factor greater than $(G_o + 1)$. In actual fact a difficulty appears, since, with large values of n, the value of p also becomes large and the assumption that the voltage at the cathode of the phase splitter valve is V_o no longer holds to a sufficiently close approximation. If the characteristics of the triode employed are μ and R_a the voltage at the cathode is given by :—

$$V_c = V_o \cdot \frac{R_1}{\frac{R_a}{\mu} + \frac{(p + \mu + 1) R_1}{\mu}}$$

$$\simeq V_o \cdot \frac{\mu}{\mu + p + 1}$$

Thus only a fraction of the output voltage is fed back to the input with the result that a greater value of G_o is required to produce a given Q-factor than is indicated by the above formula for the Q-factor. Now p must be equal to $(n + m/n)$, and hence, from the condition for maximum Q-factor, p becomes $(2n + 1)$; therefore the voltage at the cathode is :—

$$V_c = V_o \cdot \frac{\mu}{2n + \mu + 2}$$

Thus the value of the feedback is actually given by :—

$$\beta = \frac{\mu}{2n + \mu + 2} \cdot \frac{jy}{\frac{2\sqrt{n+1}}{n} + jy}$$

where m has been replaced by the value $n(n + 1)$ so that the Q-factor is maximized as far as m is concerned. From this value of β the gain of the selective amplifier, and hence the Q-factor near resonance, can be derived. They are :—

$$G = \frac{G_o\left(1 + jy \cdot \frac{n}{2\sqrt{n+1}}\right)}{1 + jy\left(G_o + \frac{1}{\mu}\right) \cdot \frac{n}{2\sqrt{n+1}} \cdot \frac{\mu}{2n+\mu+2}}$$

and $$Q \simeq G_o \cdot \frac{n}{2\sqrt{n+1}} \cdot \frac{\mu}{2n + \mu + 2}$$

If the expression for Q is differentiated with respect to n and the result placed equal to zero, the condition for maximum Q is $n^2 - n(\mu/2 + 1) - \mu = 0$. Thus for a valve having a μ of 20 the value of n is 12·6 and the value of the Q-factor is $0{\cdot}72G_o$. The circuit is now more dependent on the constancy of the μ of the triode valve and is therefore not so stable as in the case when m and n are equal to unity and p is equal to 2. Other troubles also arise. For example the output impedance at the anode of the triode increases as the value of p increases, therefore an increasing allowance must be made in the series resistor of the selective network to correct for this. For example, if the μ of the triode is 20, its R_a is 10,000 ohms and the value of R_1 is 10,000 ohms and the value of p is chosen as 10 than the output impedance is approximately 70,000 ohms.

10.0. Component Tolerances

Up to this point it has been assumed that the components forming the selective network are exactly their correct values. In the case of a fixed frequency amplifier the components may be adjusted to any required accuracy by the use of trimmers and no difficulty arises ; the components do not form the limiting factor governing the stability except in so far as they alter with age and external conditions. When it is desired to have variable frequency characteristics the variation of the components necessary to produce the tuning presents a serious problem. Two methods of tuning are available. Either ganged resistors or ganged capacitors may be used. The former are to be preferred from the point of view of accuracy and the latter from the point of view of cost. In the case of resistors, an alignment accuracy of 0·1 per cent. can be obtained, and in the case of capacitors the alignment accuracy is about 1 per cent. for those at present obtainable. In order to assess the effects of errors in the components it is assumed that the circuit has been adjusted so that the components have the correct values at one frequency and that variable capacitors are used. When the amplifier is tuned to a new frequency, the capacitors will no longer be equal. Let one equal $(1 + k)$C as shown in Fig. 21. The factor k will be referred to as the error factor. With the same notation as before, the voltage v will be :—

$$v = V_o \cdot \frac{1}{(3 + k) + j(y + k\alpha)}$$

The value of the feedback will be given by :—

$$\beta = \frac{1}{(3 + k) + j(y + k\alpha)}$$

Thus the gain of the selective amplifier is :—

$$G = \frac{A\{(3 + k) + j(y + k\alpha)\}}{(3 + k - A) + j(y + k\alpha)}$$

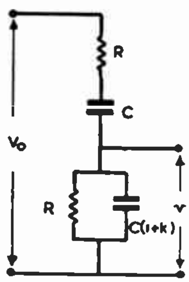


Fig. 21.—Simple selective network with the parallel capacitor $(1 + k)$ times its correct value.

This has its maximum value when $(y + k\alpha)$ is equal to zero, and it is to be noted that the resonance point has been shifted slightly. The value of the gain on resonance is given by :—

$$G_r = \frac{A(3 + k)}{(3 + k - A)}$$

This resonant gain occurs when $(y + k\alpha)=0$, that is when $\alpha = 1/\sqrt{1 + k} = 1 - k/2$ since the value of k will be small. Thus the resonant frequency is $f_r = (1 - k/2)\,\pi CR$, and the shift of the resonant frequency due to k is $k/4\pi CR$ c/s. From the expression for the gain G_o it can be seen that the possibility of oscillation arises ; this will occur if k is more negative than $(A - 3)$, remembering that in the simple case A is a little less than 3. Oscillation is only possible if k is negative, that is if the parallel condenser is smaller than the series condenser ; the condition for stability is $k > (A - 3)$. Since variations will occur in both of the capacitors in a random manner, the maximum permitted variation from the average is half of k. Alternatively, if the series capacitor is regarded as the standard, then the permitted variation is $-100k$ per cent. for the parallel one. While the foregoing shows that the maximum gain on tune and hence the selectivity obtainable must be below a certain limit to avoid self oscillation, this is not important in itself. The variation of the components from the correct values will cause variations in G_o and in the Q-factor, which are undesirable in a measuring instrument. When the amplifier is tuned to resonance the gain is $G_r = A(3 + k)/(3 - A + k)$, and therefore the rate of change of G_r with k is $dG_r/dk = -G_r^2/(3 + k)^2$. Thus the rate of change when k is zero is :—

$$\left(\frac{dG_r}{dk}\right)_{k=0} = -\frac{G_o^2}{9}$$

Hence the maximum permitted variation in G_r and the tolerance on the components determine the maximum gain which may be employed. If the maximum permitted variation in G_r is ± 5 per cent. and the tolerance on the capacitors is 1 per cent. the value of G_r that can be used is 45 ; this corresponds to a Q of 5 and a second harmonic rejection of the order of 6·5. This is about the best that can be obtained with this type. If variable resistors with an accuracy of 0·1 per cent. are employed and the same variation is allowed in G_r, the maximum value of G_r is 450 and the Q is 50. In this case the variation in A would probably be the limiting factor. The formula for the variation of G_r with k may also be written :—

$$\left(\frac{dG_r}{dk}\right)_{k=0} = -9Q^2$$

Hence the rate of change of the Q-factor and the band-width Δf_r of the amplifier with k are :—

$$\left(\frac{dQ}{dk}\right)_{k=0} = -Q^2 \text{ and } \left(\frac{d(\Delta f_r)}{dk}\right)_{k=0} = f_r$$

The rate at which the Q-factor varies with the errors in the components is dependent only on the Q.

The circuit employing the phase splitter valve will now be considered. Again it will be assumed that the parallel capacitor is $(1 + k)$ times its correct value as shown in Fig. 22. The output voltage v between the terminals 1 and 2 of the equivalent circuit shown in Fig. 23 is given by:—

$$v = V_o - \frac{3V_o}{R\left(1 - \frac{j}{\alpha}\right) + \frac{R}{1 + j(1+k)\alpha}} \cdot \frac{R}{1 + j(1+k)\alpha}$$

$$= V_o \cdot \frac{k + j(y + k\alpha)}{(3 + k) + j(y + k\alpha)}$$

Hence the value of the feedback is :—

$$\beta = \frac{k + j(y + k\alpha)}{(3 + k) + j(y + k\alpha)}$$

Thus β no longer falls to zero at any frequency; it has its minimum value when $(y + k\alpha) = 0$, that is when α is approximately $(1 - k/2)$. The resonant frequency is then $(1 - k/2)/2\pi CR$. Therefore the minimum value of the feedback β is $k/(3 + k)$ and therefore the gain on resonance is given by :—

$$G_r = \frac{G_o(3 + k)}{3 + k(G_o + 1)}$$

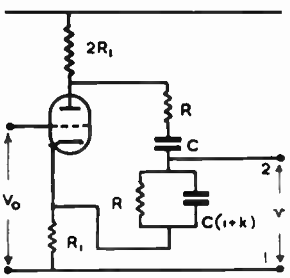


Fig. 22.—Circuit of Fig. 12 with an error in one component of the selective network.

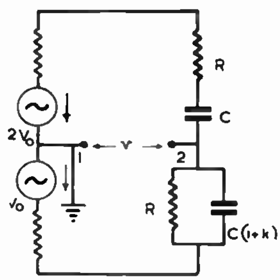


Fig. 23.—Equivalent electrical circuit of Fig. 22.

From this expression it can be seen that oscillation can occur if $k = -3/(G_o + 1)$. If G_r is differentiated with respect to k then $dG_r/dk = -3G_r^2/(3 + k)^2$, and so when k is zero the rate of change of G_r with k is

$$\left(\frac{dG_r}{dk}\right)_{k=0} = -\frac{G_o^2}{3}$$

If as before the maximum permitted variation of G_r is $\pm$ 5 per cent. and the tolerance on the capacitors is $\pm$1 per cent., the maximum value of G_r that may be employed is 15, which corresponds to a Q of 5. It is to be noted that this is exactly the same value of Q as before. Thus, as far as stability is concerned, the only advantage gained by using the more complicated circuit is the decreased gain that may be used, the variation of Q due to variations in the components being independent of the circuit employed. The rate of change of the gain with k may be written :—

$$\left(\frac{dG_r}{dk}\right)_{k=0} \doteq -3Q^2$$

Therefore the rate of change of the Q-factor and hence of the band-width Δf_r with k is :—

$$\left(\frac{dQ}{dk}\right)_{k=0} \doteq -Q^2 \text{ and } \left(\frac{d(\Delta f_r)}{dk}\right)_{k=0} = f_r$$

These are identical with the values obtained for the simple selective circuit.

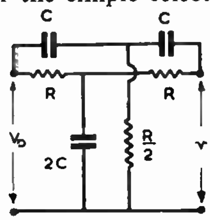


Fig. 24.—The symmetrical Twin-T network.

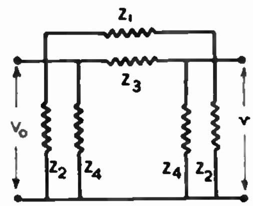


Fig. 25. — Network obtained from Fig. 24 by replacing each T network by its equivalent π network.

11.0. The Twin-T Network

The twin-T network will now be considered, and the form of the circuit is as shown in Fig. 24. The symmetrical form of the network is not essential and the general case will be considered later. It is assumed that the network is working from a source of zero impedance and that the load is of infinite impedance. This is not essential but the selectivity is greatest under these conditions, the network, being a bridge circuit, has zero output at one frequency whatever the source and load impedances. If each T network of the twin is converted to its equivalent π network the result is as shown in Fig. 25, where the impedances are given by :—

$$\begin{cases} Z_1 = 2R(1 + j\omega CR) = 2R(1 + j\alpha) \\ Z_2 = R\left(1 + \frac{1}{j\omega CR}\right) = R\left(1 - \frac{j}{\alpha}\right) \end{cases}$$

$$\begin{cases} Z_3 = \frac{2}{i\omega C}\left(1 + \frac{1}{j\omega CR}\right) = \frac{2}{j\omega C}\left(1 - \frac{j}{\alpha}\right) \\ Z_4 = R\left(1 + \frac{1}{j\omega CR}\right) = R\left(1 - \frac{j}{\alpha}\right) \end{cases}$$

The four shunt arms are of identical character, each impedance consisting of a resistance R in series with a capacitance C. The series arms are not the same. The impedance Z_1 consists of a resistance 2R in series with an inductance of magnitude $2CR^2$, Z_3 consists of a negative resistance of magnitude $2/\omega^2C^2R$ in series with a capacitance C/2. Thus the resultant series aim of the circuit of Fig. 25 consists of a rejector circuit which resonates at a frequency $f_o = 1/2\pi CR$, and whose impedance at this frequency is infinite. Fig. 26 shows the final equivalent circuit, and the impedances of the arms of the π networks are :—

$$Z_5 = \frac{2R\left(1 - \frac{j}{\alpha}\right)}{j\left(\alpha - \frac{1}{\alpha}\right)} \text{ and } Z_6 = \frac{R}{2}\left(1 - \frac{j}{\alpha}\right)$$

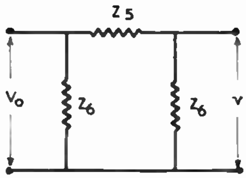


Fig. 26.—Resultant π network obtained from Fig. 25.

The fact that Z_5 does become infinite at the frequency at which α equals unity is clearly shown by the formula for this arm. The output at this frequency is therefore zero ; this results in zero negative feedback being applied to the amplifier at this frequency, and this is the case whatever the impedance of the source and load The only effect that finite generator and load impedances can have is that of reducing of the selectivity. If the voltage V_o, from a generator of zero internal impedance, is applied to the network and if the load is of infinite impedance the output voltage v is :—

$$v = V_o \cdot \frac{Z_6}{Z_5 + Z_6}$$

On insertion of the values of Z_5 and Z_6 this reduces to :—

$$v = V_o \cdot \frac{1}{1 + \frac{4(1 + j\alpha)}{\left(1 - \frac{j}{\alpha}\right)(1 - \alpha^2)}}$$

$$= \frac{V_o}{1 + \frac{4}{jy}}$$

Thus the value of β when this circuit is used as a feedback network is given by :—

$$\beta = \frac{1}{1 + \frac{4}{jy}}$$

The form of this equation is identical with that already described in the circuit employing the phase splitter valve and the same details apply in this case. If used with an amplifier of gain G_o, then the gain with feedback is,

$$G = \frac{G_o\left(1 + j\frac{y}{4}\right)}{\left(1 + jy\frac{G_o + 1}{4}\right)}$$

Thus the Q-factor obtained in this case is :—

$$Q = \frac{G_o + 1}{4}$$

This is of the same order as before, but it is obtained at the expense of six components for the selective network instead of four ; this is an important point when the problem of varying the resonant frequency is considered. In this case it is necessary to vary three components instead of two, and thus two error factors must now be taken into account. A circuit suitable for use with this network is shown in Fig. 27.

12.0. Phase Shift in the Amplifier and Self Oscillation

It is of interest to consider the effect of phase shift in the amplifier; up to this point it has been assumed that G_o is real and positive. The connections to the selective network have been so arranged that the product of β and G_o is such that the resultant gain of the amplifier is:—

$$G = \frac{G_o(4 + jy)}{4 + jy(G_o + 1)}$$

It is now assumed that the gain of the amplifier is $G_o\underline{/\theta}$. In a two-stage amplifier θ is approximately 360° over the middle range of

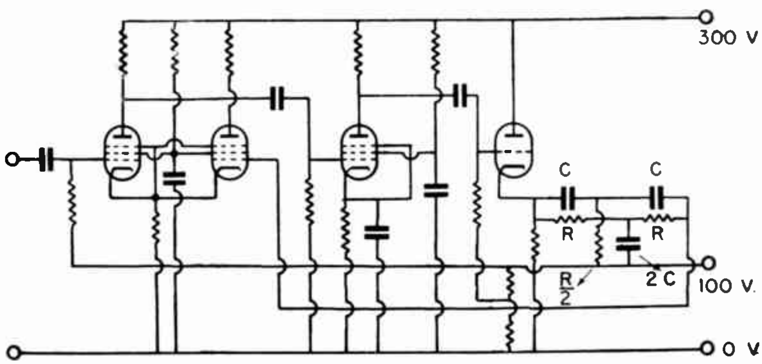


Fig. 27.—Circuit of a selective amplifier employing the twin-T network.

frequencies ; it increases to 540° as the frequency tends to zero if R-C couplings are employed, and at high frequencies it tends to 180° due to the stray capacitance shunting the anode loads. These figures assume that there are only two phase shifting networks at both low and high frequencies ; this is not always the case even in some apparent two-stage amplifiers. The gain of the selective amplifier now becomes :—

$$G = \frac{G_o\underline{/\theta}(4 + jy)}{4 + jy(G_o\underline{/\theta} + 1)}$$

Since $G_o\underline{/\theta}$ can be expressed in the form $G_o(\cos\theta + j\sin\theta)$, this expression for the gain can be written

$$G = G_o . \frac{(4\cos\theta - y\sin\theta) + jy\left(\cos\theta + \dfrac{4\cos\theta}{y}\right)}{(4 - yG_o\sin\theta) + jy(1 + G_o\cos\theta)}$$

If the denominator of this expression is zero the gain will be infinite, that is the circuit is self oscillatory. This occurs when $\cos\theta = -1/G_o$ and $\sin\theta = 4/yG_o$. It is now supposed that the selective network is adjusted so that the resonant frequency of the amplifier is f_r. Then the value of y for frequencies above f_r is positive. Therefore, for oscillation to be possible above f_r, $\cos\theta$ must be negative and $\sin\theta$ positive ; thus θ must lie between 90° and 180°, or of course, between $(90° + 2n\pi)$ and $(180° + 2n\pi)$ where n is an integer. As given above, in a two-stage amplifier at high frequencies, θ lies between 180° and 360° and thus oscillation is not possible in this case. At frequencies below f_r, the factor y is negative and therefore both $\cos\theta$ and $\sin\theta$ must be negative for oscillation to be possible, that is θ must lie between 180° and 270°, or $(180° + 2n\pi)$ and $(270° + 2n\pi)$. Again this is not possible with a two-stage amplifier, since at low frequencies θ lies between 360° and 540°, and thus oscillation is not possible below f_r. In a three-stage amplifier it would be possible for oscillation to occur.

Now it is assumed that the connections to the selective network are re-arranged so that the gain is :—

$$G = \frac{G_o\underline{/\theta}(4 + jy)}{4 - jy(G_o\underline{/\theta} - 1)}$$

This can be rearranged to give :—

$$G = G_o . \frac{(4\cos\theta - y\sin\theta) + j(y\cos\theta + 4\sin\theta)}{(4 + yG_o\sin\theta) + jy(1 - G_o\cos\theta)}$$

Oscillation will now be possible if $\cos\theta = 1/G_o$ and $\sin\theta = -4/yG_o$. If y is positive this implies that θ must lie between 270° and 360° ; this is possible in a two-stage amplifier at high frequencies, so that oscillation can occur. Similarly at low frequencies, when y is negative, θ must lie between 0° and 90° ; this is again possible in a two-stage amplifier so that oscillation can occur at low frequencies. In an actual circuit, the oscillation would probably consist of bursts of high-frequency oscillation superimposed on a very low-frequency oscillation resembling that of a multivibrator

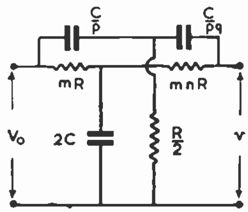


Fig. 28.—General case of the twin-T network.

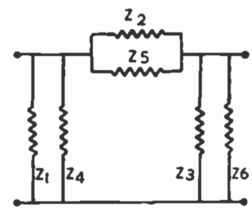


Fig. 29. — Network obtained from Fig. 28 by replacing each T network by its equivalent π network.

13.0. General Case of the Twin-T Network

The asymmetric twin-T network will now be dealt with, and the form of the network is as shown in Fig. 28. As in the case of the symmetrical network each T network is converted to its equivalent π network as shown in Fig. 29. The values of the impedances which make up the π networks are :—

$$Z_1 = R\left\{m - j\left(1 + \frac{1}{n}\right) . \frac{1}{2\alpha}\right\}$$

$$Z_2 = mR\left\{(n + 1) + j2mn\alpha\right\}$$

$$Z_3 = R\left\{mn - j(n + 1) . \frac{2}{\alpha}\right\}$$

$$Z_4 = \frac{R}{q}\left\{\frac{(q + 1)}{2} - j . \frac{pq}{\alpha}\right\}$$

$$Z_5 = \frac{p}{j\omega C}\left\{(q + 1) - j\frac{2pq}{\alpha}\right\}$$

$$Z_6 = R\left\{\frac{(q + 1)}{2} - j . \frac{pq}{\alpha}\right\}$$

The series arm Z_2 consists of a resistance and an inductance in series ; the series arm Z_5

consists of a negative resistance and a capacitance in series. Thus when $(Z_2 + Z_5)$ is equal to zero, the impedance of the total series arm of the π network is infinite and the output from the network is zero. This will occur when :—

$$mR\{(n+1)+j2mn\alpha\}+\frac{p}{j\omega C}\left\{(q+1)-j.\frac{2pq}{\alpha}\right\}=0$$

The real and imaginary parts of this yield:—

$$\alpha^2 = \frac{2p^2q}{m(n+1)} \text{ and } \alpha^2 = \frac{p(q+1)}{2m^2n}$$

These two expressions give the frequency at which the output is zero ; the result must be the same from either, and therefore, by equating these two expressions, the condition for zero output is obtained. This is :—

$$4mnpq = (n+1)(q+1)$$

The value of the resonant frequency is given by :—

$$f_r = \frac{1}{2\pi CR}\cdot\sqrt{\frac{2p^2q}{m(n+1)}} = \frac{1}{2\pi CR}\cdot\sqrt{\frac{p(q+1)}{2m^2n}}$$

The manner in which the output voltage v varies with frequency is given by :—

$$v = \frac{V_o}{1+\dfrac{Z_2Z_5(Z_3+Z_6)}{Z_3Z_6(Z_2+Z_5)}}$$

Insertion of the values of the various impedance yields :—

$$v = \frac{V_o}{1+\dfrac{2(2mn+q+1)\left\{1-j.\dfrac{(2pq+n+1)}{(2mn+q+1)}\right\}}{(n+1)-j2mn\alpha-j.\dfrac{p(q+1)}{m\alpha}-\dfrac{2p^2q}{m\alpha^2}}}$$

By suitable manipulation, and remembering the condition for a zero of output occurring at one frequency, this can be reduced to :—

$$v = \frac{V_o}{1+\dfrac{2(2mn+q+1)}{j2mn\dfrac{\alpha}{p}-j.\dfrac{1}{\alpha}.\left(\dfrac{q+1}{m}\right)}}$$

The α in this equation has a value f/f_o, where $f_o = 1/2\pi CR$; this must be converted to α_r which is f/f_r. The frequency f_r is the resonant frequency at which zero output occurs. Thus the following are obtained :—

$$\frac{2mn\alpha}{p} = \alpha_r\sqrt{\frac{2n(q+1)}{p}}$$

$$\text{and } \frac{1}{\alpha}\left(\frac{q+1}{m}\right) = \frac{1}{\alpha_r}\sqrt{\frac{2n(q+1)}{p}}$$

The insertion of these values in the equation for v yields :—

$$v = \frac{V_o}{1+\dfrac{2(2mn+q+1)}{\sqrt{\dfrac{2n(q+1)}{p}}\left(j\alpha_r-\dfrac{j}{\alpha_r}\right)}}$$

Therefore

$$v = \frac{V_o}{1+\dfrac{\sqrt{\dfrac{2m(n+1)}{q}}+\sqrt{\dfrac{2p(q+1)}{n}}}{jy}}$$

Where use has again been made of the balance condition, $4mnpq = (n+1)(q+1)$. Thus the feedback β is of the form $1/(1 + D/jy)$, which is of exactly the same form as in the special case already dealt with. Thus the gain of the selective amplifier and of the Q-factor are given by the expressions :—

$$G = \frac{G_o\left(1+\dfrac{jy}{D}\right)}{1+jy\left(\dfrac{G_o+1}{D}\right)} \text{ and } Q = \frac{G_o+1}{D}$$

when the value of D is :—

$$D = \sqrt{\frac{2m(n+1)}{q}}+\sqrt{\frac{2p(q+1)}{n}}$$

It is of interest to determine the minimum value of D, since this will give the maximum Q-factor. If n and q are fixed and the resonant frequency is also fixed, then, since $4mnpq = (n+1)(q+1)$, the product mp must be constant ; thus D reduces to :—

$$D = \sqrt{\frac{2m(n+1)}{q}}+\sqrt{\frac{2k(q+1)}{mn}}$$

Where p has been replaced by k/m and k is constant. If this expression is differentiated with respect to m and the result placed equal to zero, the condition for a minimum of D is obtained ; this is, $m = (q+1)/2n$. In a similar manner, differentiating with respect to p, it is shown that p must equal $(n+1)/2q$ for a minimum of D. If these values are substituted in the expression for D the result is :—

$$D = 2\sqrt{\left(1+\frac{1}{n}\right)\left(1+\frac{1}{q}\right)}$$

From this equation it can be seen that n and q should be as large as possible. In the limit when $n = q = \infty$ and $m = p = 1/2$ the value of D is 2. Thus the maximum Q-factor that can be obtained with this circuit is $(G_o + 1)/2$. In a practical case if $n = q = 10$ and $m = p = 0{\cdot}742$ the value of D is 2·56 and the Q obtained is $(G_o + 1)/2{\cdot}56$.

It is to be noted that the input impedance is no longer equal to the output impedance ; the output impedance becomes very large and this may be of importance if the input impedance of the circuit into which the twin-T network is to work is not extremely large.

14.0. Component Tolerances

The effect of errors in the sizes of the components forming the selective network can best be determined by using the equation for v which is reproduced here :—

$$v = \cfrac{V_o}{1+\cfrac{2(2mn+q+1)\left\{1-j\left(\frac{2pq+n+1}{2mn+q+1}\right)\frac{1}{\alpha}\right\}}{(n+1)+j2mn\alpha-j.\frac{p(q+1)}{m\alpha}-\frac{2p^2q}{m\alpha^2}}}$$

Thus the feedback is of the form $\beta = 1/(1+X)$, and the value of X is :—

$$X = \frac{2(2mn+q+1)\left\{1-j\left(\frac{2pq+n+1}{2mn+q+1}\right)\frac{1}{\alpha}\right\}}{(n+1)+j2mn\alpha-j.\frac{p(q+1)}{m\alpha}-\frac{2p^2q}{m\alpha^2}}$$

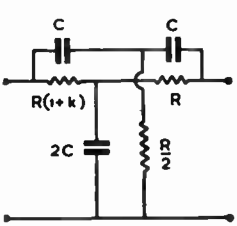


Fig. 30.—The twin-T network with one of its series resistors $(1 + k)$ times its correct value.

Now, considering the circuit shown in Fig. 30, the input series resistor is $(1 + k)$ times the correct value ; $100k$ is the percentage error. In this network m is equal to $(1 + k)$, and n is equal to $1/(1+k)$, which is approximately $(1-k)$ since k is small, and $p = q = 1$. Insertion of these values in the formula for X yields :—

$$X = \frac{4\left\{1-j\left(1-\frac{k}{4}\right)\frac{1}{\alpha}\right\}}{\left(1-\frac{k}{2}\right)+j\alpha-j(1-k)\frac{1}{\alpha}-(1-k)\frac{1}{\alpha^2}}$$

Therefore

$$X = \frac{4}{j\alpha-j(1-k)\frac{1}{\alpha}-\frac{k}{4}\left(1+\frac{1}{\alpha^2}\right)\left\{\frac{1}{1-j\left(1-\frac{k}{4}\right)\frac{1}{\alpha}}\right\}}$$

The term $(1 - k/4)$ in the last term in the denominator may be assumed equal to unity with no appreciable error, since the whole term is only a correction term, thus :—

$$X = \frac{4}{j\alpha-j\left(1-\frac{3k}{4}\right)\frac{1}{\alpha}-\frac{k}{4}}$$

The denominator of this expression has its minimum value, and hence also β has its minimum value, when $\{j\alpha - j(1 - 3k/4)/\alpha\}$ is equal to zero. Thus the amplifier has its maximum gain, that is it is on tune, when $\alpha^2 = (1 - 3k/4)$; thus $\alpha \triangleq (1 - 3k/8)$ since k is small. The minimum value of β is $1/(1-16/k)$, and therefore the gain of the amplifier on resonance is :—

$$G_r = \frac{G_o(k-16)}{k(G_o+1)-16}$$

where G_r is used to denote the gain at the new resonant frequency which is $f_r = (1 - 3k/8)/2\pi CR$. From the expression for the gain G_r, it can be seen that if k is positive and larger than $16/(G_o + 1)$ the circuit will oscillate; thus the error factor must be less than $16/(G_o + 1)$. The important factor is the rate of change of gain with small values of the error factor and its resultant effect on the Q-factor and the bandwidth of the amplifier. If the same procedure is adopted as in the previous section on errors, the following rates of change are obtained :—

$$\frac{dG_r}{dk} = \frac{16G_r^2}{(k-16)^2} \text{ and } \left(\frac{dG_r}{dk}\right)_{k=0} = \frac{G_o^2}{16}$$

Since the value of the Q-factor in this case is $(G_o + 1)/4$ the latter equation may be written:—

$$\left(\frac{dG_r}{dk}\right)_{k=o} = \left(\frac{G_o}{G_o+1}\right)^2 . Q^2 \triangleq Q^2$$

Thus the variation of the Q-factor and the band-width with the error factor k are :—

$$\left(\frac{dQ}{dk}\right)_{k=o} = \frac{Q^2}{4} \text{ and } \left(\frac{d(\Delta f_r)}{dk}\right)_{k=o} = \frac{f_r}{4}$$

If these are compared with the corresponding quantities obtained previously then this circuit is apparently four times as good ; it must be remembered, however, that in this circuit there are three components to be varied, and hence two error factors must be taken into account. If the shunt resistor is taken as the standard the errors in the other two resistors may be such that they add and therefore all the rates of change found above must be multiplied by two in order to obtain the relevant quantities for the twin-T network. Thus this network is twice as good as those previously dealt with.

The case of variable capacitors may be dealt with in exactly the same manner ; if, as shown in Fig. 31, one of the series capacitors is assumed to be $(1 + k)$ times its correct value then the corresponding rates of change are :—

$$\left(\frac{dG_r}{dk}\right)_{k=o} = -\frac{G_o^2}{16}, \left(\frac{dG_r}{dk}\right)_{k=o} = -Q^2$$

$$\left(\frac{dQ}{dk}\right)_{k=o} = -\frac{Q^2}{4}, \left(\frac{d(\Delta f_r)}{dk}\right)_{k=o} = -\frac{f_r}{4}$$

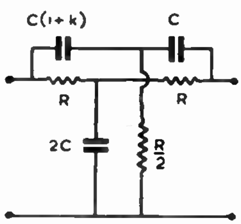


Fig. 31.—The twin-T network with one of its series capacitors $(1 + k)$ times its correct value.

15.0. Circuits Employing Only a Fraction of the Available Feedback

At this point it is convenient to consider the arrangement employed in certain selective amplifiers, namely that of applying only a fraction $1/n$ of the output of the selective network to the input. In this case the value of β is $jy/n(D + jy)$ with the result that the gain of the amplifier is :—

$$G = \frac{G_o\left(1 + j \cdot \frac{y}{D}\right)}{1 + jy \cdot \frac{\left(\frac{G_o}{n} + 1\right)}{D}}$$

Thus the Q-factor obtained in this case is $(G_o/n + 1)/D$, and therefore in order to obtain the same value of Q as before, it is necessary that the amplifier gain G_o without feedback should be n times as large. This makes the problem of stabilizing the circuit more difficult. To secure the same stability against changes in the valves, etc., the feedback factor $\beta^- M$ must have the same value as before. Now β^-, which is the negative feedback employed, is approximately equal to $1/G_o$ and therefore, since G_o must be n times as large, the value of the negative feedback has been decreased by a factor n. Therefore if the product $\beta^- M$ is to have the same value, M must be approximately n times as large.

If the effect of the errors in the components used in the selective network is considered, the result of the same steps as were taken in the previous case yields :—

$$\left(\frac{dG_r}{dk}\right)_{k=o} = \frac{G_r^2}{16n}$$

Again it must be remembered that G_o is n times its previous value. The variation in the Q-factor and the band-width are :—

$$\left(\frac{dQ}{dk}\right)_{k=o} = \frac{Q^2}{4} \text{ and } \left(\frac{d(\Delta f_r)}{dk}\right)_{k=o} = \frac{f_r}{4}$$

These are exactly the values found before and thus no improvement has been effected here. The only improvement is that G_o is not so small, which may be of interest if the signals are very small and additional amplification is required in any case. It must be remembered, however, that there is a limit to the amplification M that can be obtained with a two-stage amplifier and to use more than two stages immediately introduces serious difficulties with regard to self-oscillation of the circuit as a whole.

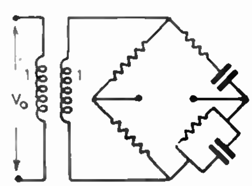


Fig. 32.—The Wien bridge as the frequency sensitive feedback network.

16.0. The Wien Bridge

The final network which will be dealt with is the Wien bridge, the circuit of which is as shown in Fig. 32 ; a transformer must be used since the input and output have no common point. It is not possible to use a pair of valves in push-pull and then to connect the bridge between the anodes, as it is quite easy to show that a very small inequality in the anode voltages in the wrong direction can cause oscillation. The use of a transformer leads to difficulties since, instead of 90° being the maximum phase change at high and low frequencies as with R-C couplings, the phase change can rise to 180°. Thus, if the feedback is applied over more than one stage, considerably care is necessary in the design of the coupling networks generally in order to avoid self oscillation.

By using the same methods and notation as previously employed the output voltage v from the bridge is found to be :—

$$v = \frac{V_o}{2} \cdot \frac{jy}{4 + jy}$$

Hence this corresponds exactly to the case dealt with in the last section with the value of n equal to 2. Alternatively, since a transformer must be used, if a 2 : 1 step-up is employed the value of the feedback becomes :—

$$\beta = \frac{jy}{4 + jy}$$

The results of using this network are exactly similar to those obtained with the twin-T network except that four components and a transformer must be used in place of six components.

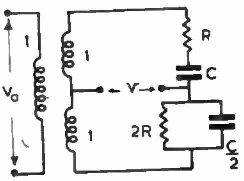


Fig. 33.—Modified form of the Wien bridge using a centre-tapped transformer in place of the resistive ratio arms.

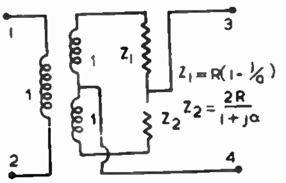


Fig. 34.—Rearrangement of the circuit of Fig. 33.

The fact that this circuit is the equivalent of the twin-T network can be shown in the following way. The resistive ratio arms of the bridge are not essential, and they may be removed if the transformer is centre-tapped as shown in Fig. 33, the ratio of the primary to half the secondary being 1 : 1. This circuit is redrawn in Fig. 34. Figure 35 shows a lattice network which, by suitable choice of the impedances Z_a and Z_b, will be the exact equivalent of the circuit of Fig 34. When the two are the equivalent of one another, the open and short circuit impedances of the two are the same. This gives the equations :—

$$Z_{oc} = \frac{Z_1 + Z_2}{4} = \frac{Z_a + Z_b}{2}$$

$$Z_{sc} = \frac{Z_1 Z_2}{Z_1 + Z_2} = \frac{2 Z_a Z_b}{Z_a + Z_b}$$

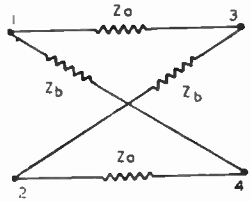


Fig. 35.—Lattice equivalent of the circuit of Fig. 34 and hence of that of Fig. 33.

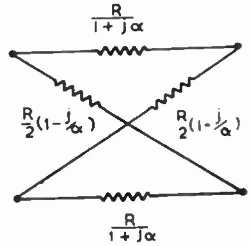


Fig. 36.—Lattice equivalent of Fig 33 with the values of the impedances inserted.

These are obtained viewing the networks from terminals 1 and 2 while terminals 3 and 4 are open- and short-circuited respectively. Exactly the same equations are obtained viewing from 3 and 4, while 1 and 2 are open and short-circuited since the networks are symmetrical. One solution of these equations is :—

$$Z_a = \frac{Z_2}{2} \text{ and } Z_b = \frac{Z_1}{2}$$

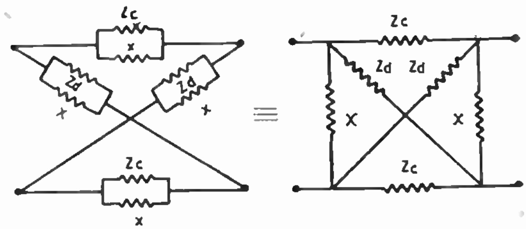


Fig. 37.—A lattice identity showing the removal of a common shunt element.

Thus the circuit shown in Fig. 36 is the equivalent of that in Fig. 33. Now in the case of a lattice network, if the arms contain a common element this may be removed from the lattice and placed at the input and output of the

Fig. 38.—Symmetrical H network obtained by treating the whole of the lattice arms of Fig. 35 as common shunt elements.

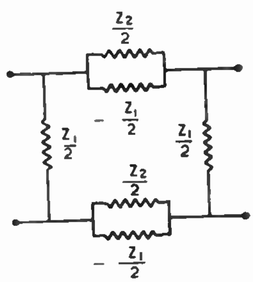
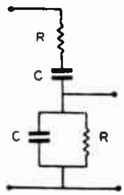
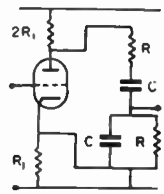
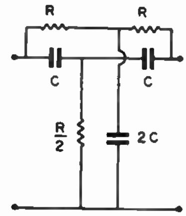
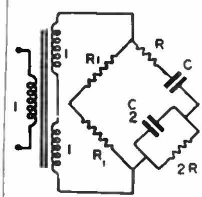


TABLE 1

Circuit				
Value of the feedback factor β	$\dfrac{1}{3+jy}$	$\dfrac{jy}{3+jy}$	$\dfrac{jy}{4+jy}$	$\dfrac{jy}{4+jy}$
Gain of the selective amplifier	$\dfrac{G_o\left(1+j\dfrac{y}{3}\right)}{1+jy\,.\,\dfrac{G_o}{9}}$	$\dfrac{G_o\left(1+j\dfrac{y}{3}\right)}{1+jy\left(\dfrac{G_o+1}{3}\right)}$	$\dfrac{G_o\left(1+j\dfrac{y}{4}\right)}{1+jy\left(\dfrac{G_o+1}{4}\right)}$	$\dfrac{G_o\left(1+j\dfrac{y}{4}\right)}{1+jy\left(\dfrac{G_o+1}{4}\right)}$
Rate of change of gain at resonance with error factor k ..	$\pm\dfrac{G_o^2}{9}$	$\pm\dfrac{G_o^2}{3}$	$\pm\dfrac{G_o^2}{16}$	—
Rate of change of the Q factor with error factor k	$\pm Q^2$	$\pm Q^2$	$\pm\dfrac{Q^2}{4}$	—
Value of the Q-factor obtained	$\dfrac{G_o}{9}$	$\dfrac{G_o+1}{3}$	$\dfrac{G_o+1}{4}$	$\dfrac{G_o+1}{4}$
Notes :—	Two components only to be varied.	Conveniently applies negative feedback. No more stable than previous circuit.	Three variables required. Two error factors to take into account. Twice as stable as first two.	No common earth point.

Positive signs apply when variable elements are resistors, negative signs with capacitors.

network. This is seen in the relevant case shown in Fig. 37.

If the network of Fig. 36 is treated in this way by removing the whole of the lattice arms, the result is as shown in Fig. 38. The series arms which remain after this operation consist of $Z_2/2$ in parallel with $-Z_1/2$ and the impedance of the series arms is therefore :—

$$\frac{-\frac{Z_2}{2}\cdot\frac{Z_1}{2}}{\frac{Z_2}{2}-\frac{Z_1}{2}} = \frac{-\frac{1}{2}\cdot\frac{2R}{(1+j\alpha)}\cdot\frac{R}{2}\left(1-\frac{j}{\alpha}\right)}{\frac{R}{(1+j\alpha)}-\frac{R}{2}\cdot\left(1+\frac{j}{\alpha}\right)}$$

This reduces to :—

$$R\cdot\frac{(1+j\alpha)}{(1-\alpha^2)}$$

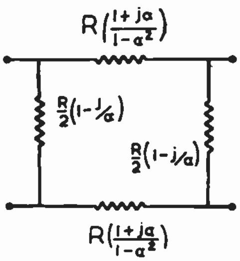


Fig. 39.—Symmetrical H network which is the equivalent of the circuit of Fig. 33.

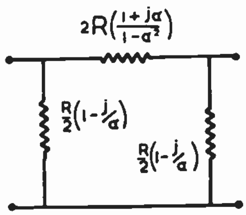


Fig. 40.—The π network obtained from the H network of Fig. 39.

Thus the resultant network is as shown in Fig. 39. This H-network may be converted to the π network as shown in Fig. 40 ; this is identical with the equivalent π network which was derived for the twin-T, and therefore the two networks are identical in all respects. This network will not be dealt with to any further extent, since the presence of the transformer renders it far less useful than the twin-T network.

17.0. Conclusion

The main details of the foregoing are shown in Table 1. This shows quite clearly that when constancy of the Q-factor of the amplifier is of paramount importance, the Twin-T network should be employed. Against this is the fact that three components must be varied and these must be accurately ganged. The fact that the Twin-T network results in an improvement by a factor of at least 2 over the simpler network is of importance. On the other hand, in cases where some variation of the Q-factor can be permitted the simpler circuit requiring only two variable elements represents a considerable saving in cost and in adjustment of the network. It is to be remembered that not only must the variable elements be accurately ganged, but also that the fixed components must be adjusted to their correct values.

NOTICES

Awards for the Study of Technology in U.S.A.

The Ministry of Education has announced that some 50 more post-graduate awards for the study of technology will be made available in the session 1950-51 at selected Universities and Technological institutions in the United States.

The first awards under this scheme were made last year. Some holders are now in the United States ; the remainder are in the process of taking up their awards.

The cost of the awards is met from funds provided by the American Economic Co-operation Administration. The aim of the scheme is to produce a small group of well-qualified men who will return to industrial posts or universities and technical colleges in this country.

The awards are tenable initially for a minimum of one year, but is is hoped they will be extended for a further year. They are open to students who hold a good Honours Degree in Pure Science or Technology, who have had at least two years' industrial experience, and who are now working in industry or research associations, or are teaching in Universities or Technical Colleges.

A substantial part of the time will be spent in selected industrial establishments. The awards will provide the tuition fees, books and travelling expenses in the United States and a maintenance allowance of $1,800 per annum. The cost of the return passage to the United States will be met from public funds. Successful candidates will be expected to leave the United Kingdom early in September next.

Full details may be obtained from the Ministry of Education (F.E. Division 1), Curzon Street House, Curzon Street, London, W.1.

Standard Frequency Transmissions

In September 1948, the Department of Scientific and Industrial Research announced that arrangements were being considered for an experimental service of standard frequency transmissions from the United Kingdom. A committee under the chairmanship of Dr. R. L. Smith-Rose established the need for such a service and, at the request of the Department, the General Post Office has assumed technical responsibility for the transmissions, which will take place from the Rugby radio station. The service began on February 1st, 1950. The frequencies used are 60 kc/s, 5 Mc/s and 10 Mc/s. The transmissions on 60 kc/s should be received throughout the United Kingdom and Western Europe and enable local standards to be calibrated with high precision. The transmissions on 5 and 10 Mc/s form part of an international programme designed to give reliable world coverage on one or other of the frequencies 2·5, 5, 10, 15, 20, 25 Mc/s which have been allocated to standard frequency services. The transmissions on these frequencies from the U.S.A. National Bureau of Standards station WWV are not always satisfactorily received in the United Kingdom and farther east. It is hoped to learn from the experimental service now being initiated to what extent reception in the European area is improved by transmissions from the United Kingdom and also to what extent the usefulness of both the U.S.A. and U.K. transmissions may be impaired by mutual interference.

The frequencies, which are to be maintained within two parts in one hundred million of the nominal values, are monitored at the National Physical Laboratory and all enquiries or comments concerning the transmissions should be addressed to the Director, National Physical Laboratory, Teddington, Middlesex, England. Information about reception conditions and interference with the U.S.A. transmissions will be particularly useful.

Bellahouston Park

On December 12th, a new transmitting station in Bellahouston Park started radiating the Third Programme for Glasgow listeners. It took over the service from the transmitter that was installed in Broadcasting House, Glasgow, as an emergency measure in 1940.

The new station, which is on a 2½-acre site, has two transmitters, each capable of 2 kilowatts output power, and a T aerial supported by a pair of tubular masts, 126 ft. high. By having duplicate transmitters, a very reliable service is ensured.

The twofold increase in power and the more efficient aerial system at the new station gives listeners to the Third Programme a stronger signal in most parts of Glasgow.

Liverpool Transmitting Station

During the week beginning January 22nd, the Third Programme transmitter at Dryden Road, Liverpool, which was installed in 1940, closed down and the service was transferred to the B.B.C.'s new station in Mersey Road, Birkenhead. This station is on a two-acre site, and comprises a 1-kW transmitter housed in a temporary building and feeding into a T aerial supported by two tubular steel masts, 126-ft high. Ultimately, when the permanent building has been completed, the station will have two 1-kW transmitters, one for service and one in reserve. These transmitters will have no engineers on the spot to operate them, but will be remotely controlled over telephone lines from Broadcasting House, Manchester. By means of the remote-control system the engineers in Manchester will be able to bring either transmitter into service and check its performance, output power, etc. They will also be able to correct the carrier frequency of the transmission, should it deviate from its correct value, which is at present 1,474 kc/s, corresponding to a wavelength of 203·5 m.

The new station is ten times more powerful than the original, and this increase, together with the more efficient aerial that has been erected, will result in better reception of the Third Programme in most parts of the district.

Educational Tours

The Engineers' Educational Travel Club has been formed by a body of engineers for the purpose of providing facilities for fellow engineers to visit engineering installations of interest abroad. The tours combine education and recreation and membership is open to all engaged in engineering.

The 1950 programme includes visits to Italy, Norway, France, Switzerland, Belgium, etc., and details may be obtained from the Secretary, 35 St. George's Square, London, S.W.1.

Television Transmission Equipment

The BBC announces that orders have been placed for the supply of two 50-kW vision transmitters and the associated two 12-kW sound transmitters for television stations to be built in Scotland and in the Bristol Channel area. It is hoped to acquire a site for the Scottish station at Kirk of Shotts, near Harthill and almost midway between Edinburgh and Glasgow. The location of the Bristol Channel station will depend on site tests now in progress in South Wales and North Somerset.

These orders complete the transmitter equipment required for the first stage of the announced programme of television expansion, consisting of five high-power stations.

R.E.C.M.F. Technical Bulletin

The Institution has received a copy of the Technical Bulletin issued by the Radio and Electronic Component Manufacturers Federation for January 1950.

Included in the contents is an article on the rules adopted by the B.S.I. and the N.P.L. for rounding-off of derived dimensions in all standardization work. Other items include details of Technical Panels, and R.C.S.C. and B.S.I. information.

BBC Studio in Dundee

The BBC has recently completed and brought into service a studio in Dundee.

The studio is in the Dundee Corporation's Coldside Library. It is approximately 50 ft. long and 25 ft. wide, and is equipped with standard BBC outside-broadcast equipment capable of handling four microphones. The acoustics were found to be sufficiently good to make any special treatment of the walls and ceiling unnecessary. Engineers from the BBC studios at Edinburgh or Aberdeen operate the equipment.

The Radio Trades Examination Board

Entries for the May 1950 Radio Servicing Certificate Examination, which are not yet complete, have exceeded 240. This increase in the number of candidates is indicative of the wide recognition being afforded to holders of this Certificate. It is interesting to compare the number of candidates who have entered for the examination since 1946.

Date	Candidates
1946	72
1947	94
1948	114
1949	155

For the first Television Servicing Certificate Examination 31 candidates have been accepted.

THE MECHANICAL DEVELOPMENT OF A RADIO ALTIMETER*

by

R. T. Croft† (*Associate*)

SUMMARY

The paper discusses the problems involved in developing a piece of radio equipment, intended for service use, from the circuit designer's laboratory model to a reliable, " Engineered " version suitable for production.

Although the principles involved are applicable to a variety of equipments, it was thought advisable to make the paper specific, and for this reason a radio Altimeter designed by the research staff of the Telecommunications Research Establishment has been taken as a particular example.

1. Introduction

The Altimeter makes use of the fact that a phase difference will exist at the aircraft between the modulation of a C.W. radio signal transmitted downward in a beam and that of its subsequent reflection from the ground (C.W. radar). This difference is, of course, proportional to its height above the reflecting surface at any particular moment.

The developed units of the Altimeter are illustrated in Plates 1-7 while Plate 8 shows a " breadboard " model of the equipment. Plate 9 is a block schematic showing the method of operation.

Service specifications governing radio equipment have wide circulation in industry, and a particularly onerous example is Specification K114,[1] which lays down mandatory conditions under which complete equipments (as distinct from components) must continue to operate.

The levels of ambient temperature, humidity, etc., specified therein are such as to defeat many domestic components unless they are protected from the full effects of the test.

The work which is subsequently described was undertaken in an effort to produce an equipment which would satisfy these requirements, using readily available components, and yet avoiding the complication and costs attendant upon pressurization.

* Manuscript received September 3rd, 1949.
U.D.C. No. 621.318.572.

† T.R.E. Malvern

2. Engineering Requirements

The ultimate aim is to ensure that the equipment will be able to successfully withstand the tests laid down in specification K114, and to this end, the following criteria were formulated.

(*a*) Individual units should be capable of being sealed, to afford maximum protection under extreme climatic conditions. This involves careful attention to heating effects (see section 9). Pressurization was considered, but seemed to indicate, for this application, an unwarrantable increase in weight.

(*b*) The best possible grades of both materials and components should be employed, to guard against deterioration in service and ensure satisfactory operation under elevated temperature conditions (see section 4).

(*c*) The equipment should be robust enough to withstand a large measure of mechanical or thermal shock, either during operation, or more particularly during transit, without the need for excessive packaging measures.

(*d*) Space saving should be important, paying due attention to the physical needs of providing sufficient surface area to radiate the generated heat efficiently.

(*e*) Weight saving should be equally important, although qualified by the preceding remarks.

(*f*) Accessibility is an essential, though an often neglected, feature in radio design, and is of great value when the equipment requires servicing under difficult conditions.

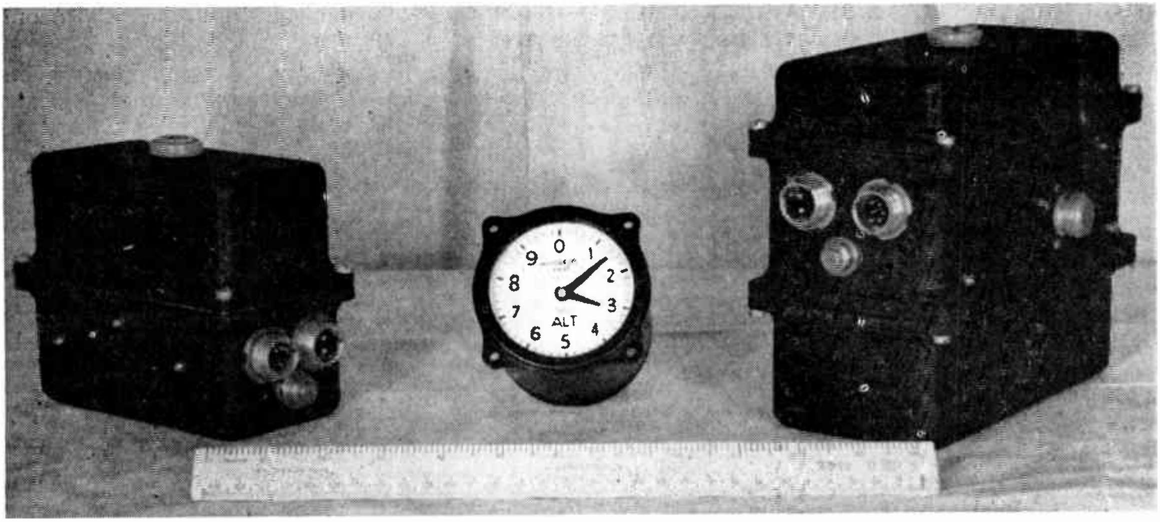


Fig. 1.—Transmitter, indicator and receiver. (The indicator shown is a space model only.)

(g) Within the limitations of available components, every care should be taken to ensure that the equipment will be reliable. This involves siting carefully chosen components efficiently, both from the point of view of the job they are doing, and the temperature conditions prevailing at any particular section of the case.

(h) The amplifier should be capable of mounting in S.B.A.C. racking, which is the standard method of mounting equipment in civil aircraft.

(i) The transmitter and receiver are best installed as close as possible to their respective aerials without regard to the racking system used.

(j) If possible, the use of blower motors and fans should be avoided, because of their inherent unreliability.

3. Materials and Construction

Aluminium alloys seemed the most logical choice for the casings, chassis, etc., of a light-weight equipment, and an alloy such as DTD 606[2] has been found to have good corrosion-resisting properties. Since a limited anodizing action occurs if a surface is scratched, these alloys may be termed self-healing.

Spotwelding of aluminium alloys is often unreliable as a means of construction, especially in marine atmospheres, so light alloy rivets have been used throughout. Aluminium welding, however, can be entirely satisfactory.

It was felt that the use of castings, although desirable as a medium for the type of construction envisaged, would cause too great a delay in the production of prototypes. The design was, therefore, made capable of translation in welded sheet metal; conversion to castings without excessive redesign would be possible if quantity production was contemplated. In this event, weight could be further reduced by the use of a zirconium/magnesium alloy.

Contact potentials between dissimilar materials have been carefully considered, Inter-Service Specification RCS/1000[3] gives a list of these potentials for various metals against an arbitrary element. A contact potential exceeding ·25 volt is liable to set up voltaic action under moist conditions, leading to corrosion of the more negative material.

The gauges of material used are somewhat heavier than are usually encountered in radio design, for various reasons, notably the following :—

(a) The shape chosen for transmitter and receiver (a box with large radii on all corners) is effective in resisting deformation by falling on corners, etc., especially if the wall is reasonably thick.

(*b*) If the units are mass produced it is intended that the shells should be die castings. Since it is not usual to produce die castings with less than $\frac{1}{16}$-in. wall thickness the transition is simplified.

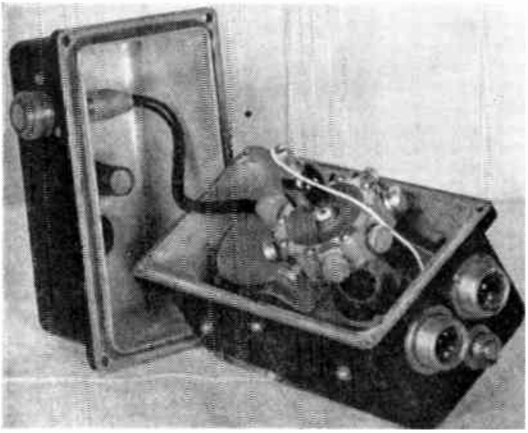


Fig. 2.—Transmitter, showing internal construction.

(*c*) It is possible to increase the area of cut-outs in chassis parts if the metal is reasonably thick, and cutouts improve natural air circulation and thereby reduce weight.

(*d*) If a box is hermetically sealed at sea-level, it will contain air at normal atmospheric pressure. If the box is flown to a great height, the ambient pressure will be reduced, and a pressure will be exerted within the box corresponding to 14·7 lb., minus the ambient pressure at the given height. The container must, therefore, be either :—

(i) Strong enough to withstand this pressure.

(ii) Be fitted with a safety valve.

(iii) Possess an expansion device similar to that fitted to large transformers.

Suggestion (ii), if adopted, means that on return to sea-level a negative pressure exists within the box unless another valve is fitted in opposition to the first. This in turn would allow the infiltration of moist air at each landing so destroying the original advantage of sealing. It is possible to arrange a dessicating agent in the inlet valve, but this would have to be large, and probably require frequent reactivation. A safety valve would therefore be rather complicated.

Suggestion (iii) could be made effective, but the expansible volume would need to be greater than the volume of the units themselves. A development might be the provision of an air-tight moulded bag of large dimensions into which all the units in a rack might exhaust their air and redraw as necessary.

Suggestion (i) therefore appeared the most economical approach, and in small can sizes this view is undoubtedly justified, in the author's opinion, as the slight weight increase would be more than offset by the weight and cost of air valves or breathing devices.

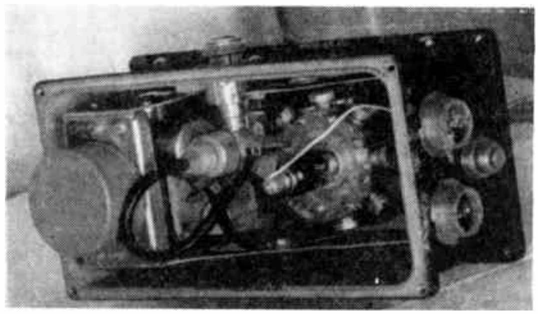


Fig. 3.—Receiver, showing internal construction from topside.

4. Components

The Mk. 4 miniature range of plugs and sockets has been used on the transmitter and receiver. They are capable of being sealed, and are of low weight and small dimensions.

The S.B.A.C. range of back plugs and sockets is used for the amplifier, since these are standard for S.B.A.C. equipment. They are not strictly sealed, however, and a modified version is required to ensure reliability.

Valve holders are of the ceramic insulation type, although the metal portions of even these components are often regrettably poor in finish. Valve holders made from polytetrafluoroethylene (P.T.F.E.), when available in quantity, should effect an improvement.

RMA16 resistors are used wherever they provide sufficient power dissipation, the high-wattage resistors being of the very reliable wire wound type.

Both potentiometers and multi-way switches are available in midget sealed form and examples of the sealed potentiometers and the switches may be seen on the prototype equipment.

Fixed condensers are all of either the sealed metal tube or ceramic types which are known to give good performances under tropical conditions.

Electrolytic condensers are not considered to be reliable at high ambient temperatures, but two are fitted in the power pack in the coolest positions available.

Dessicating elements of the silica gel cartridge type are fitted to each unit. They contribute a limited drying action, but are included mainly as a means of checking the humidity conditions prevailing within the cases, without the need for frequent dismantling.

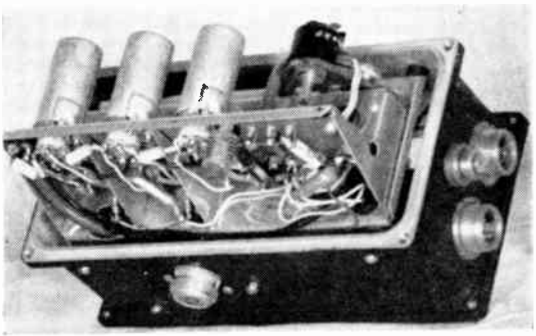


Fig. 4.—Receiver, showing internal construction from underside.

5. Transmitter and Receiver Aerials

The original experimental aerial was of the slotted waveguide type. It consisted of a 12 in. length of 3 in. × 1 in. brass waveguide having a launching probe towards one end and four radiating slots in the bottom face. The weight in this form approximated to 3 lb. per aerial, and a desired increase in performance called for a much larger aerial possessing 20 slots. The weight of such an aerial in brass was estimated at $9\frac{1}{2}$ lb.

Calculations were made for an aerial to be constructed from silvered polythene strips assembled in a metal frame. The cross-sectional area of a waveguide is reduced when air dielectric is replaced by polythene; and the necessary transverse slot spacing was to be maintained by slips of expanded perspex. The estimated weight of this design was 2·77 lb., showing considerable saving in weight but becoming mechanically complicated.

A satisfactory method of fabricating thin aluminium sheet into a large slot aerial was therefore sought, and the result may be seen in Plate 7. The problem involved is to weld, or otherwise continuously join, the metal edges of the guides without heat distortion causing dimensional inaccuracies outside the allowable tolerances. This is achieved by raising flanges on thin sheet waveguide sections, and edge welding away from the close tolerance sections. The weight per aerial, air filled, is 1·125 lb.

It was intended that this aerial should be filled with expanded rubber for tropic proofing, but the material was found to give random attenuation owing to fairly large air pockets within the section, so that recourse was made to sealing the slots with polythene film.

6. Transmitter and Receiver

It was necessary to consider these units jointly since they were to be similar in many respects; notably, their location in aircraft was to be as close to their respective aerials as possible, and they would therefore be inaccessible, etc. Physically the receiver design is like that of the transmitter, with the addition of a crystal mixer, and a head amplifier.

An account has already been given of the choice of shape for these units and they will be seen to take the form of rectangular boxes (Plates 1, 2, 3 and 4).

The transmitter is approximately 6 in. × 3 in. × 5 in. high and is split into two sections. It weighs 2·25 lb.

The receiver is 8 in. × 4 in. × 7 in. high and split into three sections. It weighs 4·5 lb.

The method of sealing the case joints throughout is to have a stout flange welded to each half of the casing. One flange contains a groove, into which a square section Perbunan sealing gasket is pressed, and the inner wall of this groove (it is actually the box wall itself) stands proud to act as a locating spigot when assembling the halves.

The case wall of the other half also stands proud of its flange, and acts as a knife-edged rim pressing into the gasket when the four screws are inserted. The surface width of the gasket is twice the thickness of the rim, and this fact, together with the locating effect of the

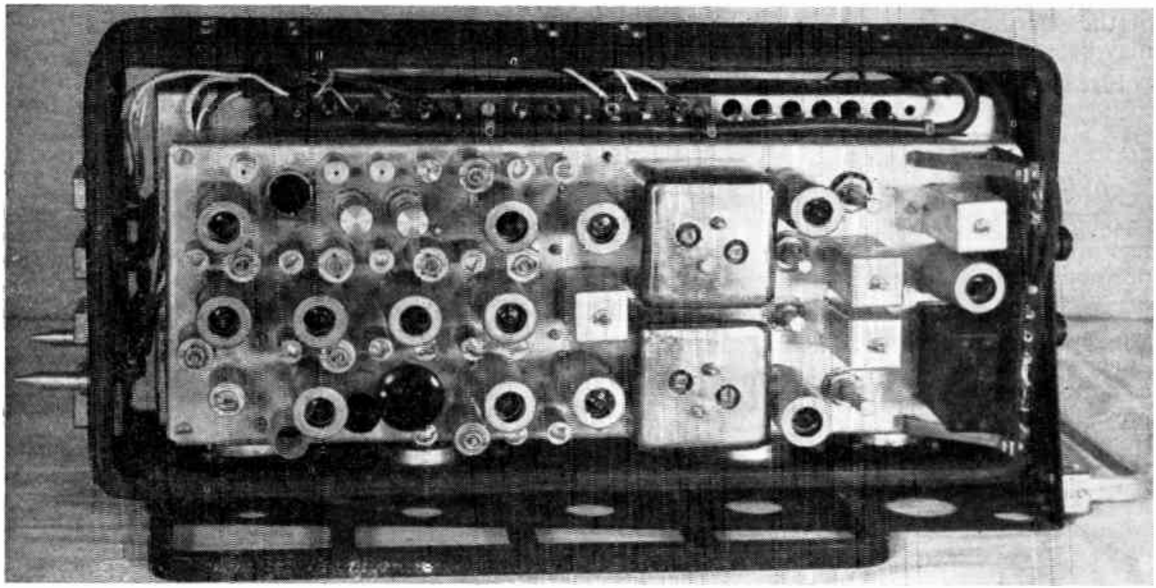


Fig. 5.—Amplifier, showing (from left to right) 1st and 2nd I.F. stages, phase discriminators and crystal-controlled modulator.

spigot, ensures accurate making and breaking of the seal.

The chassis is suspended from the casings by means of Silentbloc Rod type anti-vibration mountings. This avoids the necessity of providing spring mounting trays in the aircraft, and has proved useful in combating transit shocks.

Both units contain a klystron oscillator ; the receiver also has a crystal mixing chamber as the first I.F. followed by a three-stage head amplifier. The filament transformers used are interchangeable, and are lightweight, midget toroids. Their cases measure less than 2 in. dia. $\times$ $\frac{7}{8}$ in. high and may be rated at up to 30 VA.

In the receiver, the top and bottom sections are merely covers, while the centre section mounts the plugs and sockets with the chassis suspended from it. At one end is a screw-on cover which provides access to the crystal holder in the event of crystal replacement becoming necessary.

The anode circuits are screened from one another by being sited in the well of a stepped deck, and by stiffening screens. All adjustments may be made from one side of the unit.

The input circuit to the head amplifier is contained on a tagboard adjacent to VI, but intervalve components are small enough to be placed directly in position.

The original intention was to combine these units with their respective aerials. This offered a neat method of construction, when the aerial was of modest dimensions. A receiver was built along these lines in which the overall size in plan was little greater than the aerial which formed its backbone. A small deck mounted the head amplifier, and was arranged to hinge outward for servicing. The whole was suspended from its casing by means of Silentbloc mountings arranged with their axis at 45 deg. to the vertical. This brought the rubber nominally in shear whatever direction of thrust was induced by a mechanical shock.

The great increase in aerial size later found necessary rendered this scheme obsolete.

7. The Amplifier

No existing box, designed for fitting S.B.A.C. racking, had been produced as a sealed unit.

A difficulty is encountered due to the fact that this system of racking calls for back mounting plugs on the unit, and it will be readily seen that it is not possible to use the standard method of inserting a front panel and cantilever chassis into a sealing can when that sealing can carried the input plugs on its rear face.

It is suggested that the difficulty can be overcome in several ways :—

(a) By designing a connector for the rear of the can which could plug into the rack sockets and would also act as a socket to a further plug mounted at the end of the cantilever chassis, i.e., tandem plugs and sockets.

(b) Mounting the standard plug item on a detachable sealing plate at the rear of the can. This would involve releasing the front panel seal and the back panel seal before withdrawing the unit from the case. It has a serious disadvantage in that manufacturing tolerances over the length of a chassis and a sealing can may render the fitting and sealing unreliable.

The overall dimensions of the box are 9 in. × 8 in. × 16 in. long (see Plates 5 and 6), the total weight being 15 lb.

The centre section provides a strong shell supporting a central chassis web through Equiflex spring mountings. The rear face carried the back mounting plugs and the front face the co-axial sockets.

A sheet metal channel section is edge welded to each side of the shell, forming troughs into which square-section perbunan gaskets are pressed. Into these, the two side cans were embedded.

The cans are drawn up on rods passing from one can to the other, to ensure equal pressure on both gaskets.

One edge of the channel section stands proud of the shell so as to give definite location vertically and longitudinally to the cans when they are being fitted.

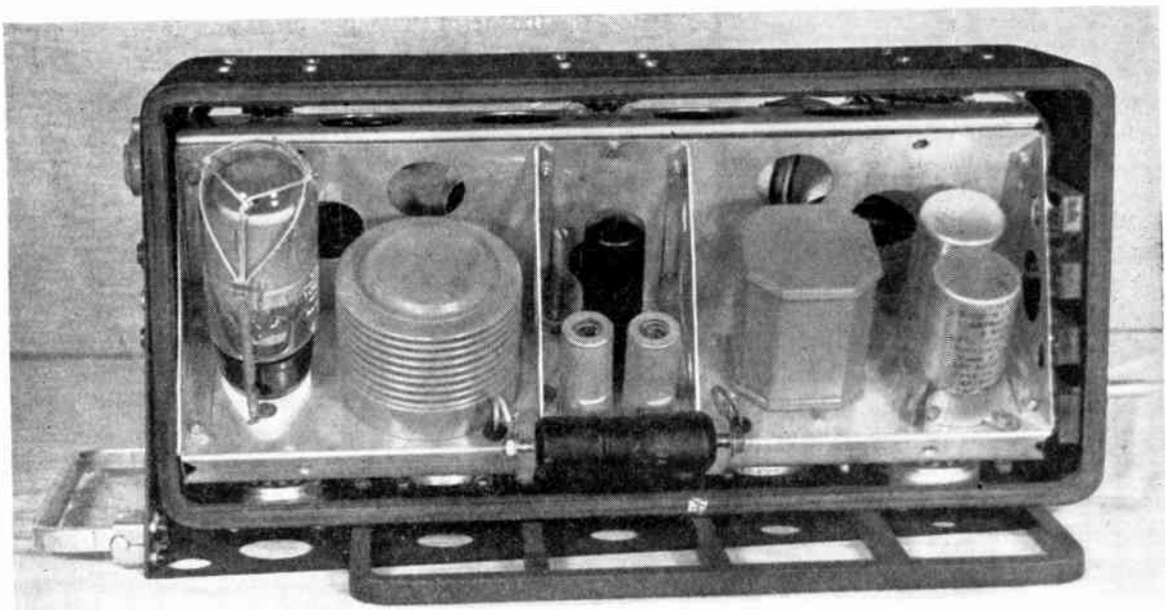


Fig. 6.—Amplifier, power supply for complete equipment.

(c) By dividing the unit into three sections in a vertical plane and mounting the plugs in the centre section. This has the disadvantage of increasing the length of seal required, but improves accessibility and makes it possible to use standard S.B.A.C. back plugs and sockets. This last alternative was chosen for the amplifier unit.

The amplifier, modulator, and phase discriminator circuit element are all mounted on a chassis deck which is easily detachable from one side of the central web. Supplies are taken from this web to the chassis by means of two multi-way tag strips bridged across by a safety pin type of jumper which springs off each tag when touched by a hot iron. Alternatively, a multiplug and socket may be used.

A switched monitoring point is provided which makes 11 test points readable without disturbing the chassis.

The opposite side of the central web is used to support the power pack. This is reasonably standard in its present form, although the transformer is interesting in that it is of the silicone filled type and is able to work at a high operating temperature thus allowing a saving in weight and size. It is fitted with an expansible metal bellows to allow for the high expansion of silicone fluids.

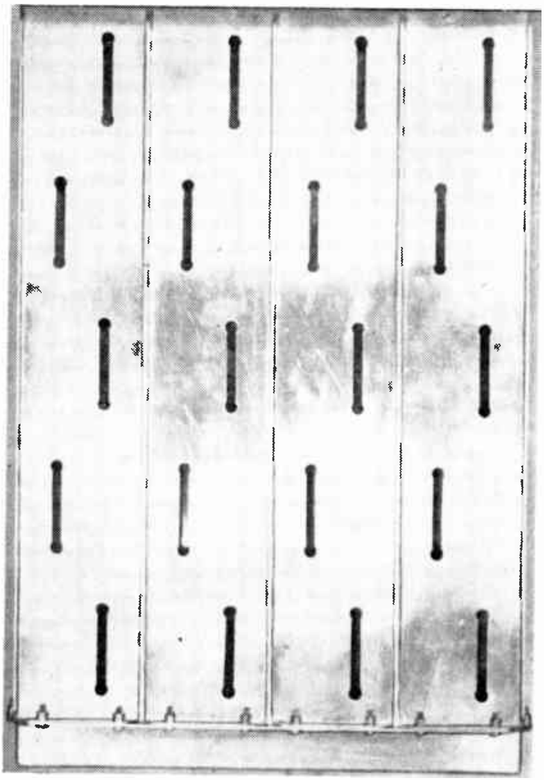


Fig. 7.—" S " band 20-slot directional aerial.

When an S.B.A.C. unit is mounted in its rack it is supported by the bottom edges of :—

(*a*) The front panel.

(*b*) The dust cover.

With this type of construction it is undesirable to allow the side cans to support the weight of the unit, so a grid is fastened to the underside of the centre section to transmit the load to that portion. A false front panel is fitted to carry the lifting handle and the rack fixing elements.

8. The Indicator

A suitable phase meter is already manufactured in this country and is a slight modification of the type used in a proprietary navigator instrument. The weight is approximately 1 lb.

9. Heating Effects

This is not a subject which can be definitely resolved by mathematical means, but there are certain formulae which may be used to provide an indication of likely temperature rises under given conditions.

Such a formula is given below, and has been used as a basis for " temperature stressing " this equipment. (The author is not aware of the term " temperature stressing " being in current use but it bears a direct analogy to the more normal mechanical and electrical stressing used in connection with safe loads in those fields.)

The values obtained are not actual temperature rises at any given point, but represent an average value.

Higher témperatures are more likely to be encountered at the top of a unit than lower down. The gradient will vary according to the position of local heat generation, obstructions, and external cooling effects.

Conclusions may be drawn regarding the siting of both local hot spots and of items particularly susceptible to excessive heat, and an endeavour may be made to forecast a particular thermo-syphon action by the careful siting of these components.

A figure of 129 milliwatts per sq. in. (20 mW/sq. cm.) was taken as a safe basis for temperature stressing the equipment and the unit sizes conform within the limits seen below.

(*a*) Transmitter losses.

Klystron heater	=	5·6 watts
Klystron H.T. Drain	=	8·4 watts
		14·0 watts

Surface area = 90 sq. in.

$$\therefore \text{Stressing} = \frac{14 \times 10^3}{90} = 155 \text{ mW/sq. in.}$$

(*b*) Receiver losses.

Head Amp. Heaters	=	5·67 watts
Head Amp. H T. Drain	=	4·20 watts
Klystron Heater	=	8·40 watts
Klystron H.T. Drain	=	5·60 watts
		23·78 watts

Surface area = 186 sq. in.

$$\therefore \text{Stressing} = \frac{24 \times 10^3}{186} = 129 \text{ mW/sq. in.}$$

(*c*) Amplifier losses.

Amplifier Heaters	=	20·79 watts
Amp. H.T. Drain	=	15·76 watts
Power Section Losses	=	37·10 watts
		75 watts approx.

Surface area = 688 sq. in.

$$\therefore \text{Stressing} = \frac{75 \times 10^3}{688} = 109 \text{ mW/sq. in.}$$

(*d*) A forecast was made, using a formula taken from an M.I.T. report on heat dissipation, of the probable temperature rise in these units.

This formula is given :—

$$q = \frac{\cdot 002178A}{L^{\frac{1}{4}}} \Delta t^{1.25}$$

for rectangular cans up to 24 in. high.

where q = Power input, watts
A = Area of can, sq. in.
L = Height of can, in.
Δt = Temperature difference, °C.

(*e*) Transmitter

q = 14 watts.
A = 90 sq. in.
L = 6 in.

$$\therefore \Delta t = \sqrt[1\cdot25]{\frac{14 \times 6^{\frac{1}{4}}}{\cdot 002178 \times 90}} = 43^\circ\text{C rise above ambient.}$$

(*f*) Receiver

q = 24 watts.
A = 186 sq. in.
L = 8 in.

$$\therefore \Delta t = \sqrt[1\cdot25]{\frac{24 \times 8^{\frac{1}{4}}}{\cdot 002178 \times 186}} = 40^\circ\text{C rise above ambient.}$$

(*g*) Amplifier

q = 75 watts.
A = 688 sq. in.
L = 16 in.

$$\therefore \Delta t = \sqrt[1\cdot25]{\frac{75 \times 8^{\frac{1}{4}}}{\cdot 002178 \times 688}} = 35^\circ\text{C rise above ambient.}$$

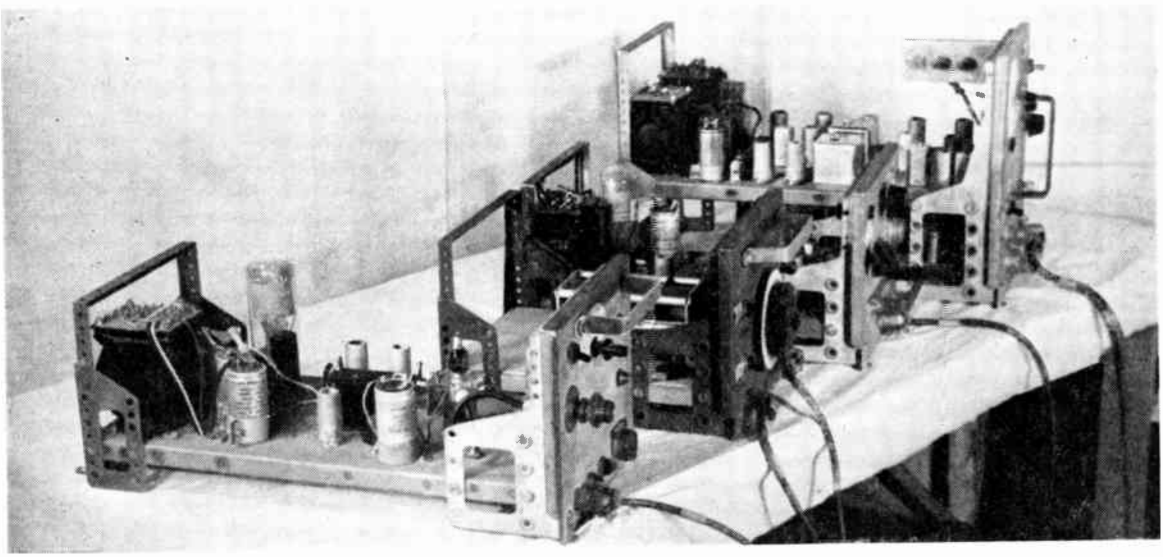


Fig. 8.—Laboratory equipment showing (from left to right) transmitter, phase shift simulator, receiver and amplifier. (Each unit has separate power supplies. Amplifier using phase discriminators designed for " engineered " version.)

If it is assumed that a temperature stressing of 20 mW/sq. cm. will cause a mean temperature rise of 40 deg C above ambient, it may be said that strata of air should exist within the unit ranging from $(40 - X)°$ to $(40 + X)°$ above ambient where X° is not greater than 40 deg C.

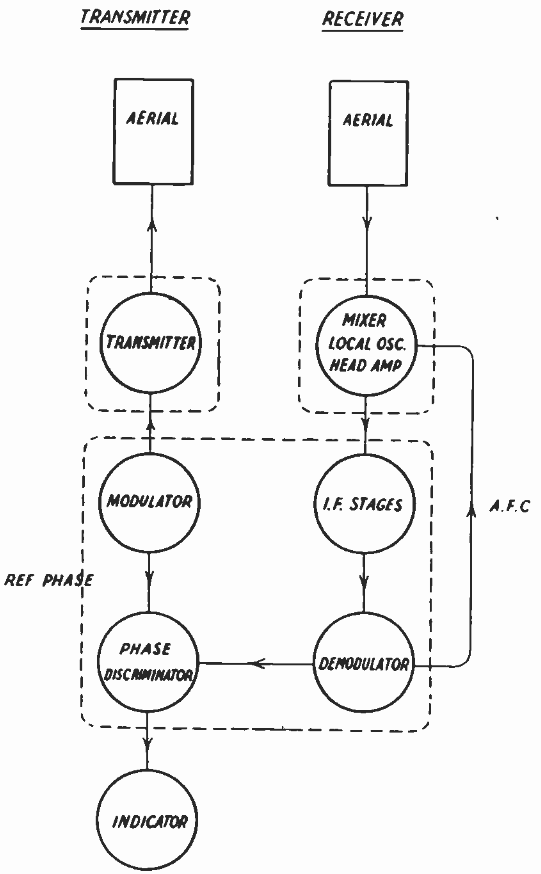


Fig. 9.—Block circuit diagram.

In this equipment, therefore, one would expect a rise above ambient of almost 80 deg C at the hottest point. This figure has been confirmed by subsequent measurements.

The standard finish for airborne equipment is matt black and this is convenient as an aid to heat dissipation. It must not be assumed that the *colour* has any direct bearing on the matter, however, since many other colours have the same emissivity ratio to a theoretical black body.

A list of various colours taken from Food Investigation Special Report No. 9, 1931, with their emissivity factors expressed as a ratio to dead black, is given below :—

	Ratio to Dead Black
Black Paint, Black Enamel, Dark Green Paint .. .. White Paint, White Enamel, Dead Black	1
Aluminium Paint	·7
Aluminium Paint rubbed down with fine sandpaper	·8
Dull Copper	·15

It is the matt surface texture which is of importance. Note that the cases should be painted inside and out foı maximum effect.

10. Tests

Two sets of equipment were built and priority was given to one of these in order to provide experimental flying data, both from an Altimeter and from an engineering point of view.

The transmitter, receiver and indicator have now been used for some hundreds of hours both in the air and in the laboratory, including casual transportation by road and hand, with only three reported cases of unserviceability.

These faults were investigated as they occurred and the results are given to emphasize the points to be watched in this type of construction.

Fault (1) was due to the miniature toroidal transformer becoming loose and breaking its primary connection. A locking washer, to prevent screws being loosened by vibration, prevented recurrence of this trouble.

Fault (2) was traced to shorting of the klystron oscillator which has risen from its socket sufficiently for its top cap to reach the case when the sprung chassis was at the top of its movement. This was cured by the fitting of a sprung retainer to the roof of the box.

Fault (3) was found to be due to a noisy joint between the screened cable, carrying the head amplifier output, and its sockets.

This was due to the flexing of the single strand inner conductor at the soldered joint during the oscillation of the chassis, causing hardening and impending fracture.

All cables had been provided with sufficient slack to avoid a direct pull on the termination but, owing to the inherent stiffness of the screened cable used, it became necessary to clamp the cable to the fixed portion of the unit at a point just before where the outer covering was stripped back for soldering.

Rather less than ordinary care was taken with these units, since they are light and small enough to be thrown from hand to hand, or into the back seat of a car for transportation. This has been deliberately done on many occasions with no ill effects.

The amplifier, transmitter and receiver have also been immersed in 18 in. of water for periods of 15 minutes with no ill effect and without leakage.

11. Conclusions

(*a*) Hermetic sealing is justified, provided sufficient attention is paid to heating effects to ensure reliability. In a C.W. system, such as the radio altimeter, it is doubly justified since it reduces leakage of energy between the transmitter and receiver within the aircraft to a minimum.

(*b*) The use of high quality components, materials and construction are necessary even in a sealed case, since the seal will have to be broken on occasions and moisture will enter. Sealing, coupled with miniaturization, often increases working temperatures to higher levels than domestic components can usefully withstand, and forced cooling is then necessary.

(*c*) The incorporation of self-springing devices may be useful in small unracked equipment provided that adequate clearances are maintained, that resonance is unlikely to occur, and that no strain is imposed on soldered connections due to cable flexing. In heavier equipments, however, the movement required to safely absorb shock loads is such as to considerably increase the physical space required ; and shock absorption is better left to adequate packing. It must also be remembered that the shock mountings will be in series with those on the mounting tray if the units are racked, and this may modify the resonances encountered under vibration.

(*d*) The use of back plugs and sockets is difficult in sealed equipment for reasons stated.

(*e*) Whether a pressurized or merely sealed container is chosen, and whether safety valves, large-scale air dryers, etc., are fitted, can be determined only after careful consideration of the equipment to be designed, and its probable field of use.

(*f*) The design of equipment which will operate successfully under pan-climatic conditions and yet, by avoiding pressurization, forced convection, etc., will remain light and reasonably economical to produce and maintain is definitely practicable.

12. Acknowledgments

The author is indebted to the Chief Scientist, Ministry of Supply, for permission to publish this paper. Crown copyright is reserved, and it is reproduced with the permission of the Controller of His Majesty's Stationery Office.

13. References

1. W.T. Board Specification, No. K.114, Climatic and Durability Testing of Service Telecommunication Equipment, H.M.S.O. (1946).
2. D.T.D. Specification No. 606, half-hard aluminium alloy sheets and coils, H.M.S.O.
3. Inter-Service Specification RCS/1000 General requirements for Service Telecommunication Equipment, Radio Standardization Committee, Ministry of Supply (1947).

GRADUATESHIP EXAMINATION, NOVEMBER 1949

PASS LIST

A total of 251 candidates entered for the whole or part of the November 1949 Graduateship Examination. This list contains the results of all the home candidates and those overseas candidates whose results were available on January 23rd, 1950. The results of the remaining candidates will be published in the March Journal.

Eligible for transfer or election to graduateship or higher grade of membership.

The following candidates have passed the entire examination, or having previously passed or been exempt from part of the examination have now passed the remaining subject(s).

ALDRED, Arthur Bernard (S)	Great Yarmouth
BECKLEY, Norman James (S)	Wembley, Middx
BELL, John Ramsay (S)	Forfar, Angus
BELL, Ronald (S)	London, N.16
BOYCE, Desmond Archibald (S)	Oxford
BROVIG, Aanen (S)	Ilford, Essex
BUDGEN, Ronald Pallister (S)	East Grinstead, Sussex
CURRAN, Arthur Joseph (S)	Dublin
FORBES, Frank (S)	West Croydon, Surrey
FRANCIS, Idris Alan (S)	Machynlleth, Wales
GILMOUR, Hugh Morton (S)	London, S.W.9
GILSON, Thomas John (S)	Tullamore, Eire
HEALY, Geoffrey Noel (S)	Blackburn, Lancs
HINTON, Maitland Joseph Gresley (S)	London, S.W.1
LEE, Fong Lim (S)	London, S.W.5
MARTIN, Michael Blackmore (S)	London, W.4
NICHOLLS, Noel Charles (S)	London, W.4
ODELL, Harold Arthur (S)	Tripoli
OWEN, Clifford RusselWilliam, B.Sc.	Barking, Essex
PANNELL, Peter William (S)	Ilford, Essex
PENDHARKAR, Neelkanth Gopal, B.Sc. (S)	Poona, India
PEARCE, Richard John (S)	Shrewsbury
QUIRK, William James (S)	London, S.W.13
RICHARDSON, John Thomas	Treforest, Glam
RIESEL, Herbert (S)	Tel-Aviv
ROBERTSON, Albert William (S)	London, N.W.7
SAYERS, John Francis	Barnehurst, Kent
SEARS, John (S)	London, N.6
SMITH, Frederick George (S)	Bristol
TAYLOR, Douglas Raymond	Hull
TOWELL, Roger Percival (S)	East Molesey, Surrey
TWIVEY, Derrick, B.Sc. (S)	London, W.5

The following Candidates passed Part I only

BAKER, Charles Thomas (S)	London, S.W.6
EVANS, Hugh Maitland (S)	Newbridge, Mon
KENDALL, Ernest Walter (S)	London, W.4
LILLEY, Robert James Hartley (S)	Invercargill, New Zealand
PAYNE, Stanley Frederick (S)	Cross Keys, Mon
RODMELL, Edward Cripps (S)	Chipping Norton, Oxon

The following Candidates passed Part II only

ANANTHARAMIAH, A. V. (S)	Bangalore, India
BHASIN, Rajinder Nath (S)	Pathankot, India
FAUTLEY, Raymond Francis (S)	Ewell, Surrey
HANCOCK, Harry James	Leek, Staffs
HARDY, John Alfred (S)	Glasgow, S.1
KRUYSHAAR, Willem (S)	Pretoria, S. Africa
LLOYD, Reginald Francis (S)	Korweinguboora, Australia
McDERMOTT, Harry Robert (S)	Darlington, Co. Durham
MENSAH-BROWN, Joseph Paschal (S)	Cambridge
PEGRUME, Peter Hillyard (S)	Whitton, Middx
PORTER, William Edmund (S)	Nottingham

Candidates who have passed Part II only (continued)

ROSS, James McGlashen (S)	Carshalton, Surrey	WIFFEN, Montagu Arthur (S)	Hawke's Bay, New Zealand
ROW, Edward Francis (S)	Birmingham	WOODFORD, Paul Ivor Keith (S)	Trieste Force
STOTT, Donald Geoffrey (S)	Ashford, Middx		
TAYLOR, Alan Parsonage (S)	Chislehurst, Kent		

The following Candidates passed Part III only

CHAND, Satish, B.Sc. (S)	Delhi	STONE, Frederick John (S)	Exeter
EAGLES, Dennis James (S)	London, W.3	THOMAS, Newton Vedanayagam, B.Sc. (S)	Bangalore

The following Candidates passed Part IV only

BUNDY, Douglas Ernest (S)	Farnborough, Hants	MELLOR, Jim (S)	Dover
CAMERON, Peter McDougall (S)	London, N.22	MORGAN, Stephen Lascelles (S)	Johannesburg
DA GUPTA, Dibyendu, B.Sc. (S)	Patna, India	PROWSE, Michael John (S)	Staines, Middx
DE SILVA, Appu Hennedige (S)	Moratuwa, Ceylon	RAGHUNATH RAO, Arani Laxmi, B.Sc. (S)	Bezwada, India
DOWNEY, Terence Brian (S)	London, W.7	ROBB, Ronald Laidlaw (S)	Harrow, Middx
HALES, Anthony William (S)	Nuneaton, Warwicks	SELINGER, Cyril S. (S)	London, N.W.6
HOPKIN, Peter Roy (S)	Hadleigh, Essex	SINCLAIR, David (S)	Wirral, Cheshire
JOSLIN, Charles Albert Frederick (S)	Chelmsford, Essex	SNOWSILL, Alan Harold (S)	London, S.E.21
		EMBE, Sakharam Bhaskar, B.Sc. (S)	Gwalior, India
		TWIGG, George Sydney (S)	Wembley, Middx

The following Candidates passed Parts I and II only

DUNN, John William (S)	Prestwick, Ayrshire	SHIPGOOD, Frederick John (S)	North Shields
REID, Andrew Michael (S)	London, W.3	STEVENSON, Roy Neville (S)	London, N.16
ROBINSON, John (S)	Hull	TITHERADGE, John Percival (S)	London, S.W.16

The following Candidate passed Parts I, II and III only

REANEY, Donald (S)	Liverpool

The following Candidates passed Parts I, II and IV only

BRANDT, Arthur Raymond Anthony (S)	Sevenoaks, Kent	FENWICK, William Alfred (S)	Cambridge

The following Candidate passed Parts I, III and IV only

HARRIS, John Edward (S)	Dublin

The following Candidates passed Parts II and IV only

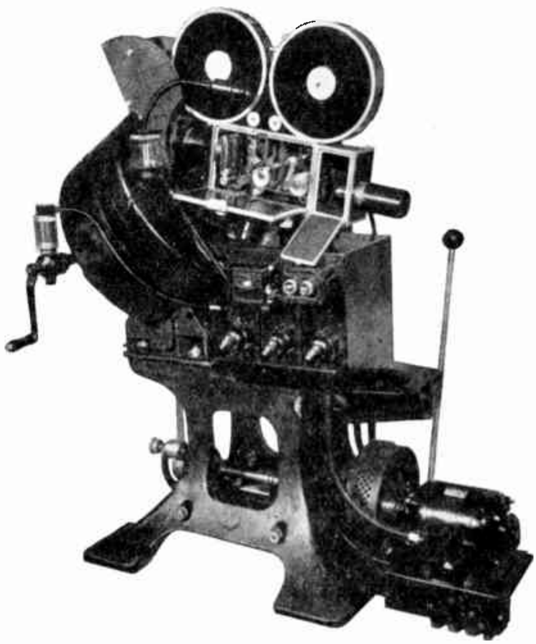
GOULD, Robert Alan (S)	London, S.W.6	SMAILES, George (S)	London, W.1

TELEFILM EQUIPMENT

The new telefilm equipment at Alexandra Palace and the television studio space recently acquired at Shepherds Bush were inspected by a representative of the Institution last month at the invitation of the BBC.

The telefilm equipment is a major addition to the technical facilities of the BBC television service and it will no doubt form the basis of the interchange of television programmes between the Television Broadcasting Organizations in the various countries. The system in operation was devised and developed entirely by BBC engineers and was first used in November, 1949.

The recording installation consists of duplicate recording cameras facing a pair of picture monitors. The film is drawn continuously through the camera and the use of a shutter is avoided by a system of rotating and rocking mirrors which keep the image stationary relative to the film (see photograph). In this way a succession of television images is formed on the film as it passes through the gate, the brilliance of each image rising from zero to a constant intensity over the central part of the gate, finally fading out to zero on reaching the bottom. With this arrangement, each television picture consisting of two interlaced frame scans is fully recorded on one frame of film. To eliminate the line structure which would possibly result in an interference pattern when the film was subsequently scanned for transmission, the scanning spot of the monitor tube is deflected in a vertical direction by a sine wave displacement, the frequency of which is several times the maximum detail frequency of the system.



One of the cameras used for recording television programmes on film. Inside the casing at the front are rotating mirrors which cause the image of the television pictures to follow the film in its downward progress through the gate of the camera.

The film scanning equipment at present in use is also relatively new. The film runs continuously through the projector and each frame is exposed to two scanning images projected from a cathode ray tube in turn by a rotating shutter. The light passing through the film is focused on a photo multiplier tube where the two scanning images give an interlaced signal. These equipments give a very much better television reproduction of film than has previously been possible, especially in respect of definition and tonal gradation.

The recording of sound on films is also of prime importance and it is most important that the time taken to record commentaries on the films should be reduced to a minimum. For this reason experiments are being carried out in recording the sound on a magnetic strip on the film instead of photographically and this cuts the time of production from six hours to one hour.

The Lime Grove Studios at Shepherds Bush, are the solution to one of the BBC Television Service's biggest problems, that of studio space. The new studio space available will be 28,100 sq. ft. in five studios compared with 3,980 sq. ft. in the two studios at Alexandra Palace. It is hoped that the first studio will be in operation in late April or early May this year.

At the conclusion of the tour, the Controller of the B.B.C. Television Service stated that the number of licence holders in Great Britain is in excess of 250,000, and the estimated annual production of television receivers for 1950 is 480,000.

TRANSFERS AND ELECTIONS TO MEMBERSHIP

Subsequent to the publication of elections to membership which appeared in the January issue of the Journal, a meeting of the Memberhip Committee was held on February 7th, 1950. Twenty-five proposals for direct election to Graduate or higher grade of membership were considered, and eighteen proposals for transfer to Graduate or higher grade of membership.

The following list of elections was approved by the General Council : Nineteen for direct election to Graduate or higher grade of membership, and fourteen for transfer to Graduate or higher grade to membership.

Direct Election to Associate Member

Jago, Patrick Desmond	London, N.W.5
Lunt, John, B.Eng.	London, W.5
Sexton, Henry George Arthur, Major	Belmont, Surrey
Theodossiades, Dimitri Fintias, W/Cdr.	Athens, Greece
Travers, Roy	Nottingham
Owen, Clifford Russell William, Instr./Lt., B.Sc.	Barking, Essex
Taylor, Douglas Raymond	Hull, Yorks.

Direct Election to Associate

Cuppleditch, George Alfred	Louth, Lincs.
Elvy, Montague Terrell	Stone, Staffs.
De Hass, Jacobus	Bogota, Colombia
Morgan, Hubert Bernard Whitehead	Liverpool
Norton, Leslie Herbert,	Chadwell Heath, Essex
Scott, William Ernest George	Sutton, Surrey

Direct Election to Graduate

Cattermole, Kenneth William	Ilford, Essex
Couroussopoulos, Andrew	Athens, Greece
Dobson, Frank	Maidstone, Kent
Donovan, Timothy Denis	Marazion, Cornwall
Edwards, Norman Edward	Cranham, Essex
Marklew, Stanley Sidney Peter	Birmingham

Transfer from Associate to Associate Member

Rao, Sattiraju Sambasiva	Bombay, India

Transfer from Student to Associate Member

Long, William Joseph	Sidmouth, Devon

Transfer from Student to Graduate

Beckley, Norman James	Wembley, Middlesex
Bell, John Ramsay	Angus, Scotland
Bell, Ronald	London, W.16
Budgen, Ronald Pallister	E. Grinstead, Sussex
Gilmour, Hugh Morton	Weston-Super-Mare
Lindsey, Peter William	King's Lynn, Norfolk
Martin, Michael Blackmore	London, W.4
Nicholls, Noel Charles	London, W.4
Pannell, Peter William	Ilford, Essex
Pintoff, Edward Ephraim	London, E.5
Sail, Jagdish Parshad, B.A.	Jullendher, India
Towell, Roger Percival	E. Molesey, Surrey

Proposals from the Union of South Africa

A provisional Committee has been appointed as a first step toward inaugurating a section of the Institution in the Union. In accordance with the recommendations of the Professional Purposes Committee, as published in the last Annual Report, applicants in South Africa are now required to submit their proposals to the Secretary of the South African Committee. That Committee will then submit a report to the main Membership Committee, together with the proposal.

This arrangement will be of advantage both to the Institution and to the candidate in ensuring that the support of members will be taken into account in considering the application.

STUDENTSHIP REGISTRATIONS

In addition to the list of Studentship Registrations published in the January issue of the Journal, the following fifty-nine studentship proposals were dealt with at the meeting of the Membership Committee held on February 7th, 1950. The following registrations have been approved and include names not published in the January issue.

Andrews, William Gordon	Liverpool
Anikhindi, Ramesh Ganesh, B.Sc.	Poona, India
Balasubramanyam, T. R.	Madras, India
Bennett, Wilfred Denis	Manchester
Bhatia, Tilak Raj	Delhi, India
Brahm Dev	Jullundeer, India.
Brooking, Kenneth William	Newport, Mon.
Chattopadhyay, Kalyaneswar, B.Sc.	Kewar, Pakistan
Charlton, John Augustus	London, S.W.16
Chopra, Prem Nath, B.Sc.	Simla, India
Clarke, Leonard Roy	Ruislip, Middlesex
Clough, Reginald Arthur	Cheadle, Cheshire
Comfort, Theodore Valentine	Colwyn Bay, N. Wales
Dawes, Ian	Glasgow, S.2
Deshmukh, Ananda Kumar G., B.Sc.	Nagpur, India
Deshpande, Purushsttam D., B.Sc.	Nagpur, India
Donnison, John William	London, S.E.24
Dutta, Rajendra Nath, B.Sc.	Howrah, India
Ellefsen, Angus Olat	Heysham, Lancs.
Gabzdyl, Stanislaw Jan	Eindhoven, Holland
Ghose, Amiya Kumar, B.Sc.	New Delhi, India
Glassbrook, George James	London, S.W.19
Guildys, Jonas Algirdas	Toronto, Canada
Grove, Cyril John	Feltham, Middlesex
Handa, Ram Saroop	Phagwara, India
Hann, Arthur George	Bournemouth, Hants.
Harris, John Edward	Southampton, Hants.
Hart, Francis Sydney	Auckland, New Zealand
Hadjakis, George	Athens, Greece
Harvey, John Copinger	London, N.W.3
Head, Reginald Edward Albert	Wolverhampton, Staffs.
Howard, Jack	Oldham, Lancs.
Jones, Bernard Harley	Ryde, Isle of Wight
Kalyanasundaram, C. R., B.Sc.	Aduturai, India
Kanojia, Babubhai	Dakor, India
Kaskabas, Constantine	Mytilene, Greece
Katrak, Behram J.	Bombay, India
Kawatra, Rajindra K.	New Delhi, India
Kay, John Douglas	Bournemouth, Hants.
Khatry, Krishna Bahadur	Nepal, India
Knight, John Alexander	Vancouver Island, Canada
Kolotouros, Nicholaos	Athens, Greece
Lewis, John Douglas	Tontesbury, Shropshire
McCorkindale, Thomas Ramsey	Ferry Road, Renfrew
Maswood, Ahmad	Karachi, Pakistan
Mundkur, Muralidhar Rao	Bombay, India
Owles, Leonard Eustace	Potchefstroom, S. Africa
Paterson, Anthony Gordon	Fareham, Surrey
Power, Vicent Talbot	Liverpool
Ranjit Singh Chimni, Sadar	New Delhi, India
Reece, Douglas Clagne	Wallasey, Cheshire
Sailendra Nath Das	Calcutta, India
Saini, Sambasiva Rao	Phagwara, India
Thapar, Har Prakash, B.Sc.	New Delhi, India
Wilkinson, William Dinsdale, B.Sc.	London, S.W.16
Williams, Ronald Herbert George	Caerphilly, Glamorgan
Winter, Arthur Lewis	Bristol
Young, John Edwin Percival	Tarporley, Cheshire
Charlton, Leslie (Reinstatement)	Johannesburg, S. Africa