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The Application of Transistors to Sound Broadcasting

by

S.D. Berry, Associate I.E.E.
(Designs Department, BBC Engineering Division)

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BRITISH BROADCASTING CORPORATION

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FOREWORD

THIS is one of a series of Engineering Monographs published by the British Broadcasting Corporation. About six are produced every year, each dealing with a technical subject within the field of television and sound broadcasting. Each Monograph describes work that has been done by the Engineering Division of the BBC and includes, where appropriate, a survey of earlier work on the same subject. From time to time the series may include selected reprints of articles by BBC authors that have appeared in technical journals. Papers dealing with general engineering developments in broadcasting may also be included occasionally.

This series should be of interest and value to engineers engaged in the fields of broadcasting and of telecommunications generally.

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THE APPLICATION OF TRANSISTORS TO SOUND BROADCASTING

SUMMARY

This monograph describes a number of transistor audio-frequency amplifiers and associated apparatus designed for general use in the sound broadcasting services of the BBC. Earlier developmental work is referred to and the system requirements which have influenced the final designs are discussed. The characteristics of some frequently used circuit configurations are considered and a treatment is given of some properties of the input circuits of transistor microphone amplifiers with regard to noise.

1. Introduction

An earlier monograph, No. 26[1] of this series, described a number of audio-frequency amplifiers which had been designed for various requirements of sound broadcasting. This preliminary work was undertaken in order to investigate how transistor amplifiers should be designed to satisfy the necessary high standards and, by experience of apparatus in service use, to obtain information on which later work could be based. Subsequently a programme was started to provide a full range of transistor apparatus with which the sound services would eventually be equipped. Besides the purely electronic design the programme necessarily entailed consideration of the mechanical construction of the apparatus, of bay-mounting methods, and of the future form of studio and other desks with which much of the apparatus would be associated. A preferred method of power supply had also to be decided.

It is the policy of the BBC to have available a standard range of amplifiers which will meet all requirements as far as they can be foreseen. The number of types should be as small as possible but there should be sufficient to avoid the need for redesign or modification when new schemes are planned, although occasions will arise when amplifiers of the standard range will not efficiently meet a particular need and further designs become necessary.

The range must include a microphone amplifier and others for studio desks, control positions, and outside broadcasts; line-sending and receiving amplifiers are required and a variety of miscellaneous requirements must also be met. The new standard range of transistor amplifiers is described below together with two peak programme meters and some other associated apparatus.

2. Amplifier Types and System Requirements

The apparatus considered comprises the following units:

2.1 *Line-sending Amplifier*

A low, fixed gain and a fairly high output power are required for sending programmes to outgoing lines. This amplifier also has a number of other uses where the higher output power is needed.

2.2 *Line-receiving Amplifier*

Although primarily intended to follow the equalizer of an incoming line, it also serves as a general-purpose level-raising amplifier. A fairly high adjustable gain is required.

These two line amplifiers must have a particularly high standard of performance since many are used in tandem on a long programme chain.

2.3 *Microphone Amplifier*

This will be used in studio desks and in outside broadcast assemblies as channel amplifier and also as group and echo amplifier. The primary requirement is for low noise. Small size is also an important consideration; more than sixty are likely to be needed on one of the larger studio desks.

2.4 *Outside Broadcasting Line-sending Amplifier*
Studio Output Amplifier

These amplifiers are designed to terminate outside broadcasting assemblies and studio desks respectively. They resemble each other closely and only the first will be described in detail.

2.5 *Response-selection Amplifier*

The production of some types of light music requires the facility of manipulating the frequency response to obtain a sound quality in the modern idiom. This apparatus enables a rise and fall to be obtained at each end of the audio-frequency spectrum and also a resonance peak to be inserted in order to provide the effect known as 'presence'.

2.6 *Peak Programme Meters*

Two types of transistor programme meter have been developed. The first is intended as the standard P.P.M.; it will be used where accuracy of programme measurement is essential, e.g. at studio centres and before transmitters. The second is a simplified instrument in which accuracy in the reading of programme peaks has been relaxed in favour of small size and low-power consumption. It is intended for use, for example, on outside broadcasts and for echo-channel measurement.

2.7 *Power Supply Unit*

The considerations which have affected the choice of power supply system are discussed later. The preferred method employs a 24-volt stabilized power unit for each small group of amplifiers.

3. System Requirements

The development of transistor electronic apparatus and of modern components and materials have led to the reconsideration of some details of the existing system and, conversely, the growth and enlarging nature of the facilities required for programme production have had a great influence on apparatus design. It is, of course, not practicable to exchange quickly a large existing system for a completely redesigned one; new apparatus must work in harmony with most existing features but must also be capable of fulfilling all foreseeable demands. Some matters which have arisen in this connection are discussed below.

3.1 *Studio Desks and other Control Positions*

A change from valve to transistor electronic apparatus leads to great changes in the design of studio and other desks. It is now possible, by reason of the reduction of size and far lower production of heat, to mount all amplifiers in the desk itself instead of upon an adjacent bay. At the same time there is an increasing demand for additional and more complex facilities, some of which are being supplied on existing valve-operated desks by the addition of separate units. These extra facilities include an increase in the number of input channels, from about ten in earlier designs to the thirty-eight which are envisaged for the larger studio desks of the future, more provision for frequency-response manipulation, and a larger number of echo channels. Remote control of echo-plate characteristics and of tape reproducers is also required. Volume limiting and some forms of automatic volume control are becoming necessary.

Desks of such complexity are practicable only when transistors replace valves as the active elements, and when the more compact faders and other controls that have been developed to meet these requirements are used. Even so, it is already difficult to compress all the necessary controls within the reach of a single operator, while leaving sufficient space for programme scripts.

The mixing of a large number of channels, and other circuit requirements, gives rise to a large total loss which is met by a total available amplifier gain of 163 dB from microphone to output. This is distributed among four amplifiers, the microphone, group, main, and output amplifiers. The first three of these are of the same type, the microphone amplifier described below, which also serves for use on echo and talkback circuits. The fourth is the studio output amplifier, also described later.

3.2 *Internal Wiring of Programme Centres*

A consequence of the expansion in recent years of the scale and complexity of installations has been the growing need to use long runs of small-gauge cable linking studios, control rooms, and switching equipment. The electrical constants of such a cable, together with the capacitance of the selection-switching multiples, can introduce serious frequency-dependent losses when the terminating impedances are 600 ohms, a value which has hitherto been the standard. This can be seen from the fact that a possible link is well represented by a π network having a series arm of 56 ohms in series with $0 \cdot 2$ mH and shunt arms of $0 \cdot 016$ and $0 \cdot 047$ μF respectively. Calculation shows that the frequency error caused by the worst conditions likely to be met could be reduced to a negligible amount by the use of a sending resistance of about 40 ohms. The circuit would be loaded only by the high impedance of the normal bridging amplifiers at the far end in order to avoid a basic loss which would vary with varying lengths of cable.

It was found, however, that the lowest leakage inductance obtainable in a practicable transformer designed to provide an output impedance of the required magnitude was too great to give acceptable results with the reactive load. The resonance of the leakage plus line inductance with circuit capacity produced a rise in response at the higher frequencies of an amount which depended upon the length of cable. Another adverse effect was a reduction of maximum undistorted output at the higher frequencies due to the elliptical load-line presented to the output transistors by the reactive load. These effects were largely overcome and an acceptable performance obtained for all foreseeable conditions by the insertion of resistance padding having a loss of $2 \cdot 7$ dB between the output transformer secondary and the cable load. The resulting output impedance is 55 ohms. With this arrangement the response at 15 kc/s was within $\pm 0 \cdot 25$ dB in all likely conditions of load and the reduction of undistorted output at 10 kc/s was negligibly small. An output circuit of this kind is used for the line-receiving amplifier and the studio output amplifier, both of which work in the conditions described. It is included in Fig. 4 (p. 11).

3.3 *Power Supply*

All apparatus is designed to take power at a potential of 24 volts with the positive pole at earth potential. This voltage was decided upon during earlier experimental work and has been retained as a standard. The factors which have influenced the choice are discussed in Monograph No. 26[1]; briefly, a lower voltage causes some difficulty in relation to the required output impedance of amplifiers and the need for adequate d.c. stabilization, a higher voltage cannot conveniently and safely be used with the majority of available transistors.

Power is supplied by mains-operated stabilized power units, each of which supplies a small group of amplifiers. Other methods have been considered and rejected as a standard, though they will have application in special cases. The two most important of these other methods are: (*a*) the utilization of the 50-volt relay supply with which practically all stations are equipped, and (*b*) the provision of a small mains unit in each individual amplifier. The former arrangement has been used successfully and will be used in the future on small installation where there is no reserve supply at mains voltage; the relay supply is maintained by a floating battery of sufficient capacity to deal with short failures of mains supply. At larger installations an emergency supply at mains voltage is essential for other reasons, and the large power loss caused by the reduction of 50 volts to 24 volts, together with other disadvantages, has led to the rejection of this method for general use. The individual mains unit form of supply will be used for larger, isolated pieces of equipment, such as some test apparatus, particularly where they may be required in portable form. The necessary increase in apparatus size and the need for stringent precautions against hum induction in high-gain audio-frequency amplifiers rule out this method for general use, although a form of it is used for some television amplifiers, where the above disadvantages are not so marked.

4. Some Aspects of Circuit Design

It was found during the design work leading to the present range of amplifiers that certain types of circuit configuration were well adapted to the performance required and were in consequence used in a number of the amplifiers. Two such arrangements are (*a*) a d.c./a.c. feedback pair, and (*b*) the 'Darlington' or 'Super-alpha' connection of two transistors.

4.1 *The Feedback Pair*

The basic circuit of the feedback pair is shown in Fig. 1; it is used mainly as a level-raising block at the input of an amplifier, but the general principle has other applications. The total resistance in the emitter of TR2 can be arranged to supply d.c. feedback to the base of TR1, series- or shunt-connected a.c. feedback, or a combination of these. The a.c. feedback is current-derived when the load is connected between the collector of TR2 and the common line, as is usually the case, but may be made voltage feedback when it is possible to connect the load between collector and emitter, as when a transformer is used. The circuit may be designed to give a high degree of d.c. feedback, thus stabilizing the operating points of the transistors securely, and a wide variation in the proportions and amount of series- and shunt-connected a.c. feedback in order to establish a required input impedance with stability and to provide the total feedback needed.

4.2 *The 'Darlington' Pair*

Two transistors connected in the manner of the output pair in Fig. 4, for example, provide a very convenient

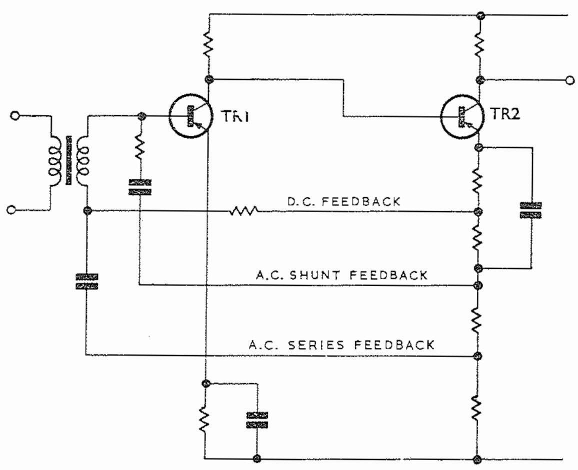

Fig. 1 — *A Feedback Pair—the Basic Circuit*

means of obtaining a high gain stage, particularly where a high input impedance is also required. The pair behaves as a single transistor with a current gain approximately equal to the product of the current gains of each, and an input impedance which may be of the order of several tens of thousand ohms, due to the feedback voltage existing across the base/emitter region of the second transistor appearing in series with the input circuit. The arrangement is usually used for these amplifiers in a push-pull output stage. The high initial gain, which is obtained without the complication and phase shift of the components necessitated by a normal tandem connection, enables adequate feedback to be applied while retaining a useful stage gain. The input impedance, in these conditions, becomes a large fraction of a megohm.

Both the feedback pair and the 'Darlington' pair are discussed in Reference 1, in which an analysis of the d.c. stability conditions of both arrangements is given.

Derivations of the 'h' matrix parameters for the common-emitter and common-base connections of the 'Darlington' pair have been given by Ghandi[2] in terms of the common collector 'h' parameters of the individual transistors.

4.3 *A Phase-splitter Driver for Push-pull Output Stages*

Where the highest performance is required a Class A push-pull output stage is almost essential. A phase-splitting transistor, with a load in the collector and emitter circuits, forms a simple and convenient means of driving such an output stage when it is preceded by single-ended gain stages. However, the impedances appearing between each of the output bases and the common line should be equal, since these impedances form part of the feedback circuits of the two halves of the output stage; but the effective impedance of the emitter circuit of the phase splitter is low compared with that of the collector circuit. Such considerations are of no account for the equivalent valve circuit where the total feedback circuit impedances are almost infinitely high, but become of importance with transistors where these impedances are comparatively low and are composed largely of the source impedances of the phase-splitter. If, then, these source impedances are not equal, the effective feedback applied to the two halves of the output stage will differ, when equal feedback voltages are used, and the stage will be unbalanced. The phase-splitter loads must therefore be designed to deliver equal currents to the two halves of the output stage from equal source impedances.

A representation of the arrangement used in several of the amplifiers to be described is given in Fig. 2. Here, TR1 is the phase splitter and TR2, TR3 the push-pull output pair. The resistance R4 is the parallel combination of the d.c. biasing resistors, Ro is the output resistance of the preceding stages, and E represents the feedback voltages, which give an input impedance Z_{in} to TR2 and TR3. E is generally large enough to make Z_{in} exceed 200 k.

It is shown in Appendix 1 that the desired conditions of equal currents into TR2 and TR3 from equal source impedances are met if the following relationships hold:

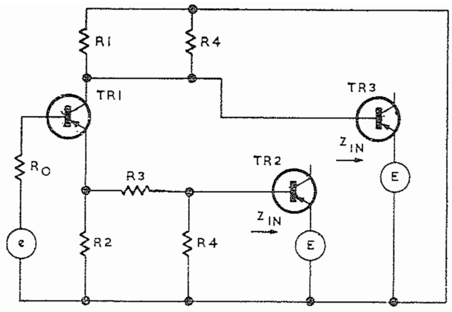

Fig. 2 — Transistor phase-splitter—Simplified Circuit

$$R_2 = R_1\left(1+\frac{R_3}{R_4}+\frac{R_3}{Z_{in}}\right)$$

$$R_3 \approx R_1 - \frac{R_0}{\beta}$$

where β is the current gain of the phase-splitter stage.

4.4 *The Microphone Amplifier: Noise Considerations*

The high-quality ribbon microphone used extensively in the BBC sound service is pressure-gradient operated and has an output of about -75 dB with respect to 1 volt/dyne/cm^2. It has an output impedance of 300 ohms and is normally followed by an amplifier with an input impedance of 600 ohms. The programme volume, as defined in Section 5.2.1 below, given by such a microphone with an acoustic input of average loudness is about -75 dB. The thermal noise which accompanies the signal has an e.m.f. of -131 dB for a 10 kc/s bandwidth at normal temperature, and therefore an input voltage across 600 ohms of $-134\cdot5$ dB (reference $0\cdot775$V. R.M.S.). On this basis, the signal/noise ratio is $59\cdot5$ dB and even if amplifier noise reduces this figure by several dB the resulting noise is inaudible during programme breaks.

However, programme material and acoustic conditions vary widely, from very quiet speakers and distant placing of microphones in musical performances, which may necessitate an increase of circuit gain by 10–15 dB, to the more exuberant dance-band performances which may result in a microphone output of about -20 dB. The signal/noise ratio may therefore fall to the region of 45 dB during some types of programme with respect to thermal noise alone. Hence this noise can become audible in critical listening conditions, and it is highly desirable that the inevitable extra noise contributed by the amplifier should be as low as possible.

The excessive output produced on other occasions is dealt with either by the insertion of a resistance pad between the microphone and amplifier, when a general very high volume can be foreseen, or by means of the calibrated interstage gain control of the microphone amplifier which enables all channels to be adjusted to the same sensitivity without risking overload of the amplifier output stage. The input stages of the amplifier must still handle a comparatively high input volume.

The requirement is, therefore, for an amplifier of low noise factor but with the ability to accept a high input volume without distortion.

An essential requirement for low noise is a low-noise type and specimen of transistor. The noise produced by such a specimen is a smooth hiss, very similar to white noise in quality. It is due mainly to shot noise in the p–n junction modified by a certain amount of 'excess' or 1/f noise due to small, and at present unavoidable, defects such as surface and other leakage currents. An inferior transistor has a larger amount of this 'excess' noise, often accompanied by crackles, which make the noise sound not only greater but rougher and more irregular.

The published data of the transistor used for the input stage of the microphone amplifier shows a narrow-band noise factor of 2 dB at 2 kc/s and above, increasing at frequencies below 2 kc/s to become $5\cdot5$ dB at 500 c/s and 12 dB at 100 c/s. At a distinctly audible level the 1/f noise causes a somewhat lower-pitched sound compared with white noise, but at the very low levels at which amplifier hiss should appear, the aural sensitivity curve falls away at low frequencies at a much greater rate than the noise rises and the more audible band in the 4 kc/s region predominates to an extent which causes both types of noise to sound very similar.

The volume of total noise is reduced as the collector current and the collector/emitter voltage fall. The reduction below about $I_c=0\cdot3$ mA and $V_{c-e}=2$V is very small or absent with low-noise types, and values much below these may lead to difficulties due to the increase of cut-off current at the higher ambient temperatures. Even with the values used in the microphone amplifier described below, where $I_c=0\cdot28$ mA and $V_{c-e}=1\cdot5$V, the high degree of d.c. stabilization given by the d.c. coupled pair is necessary to avoid such an effect. Moreover, with regard to shot and partition noise alone, there is an optimum value of collector current,[3] which depends upon other characteristics of the transistor. This optimum is of the order of $0\cdot3$ mA for the type of transistor under consideration; a reduction much below this value may cause a reduction of 'excess' noise but will cause either a reduction or an increase of the noise factor over an extended audio-frequency band depending upon the quality of the particular transistor, i.e. whether 'excess' noise or the fundamental shot and partition noise predominates.

An additional matter affecting the noise factor of a transistor is the source impedance, for which there is an optimum value, generally between 500 and 1,000 ohms.

The optimum theoretical operating conditions and the minimum noise factor to be expected have been calculated for a particular type of low-noise transistor, the GET 106, following an analysis given by Stewart.[3] This analysis is based on the theory of shot noise in p–n junctions; the thermal noise generated in the extrinsic base resistance, r_{bb}', is included but 'excess' noise is neglected. The noise factor is defined as the ratio of the total input noise power to the input noise power thermally generated in the input-

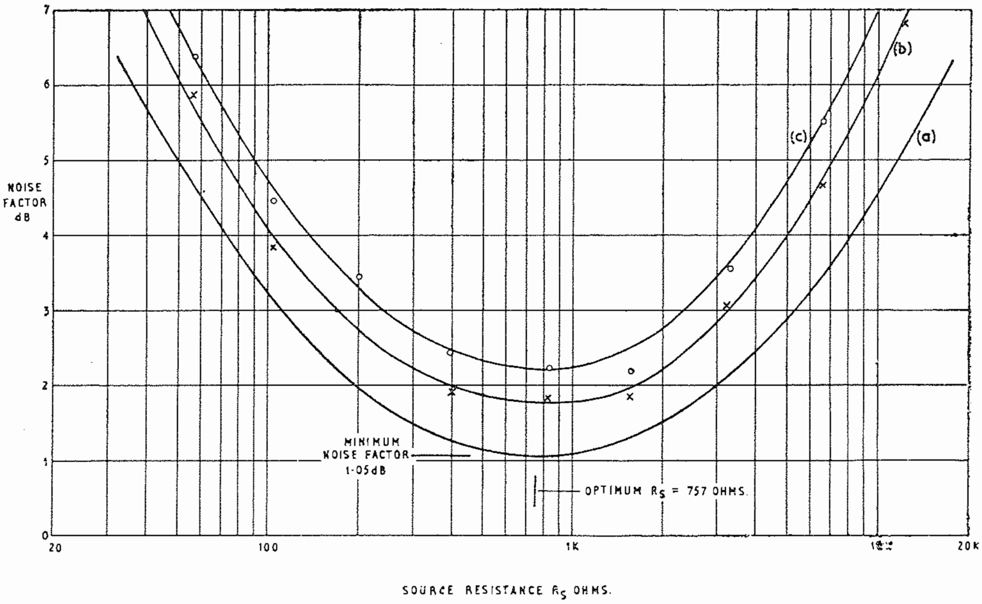

Fig. 3(a) — *Calculated noise factor plotted against source resistance for an idealized GET* 106 *transistor having the characteristics* $r_{bb'}=55$ *ohms*, $i_{co}=5\mu A$, $h_{fe}=60$, *for the calculated optimum emitter current of* 0·266 *mA*

(b) *and* (c) — *Measured noise factors of two GET* 106 *transistors in the bandwidth* 400 *c/s* — 10 *kc/s*, *at* $i_e=0\cdot3$ *mA*, $V_{c-e}=1\cdot43$ *volts.*

Transistor (a) $i_{co}=9\cdot5\mu A$, $h_{fe}=103$

Transistor (b) $i_{co}=18\mu A$, $h_{fe}=88$

terminating admittance, and in the present discussion is limited to that which exists at frequencies below about one-tenth of the cut-off frequency, f_α, a region which should well cover the audio-frequency band. It is shown[3] that for low noise factor a transistor having a high current gain and a low r_{bb}, and i_{co} is desirable, and that the three basic transistor configurations have essentially equivalent noise factors and the same optimum values of the operating conditions.

The calculated noise factors for various values of input circuit impedance are plotted in Fig. 3(a) for an ideal GET 106 type of transistor operating with the calculated optimum value of emitter current. The calculated optimum input termination is 757 ohms and for that value the noise factor is 1·05 dB. This, therefore, is the lowest noise factor which can, theoretically, be obtained from such a transistor.

For comparison with calculated results the noise factors given by two quiet specimens of GET 106 transistors in the bandwidth 400 c/s–10 kc/s are plotted in Figs. 3(b) and (c). Each was used as the first stage of a pair similar to the microphone amplifier input stages with no input transformer. The method of measurement was as follows. With the required input termination resistance the total output noise was measured with an R.M.S. meter restricted to the given bandwidth by a filter of accurately known characteristic; white noise was then added, as an e.m.f. in series with the input termination, of the amount required to cause an increase of the indicated output noise by 3 dB. The added white-noise R.M.S. voltage was then measured with the same meter-filter combination and compared with the calculated thermal noise of the input termination for a bandwidth assessed from the known filter characteristics. Since, for an increase of 3 dB, the added noise input power is equal to the original noise equivalent input power, the ratio of added noise to thermal noise is the required noise factor. The results are shown in Figs. 3(b) and (c). For the same two transistors, calculation gives:

Transistor (a): Optimum i_e $=0\cdot48$ mA
Optimum $R_s=570$ ohms
Minimum noise factor$=1\cdot15$ dB

Transistor (b): Optimum i_e $=0\cdot6$ mA
Optimum $R_s=795$ ohms
Minimum noise factor$=1\cdot4$ dB.

It will be seen from the foregoing that for an ideal transistor of this general type the best noise factor that can theoretically be obtained under perfect conditions is about 1 dB and that a quiet transistor in a practicable amplifier produces a noise factor of the order of 2 dB. The optimum source impedance lies between 600 and 1,000 ohms.

When an input transformer is needed, as is most often the case, an additional loss of noise factor is brought about by the resistance of the windings. This, and also the effect of other resistances, such as those associated with feedback voltages, must also be considered.

It is shown in Appendix 2 that the effective thermal-noise power brought about by a combination of additional shunt and series resistances such as those existing at the input circuit of the microphone amplifier, relative to the thermal-noise power produced by the source resistance, R_s, acting alone is:

$$\left\{\left(1+\frac{R_s+R_t}{R_f}\right)^2 R_e+\frac{(R_s+R_t)^2}{R_f}+R_s+R_t\right\}\Big/ R_s=N$$

and, if the noise factor of the amplifier with R_s alone, Fig. 13(b) (p. 19) is 10 logN_0, then the noise factor of the amplifier with the more elaborate input circuit, Fig. 13(a), is:

$$10 \log[N+N_0-1] \text{ dB}.$$

The symbols refer to the equivalent diagram of the input circuit of Fig. 5 (p. 12) shown in Fig. 13(a) in which R_s and R_t are the source resistance and transformer winding resistance, both referred to the transformer secondary side, and R_e and R_f are the total resistances of the series and shunt feedback circuits respectively. The result of a numerical evaluation of the formula with respect to Fig. 5 is given later when the microphone amplifier is described.

5. Apparatus Types—Description and Performance Details

The design principles considered in the preceding sections have been embodied in the amplifiers and other apparatus now to be described. They are divided into three groups: audio-frequency amplifiers, peak programme meters, and power supply.

5.1 *Audio-frequency Amplifiers*

5.1.1 *Line-sending Amplifier*

This amplifier, together with the line-receiving amplifier described later, is used for the distribution of programme over line chains, and since many are therefore likely to be used effectively in tandem, a particularly high performance is required.

A single push-pull stage of two 'Darlington' pairs is used with an input transformer and a balanced choke output circuit with isolating capacitors. The input impedance is 35,000 ohms and the output impedance is 100 ohms, which is built out to 600 ohms with resistors. The choke output arrangement is adopted in order to avoid the effects of transformer leakage inductance on the output impedance, which must be 600 ohms of small angle for the accurate termination of the line with which it is used. The 'Darlington' connection of transistors provides sufficient gain from what is effectively a single stage to enable 24 dB of feedback to be applied from secondary windings on the output choke to the emitter circuits of the transistors.

The voltage gain is 10 dB, the frequency response is flat to within $\pm 0\cdot 25$ dB between 25 c/s and 20 kc/s, and the total harmonic distortion is not greater than $0\cdot 2$ per cent between 60 c/s and 5 kc/s at the normal peak working output of $+18$ dBm. The power consumption is 60 mA at 24 volts.

The prototype of this amplifier has been more fully described and a circuit diagram is given in Section 4.4 of Monograph No. 26.[1] The present apparatus resembles the prototype closely except for the form of mechanical construction.

5.1.2 *Line-receiving Amplifier*

The prototype from which this final model was derived has also been described in the reference cited above and only a brief description, mainly concerning matters where further development has produced differences, will be attempted here. The circuit diagram is given in Fig. 4.

The amplifier is intended to follow the line equalizer at BBC centres and therefore an accurate 600 ohms input impedance and a fairly high adjustable gain are needed.

The circuit is made up of a d.c./a.c. feedback pair, followed by a phase-splitting transistor which drives a push-pull stage of 'Darlington' pairs.

The input transformer has condensers connected across primary and secondary windings which form in association with the leakage inductance an approximation to a low-pass filter section having a cut-off frequency well above the audio-frequency range. By this means the effects of the leakage inductance on the input impedance is greatly reduced and an input impedance of 600 ohms, with small angle, is obtained by a resistive load across the secondary winding. This resistance is arranged to be a potential divider giving three 10-dB steps of gain control. Both d.c. and a.c. feedback are derived from a resistance network in the emitter circuit of the second transistor. The a.c. series-connected feedback can be varied from a magnitude of 22 dB to 32 dB in half-dB steps by the adjustment of a stepped gain control which forms part of the feedback network. The final model differs from the prototype by the removal of this gain control from the d.c. path of the second transistor in order to avoid the slight clicks which the operation of the gain control originally caused. Although alteration of gain is not normally required during operation, the ability to do so without causing noise may be of importance in some applications.

The phase-splitter and output stages are the same as those of the prototype with the exception that the output transformer has been replaced by one which gives an output impedance of 55 ohms, for the reasons discussed in Section 3.2 above.

The gain of the amplifier is variable from 20 dB to 60 dB in half-dB steps, the frequency response is within $\pm 0\cdot 3$ dB from 40 c/s to 15 kc/s and the total harmonic distortion does not exceed $0\cdot 3$ per cent at the normal peak output of $+8$ dB volts (reference $0\cdot 775$V. R.M.S.). The power consumption is 45 mA at 24 volts.

5.1.3 *Microphone Amplifier*

Although the circuit details of this amplifier resemble closely those of the prototype,[1] the circuit diagram is reproduced here (Fig. 5); it illustrates clearly the matters regarding the d.c./a.c. feedback pair and noise factor treated above at some length. The prototype was developed originally in a form suitable for inclusion in a small channel assembly together with an interstage channel fader and various switches; the construction considered and illus-

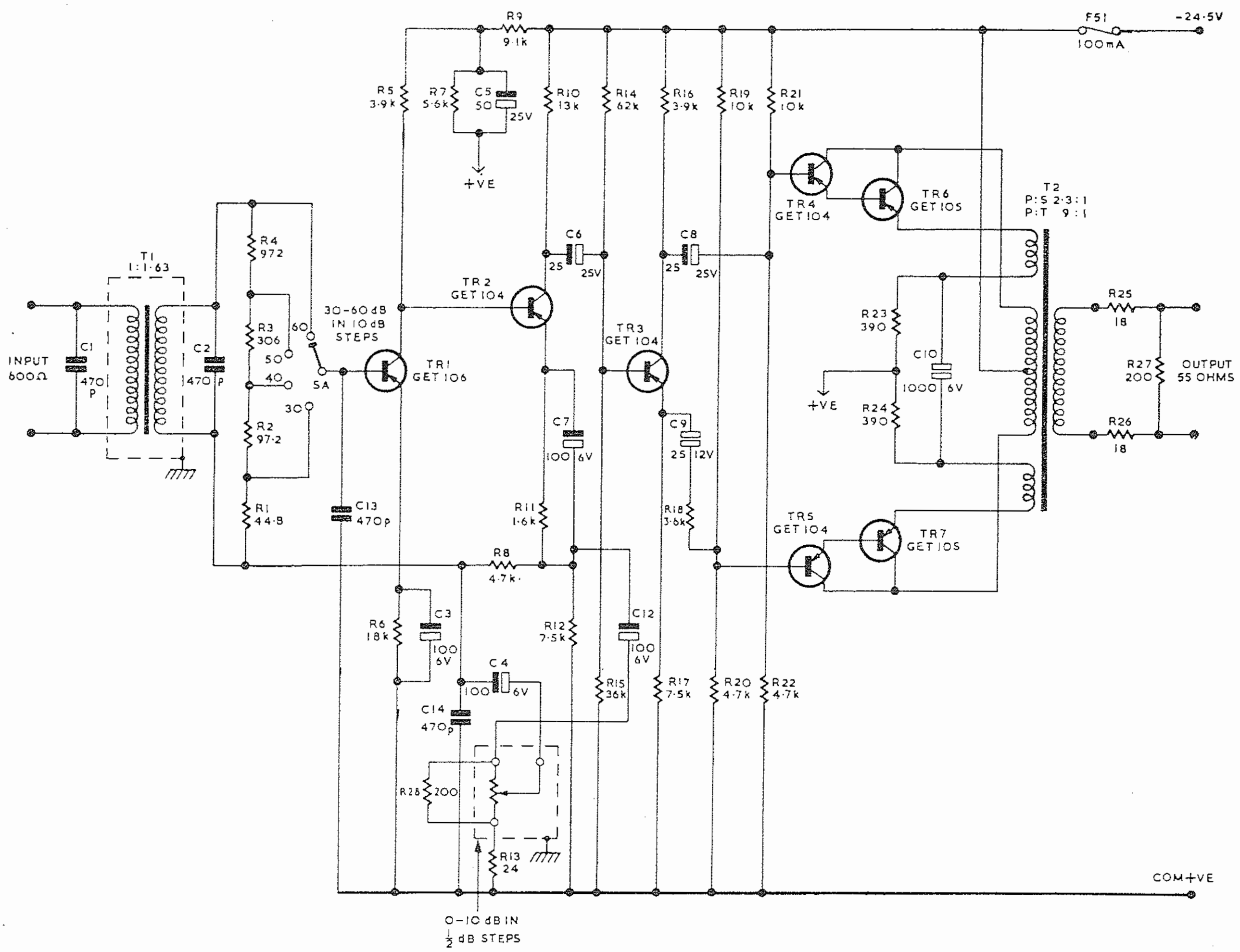

Fig. 4 — Line-receiving Amplifier—Circuit

trated here, (Fig. 11, p. 17), is intended for use in studio and other control desks as a microphone amplifier and also as group, intermediate, and echo amplifiers.

The circuit arrangement will be seen to consist of a feedback pair followed by an interstage gain control and a single-ended, unbalanced output stage. The control-desk wiring and attenuators are normally unbalanced, but in some circumstances it has been found necessary to avoid the establishment of troublesome earth-loops by the insertion of a repeating coil at some points. The output impedance is 600 ohms and the input impedance may be either 600 ohms, for use with 300-ohm microphones and for other purposes, or 60 ohms for use with 30-ohm microphones. The connections for the interstage gain control are brought out at the terminating plug from which they may be taken away to a channel gain-setting control of ten 3-dB steps mounted close to the corresponding channel fader. Alternatively, the gain may be preset to a required amount, within the maximum of 46 dB, by means of resistors at the terminating socket, so that amplifiers may be exchanged without upsetting gain conditions.

The transistors TR1 and TR2 have a considerable degree of d.c. feedback taken from the junction of R7/R8 to the base of TR1. The d.c. stability resulting from this is very great; the conditions are closely similar to those of the prototype which have been fully described in the reference quoted above. Both series- and shunt-connected a.c. feedback are also taken from the emitter circuit resistance of TR2. Together they have a magnitude of 26 dB and the ratio is adjusted to give the required input impedance.

The input circuit is discussed with regard to noise factor in Section 4.4 above and an analysis of the conditions is made in Appendix 2. When the values relating to the microphone amplifier input circuit are put into the equation it is shown that if the noise factor of the amplifier with R_s alone (Fig. 13(b)) is 2 dB, then it will be increased to 2·75 dB by the addition of the other resistances. The transformer winding resistance contributes about 0·4 dB to this increase of 0·75 dB and the feedback resistors the remainder.

It will be seen from the above discussion in conjunction with the measured results shown in Fig. 3 that the lowest noise factor to be expected from the microphone amplifier

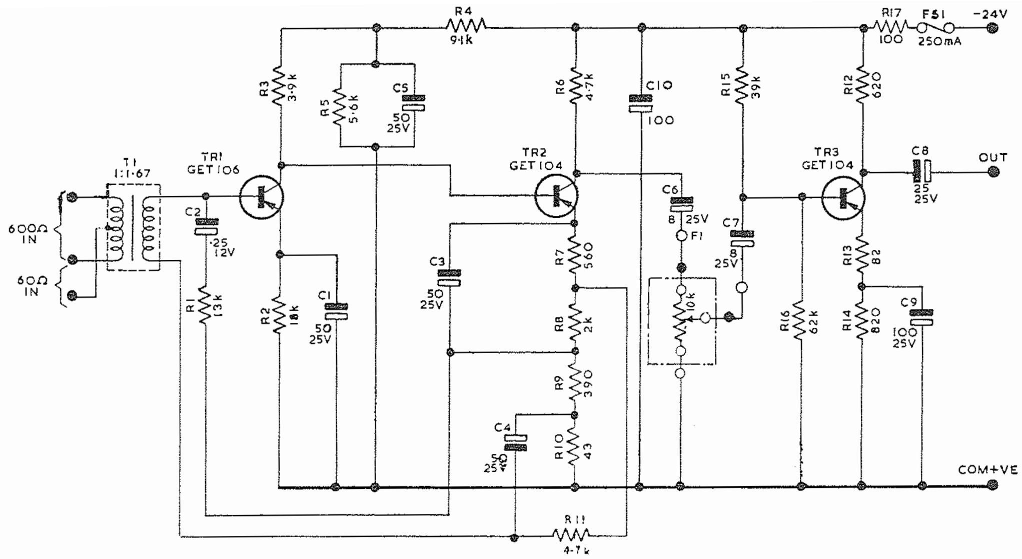

Fig. 5 — Microphone Amplifier—Circuit

will be about 3 dB and this proves to be the case in practice. This figure can be exceeded by 1 dB or more if less quiet transistors are used. The second and third transistors normally contribute an insignificant amount to the total noise but their operating conditions are necessarily not the optimum with regard to noise and the margin is not great enough to permit a noisy specimen to be used without effect on the overall noise factor.

The normal working peak input voltage of the amplifier is of the order of —70 dB (reference 0·775V. R.M.S.) and at these levels the distortion is about 0·1 per cent. The amplifier will accept an input voltage of —38 dB at maximum gain, and of —25 dB at lower gain settings, before the onset of serious distortion.

The frequency response is within +0·2 dB and —0·5 dB relative to mid-band, over the range 40 c/s to 10 kc/s. The power consumption is 15 mA at 24 volts.

5.1.4 *O.B. Line-sending Amplifier*

A line-sending amplifier for outside broadcasting requires features different from and additional to those suitable for the equivalent amplifier at studio centres. These are, mainly, an output impedance of 75 ohms, a means of varying the output volume in order to cope with abnormal line conditions, and a monitoring point which is independent of variations of line impedance.

The circuit of such an amplifier is shown in Fig. 6, from which it will be seen that an input transistor with series feedback is followed by a phase-splitter and a 'Darlington' pair push-pull output stage. The input impedance is 12,000 ohms and the output impedance 75 ohms. The normal voltage gain of 20 dB can be varied by ±4 dB by the alteration of the first stage feedback (switch SA1). This facility is used to change the normal output volume of zero dB volts by the appropriate amount. A monitoring output is taken from the emitter circuit of one side of the push-pull stage by way of transformer T2 and switch SA2 to deliver a final volume of —20 dB volts to the peak programme meter ME 12/5 with which it is used. The switches SA1 and SA2 are ganged and the potential divider R21–R24 of the monitor circuit is so arranged as to make the monitor output constant at all switch positions. The programme meter can in this way be read in the normal manner at any of the three possible output volumes.

The monitor output is derived from the voltage existing across one half of the feedback winding of the output transformer plus the voltage developed across R20 which is in series with the emitter circuit of one of the output transistors. The magnitude of these two components of the monitor voltage are therefore related to the output voltage and to the output current respectively, and their relative magnitudes are adjusted to a ratio which makes the monitor voltage very nearly independent of the main load impedance; the monitor level changes by only about 0·2 dB when the load impedance is varied between 50 ohms and open circuit. The programme meter therefore measures in effect the e.m.f. of the sending circuit.

A modified form of the amplifier is used as an output amplifier in studio and similar control desks at BBC centres. For such purposes an adjustable output volume and the monitoring output are not required, and the main output circuit is the arrangement previously described of 55-ohm impedance suitable for feeding into the long cable and switching system of a large installation. Apart from these omissions and slight modifications the circuit is that of Fig. 6.

5.1.5 *Response-selection Amplifier*

This apparatus, which provides means of manipulating the frequency response of a microphone channel to obtain

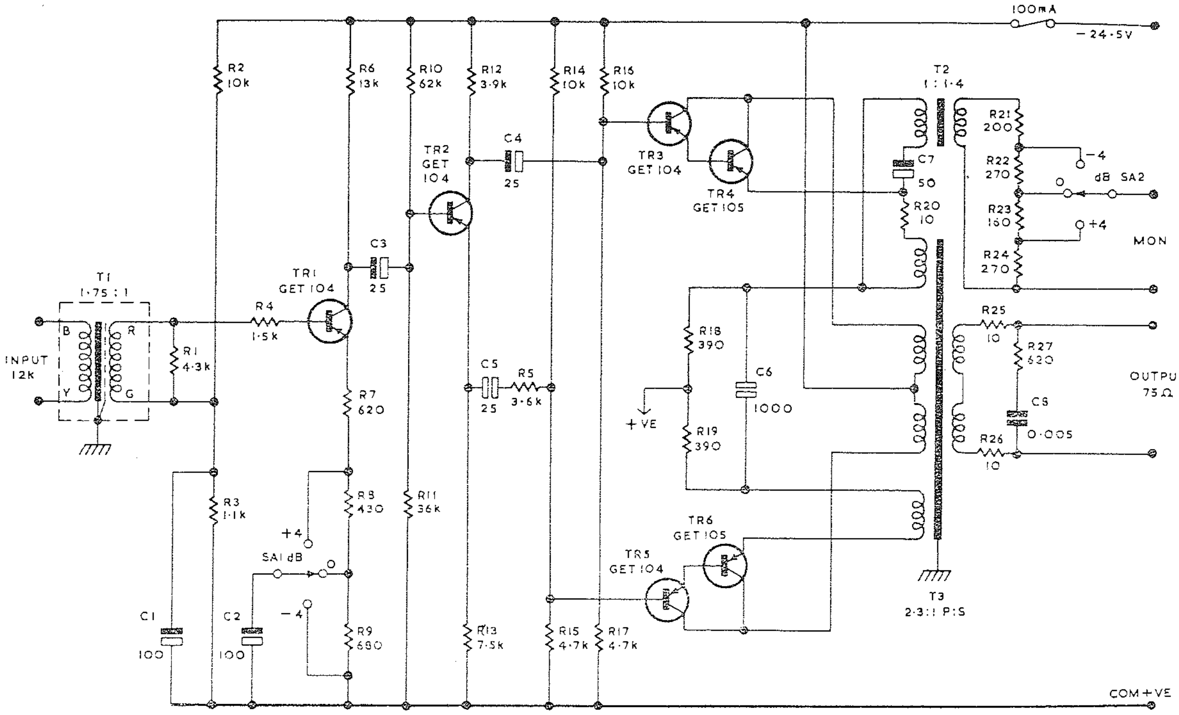

Fig. 6 — O.B. Line-sending Amplifier—Circuit

the effects required in the production of some kinds of light music, is based on a circuit arrangement due to Baxandall[4]; the circuit is given in Fig. 7.

Consider the single-stage 'Darlington' pair of transistors; the R/C network connected to the base of TR1 is fed, through C1, with the input signal and also, through C5, with a negative feedback voltage from the collector circuit. If the three capacitors C2, C3, and C4 are all open circuited and the slider of RV1 is in its mid-position, the arrangement becomes a 'collector follower' with a voltage gain of very nearly unity, and if the slider of RV1 is displaced the voltage gain becomes approximately the ratio (resistance between slider and C5)/(resistance between slider and C1). Conditions are similar with respect to RV2 if the three condensers are simultaneously short-circuited. These open-circuit and short-circuit conditions exist for the complete circuit at very low and very high frequencies respectively, and the stage gain at these extremes may therefore be varied on either side of unity by RV1 at low frequencies and by RV2 at high frequencies.

At a frequency of about 700 c/s the reactances of C2 and C3 are low and that of C4 is high, compared with the resistances with which they are associated, and each of the controls RV1 and RV2 has little effect on the stage gain. This frequency is therefore a cross-over frequency below which C2 and C3 become more and more effective as the frequency falls and above which C4 becomes increasingly significant as the frequency rises. Hence, either a progressive rise or fall in the frequency response may be obtained, by RV1 below about 700 c/s, and by RV2 above about 700 c/s.

The coupling circuit leading to the output transistor TR3 includes the tuned circuit L1, C8, and the switch SB with which it may be put into circuit to give two degrees of broad 'presence' peak in the frequency response.

The switch SA enables the whole of the response-determining elements to be put in or out of the chain at will.

When set to a flat-response condition the amplifier has an overall gain of about unity. A rising or falling frequency characteristic may be obtained at either end of the audio-frequency range with a maximum excursion of some 14 dB at 60 c/s and 10 kc/s. The gain at 700 c/s remains constant to within ± 1 dB. The 'presence' peak occurs at 2,800 c/s and may be given an amplitude of either 3 dB or 6 dB.

5.2 *Peak Programme Meters*

The early form of peak programme meter described in Monograph No. 26 was designed for a particular purpose and was not suitable for further development. It has therefore been superseded by a newly designed instrument, the Peak Programme Meter ME 12/4,* which will be the standard BBC meter for the measurement of programme volume. A need has also become apparent for an instrument which would measure programme volume in a similar manner but which would be smaller and more simple and need not have the closely controlled characteristics of the standard instrument. This simple meter is adequate for the

* Patent Application No. 1948/61.

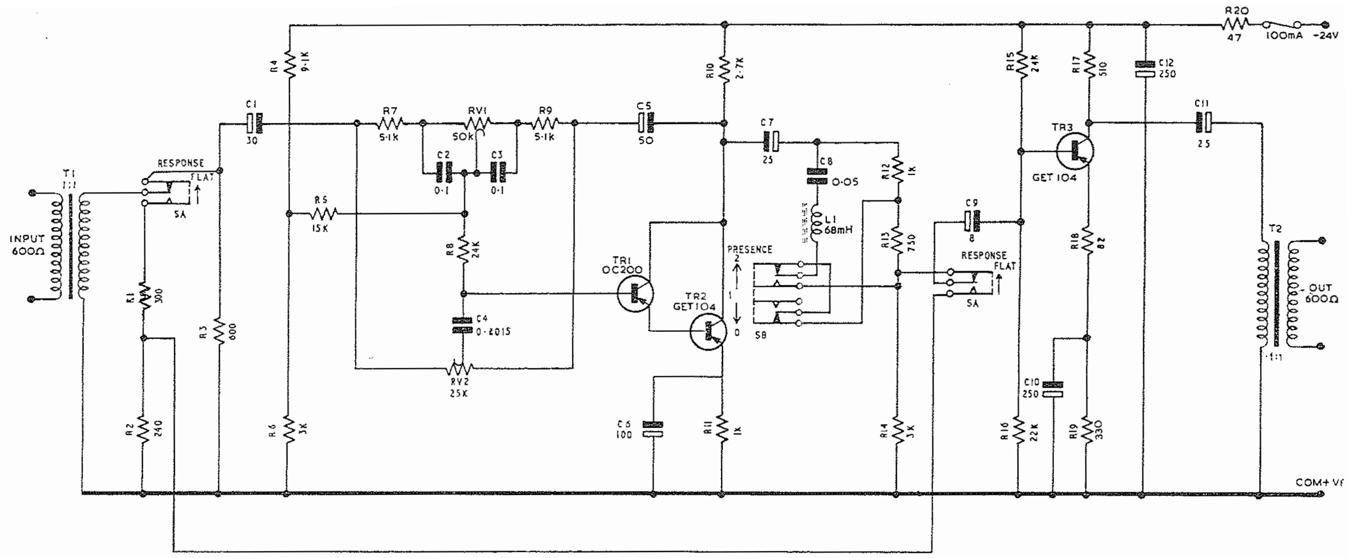

Fig. 7 — Response-selection Amplifier—Circuit

control of outside broadcast programmes, recordings fed into a studio desk, the output to echo channels, and for other situations where a standard meter is employed subsequently in the chain. Such a P.P.M. has been designed and is coded ME 12/5.

Both types of P.P.M. are used with indicating meters which are closely similar in appearance, to each other and to those used with the existing valve apparatus types, but they have a conventional left-hand zero in contrast to the right-hand no-current pointer indication used in the past.

5.2.1 *Peak Programme Meter ME 12/4*

The standard P.P.M. ME 12/4 measures programme in close conformity with long-established BBC practice, i.e. the signal is full-wave rectified and applied to a capacitive load which has a charge time-constant of 2·5 milliseconds and a discharge time-constant of 1 second. The resulting quickly rising and slowly falling pulses are amplified and displayed on a d.c. moving-coil meter which has a scale marked in approximately equal divisions from 0 to 7; each step from 1 to 7 indicates 4 dB. The instrument is normally adjusted to give a reading of '4' on the meter when zero dB volts of 1 kc/s tone (0·775 volts R.M.S.) are applied to the input. A programme volume of zero dB is then defined as that volume which causes a peak reading to the mark '6', i.e. 8 dB above the reference mark '4'.

A circuit diagram of the ME 12/4 is given in Fig. 8. The input amplifier consists of a single-stage push-pull arrangement with each half made up of two transistors connected as 'Darlington' pairs. The potential dividers which supply the bias currents are connected between the respective collectors and earth and therefore also provide balanced sources of a.c. voltage feedback connected to the transistor bases in series with the input transformer secondary windings. The magnitude of this feedback, about 16 dB, is determined by the resistance of R1+RV1; the latter forms a sensitivity control. The ganged switches SA1 and SA2 enable the gain to be varied by ±8 dB about the normal condition for convenience in checking the scale shape of the apparatus. The input impedance is 50,000 ohms.

The signal from the input amplifier is full-wave rectified by the bridge MR1–MR4 and the resulting unidirectional pulses are applied, from a source impedance of about 500 ohms, to the rectifier load network shown in the diagram, in the pass direction of the rectifiers. Consider a pulse arriving from the input amplifier. At low levels the forward resistance of all the diodes is high and the relationship between the input voltage and the voltage across the rectifier network is very nearly linear. This state continues to around the point '2' on the meter and the sensitivity control RV1 is preset to give this reading at the correct input level.

As the input increases further the current through MR5 and MR6 increases to a point where the resistance of the diodes becomes comparable with the circuit resistance, and over the next part of the range the voltage across the network rises approximately in proportion to the logarithm of the input. The first 'law' control RV2 is preset to give a meter reading of '4' at the corresponding input level. Below '4' the voltage across R17 and RV2 in series is insufficient to cause significant current to flow through MR7 and this arm of the network is therefore inoperative. As '4' is passed MR7 comes into operation and the logarithmic law is extended to around the meter point '6'. The second 'law' control RV3 is preset for accurate calibration at '6'. The third arm of the network includes MR8, a low-impedance germanium diode, which limits the reading at '7' and beyond.

The capacitor C3 is charged from an impedance consisting of the parallel combination of the amplifier output impedances and that of the logarithmic network. The charge time-constant therefore alters slightly during the charge process, but the variation is imperceptible in practice. It has an effective value of the required 2·5 milliseconds. The discharge time of C3 is controlled by the resistance of R20, R21, and the input resistance of the following d.c. stage to give the required 1-second time-constant.

The output d.c. amplifier is made up of two 'Darlington' pairs of silicon transistors in a 'long-tailed pair' bridge arrangement with the upper arm of one of the bias potential dividers made variable over a small range to form a 'zero' control (RV6). The two variable resistors RV4 and RV5 in the emitter circuit afford means of adjusting the gain and the balance of the d.c. amplifier. They, and also the two 'law' controls RV2 and RV3, are mounted internally and are used only on the initial setting-up of the apparatus.

The meter used with the apparatus has a full scale deflection of 1 mA and the controlled ballistic characteristics of the previous BBC programme meters. It is mounted in a convenient position apart from its amplifier in series with the 2,000-ohm resistor shown in the diagram. Additional meters, up to a total of four, may be used if the value of the series resistor is altered accordingly.

The instrument can be adjusted precisely at three well-spaced points on the meter calibration and little opportunity exists for error at the intermediate positions. The initial setting-up can therefore be done with accuracy and the degree of feedback employed, together with the balance adjustment of the d.c. amplifier stage, ensures that good accuracy is maintained in all likely conditions of supply voltage and temperature change. The zero drift is very small and adjustment of either the zero or the sensitivity controls should rarely be necessary.

The frequency response is flat within 0·25 dB over the range 40 c/s to 15 kc/s.

5.2.2 *Peak Programme Meter ME 12/5*

This is the simple and small instrument used in situations where the greater accuracy of the standard ME 12/4 is not necessary. Its characteristics are nominally the same as those of the standard and the circuit arrangement is similar in principle but has been considerably simplified by some sacrifice of accuracy in some programme conditions. The circuit diagram is given in Fig. 9.

The input amplifier is a d.c./a.c. feedback pair with the second transistor acting as an emitter follower in order to obtain a low output impedance. The sensitivity is adjusted by means of the feedback control RV1 and means of presetting the gain to correspond with zero, −10, and −20 dB programme volume at the input is provided by an internal link. The input impedance is 12,000 ohms. The output of the amplifier is rectified and the resultant pulses charge the capacitor C5 which discharges through the meter circuit. The two diodes MR6 and MR5 begin to shunt current away from the meter at about mid-scale and towards the top of the scale respectively. After the sensitivity control has been set to obtain a meter reading of '2' at the appro-

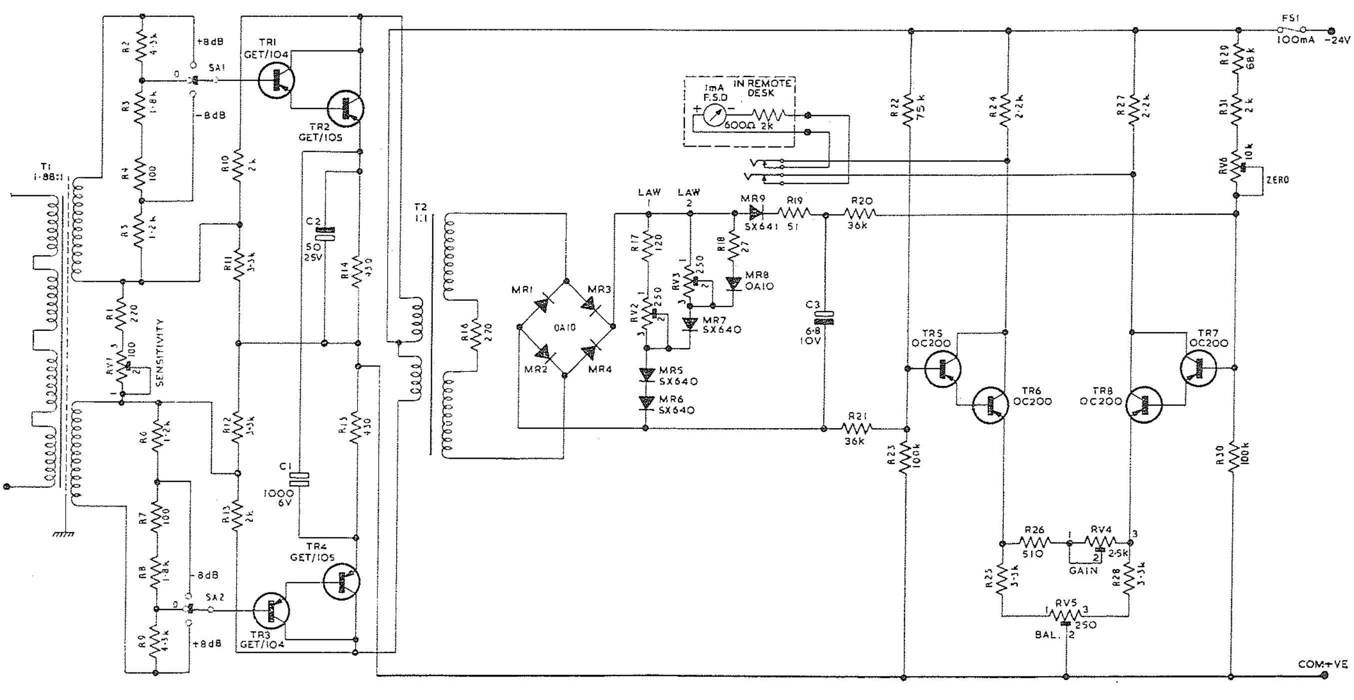

Fig. 8 — *Peak Programme Meter ME* 12/4—*Circuit*

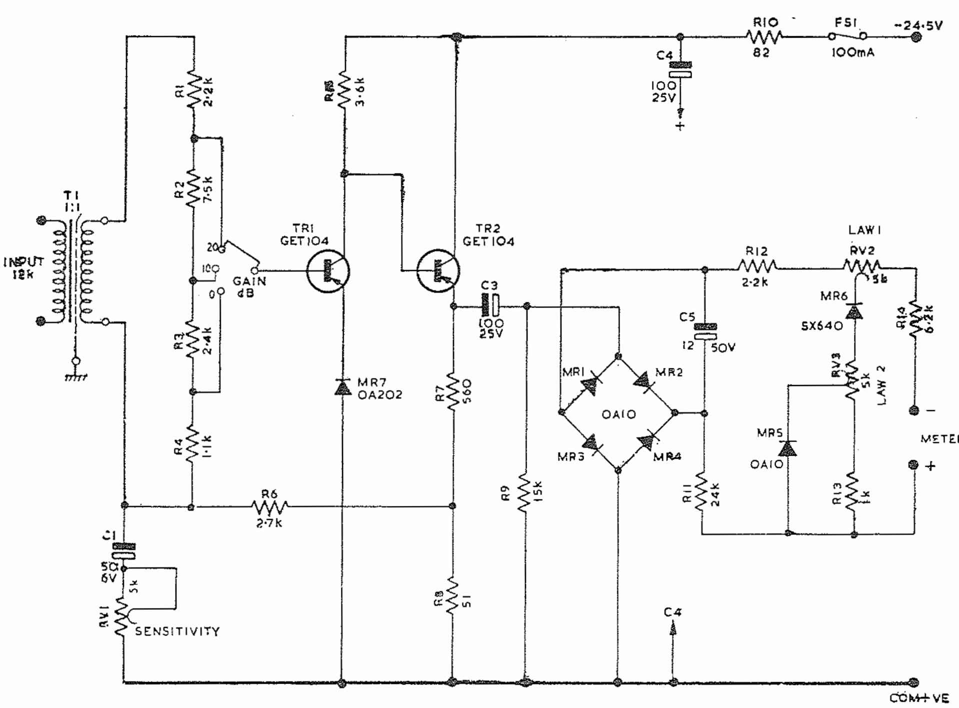

Fig. 9 — Peak Programme Meter ME 12/5—Circuit

priate input level the two 'law' controls are adjusted to give readings of '4' and '6' respectively when the input level is correspondingly altered.

The charge time-constant of the circuit is about 3 milliseconds. The discharge time-constant varies during the discharge period but the fall-back time is about the same as that of the standard instrument. The moving-coil meter used with the apparatus has a full scale deflection of 100 μA and ballistic characteristics which, with most types of programme, result in meter indications closely similar to those given by the standard instrument. With some programme material, however, peak readings can occasionally be in error, compared with the standard, by up to 2 dB in either direction. It is for this reason that the use of the ME 12/5

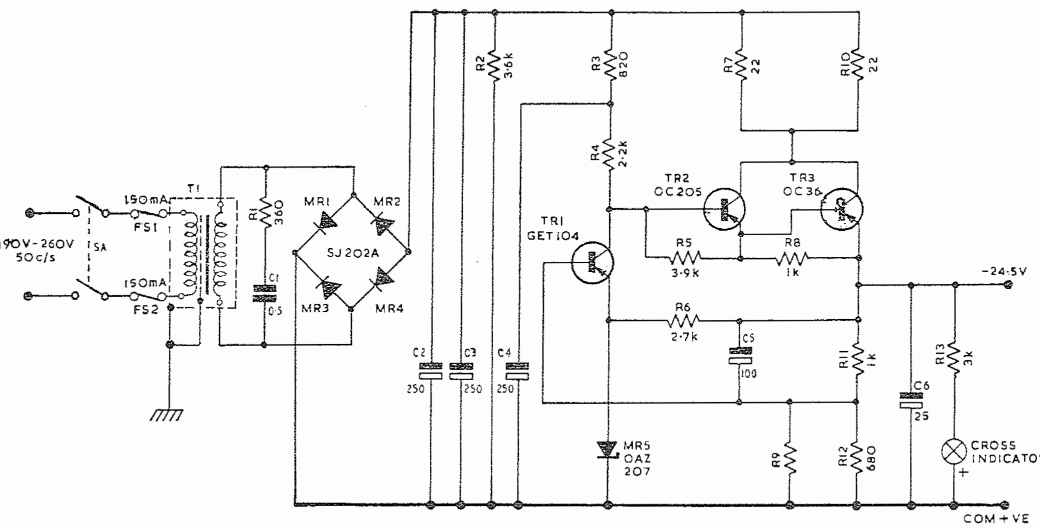

Fig. 10 — Power Supply—Circuit

is limited to situations where its small size is an advantage, and the possibility of some error in peak readings can be tolerated.

Since the method of initial adjustment of 'law' is the same as that of the standard instrument the initial accuracy obtainable is similar; stability of calibration is not inferior and there can be no zero drift if the isolating capacitor C3 has a negligible leakage current. It is a tantalum electrolytic in order to ensure this.

5.3 *Power Supply Unit*

A group of the amplifiers considered above is supplied with power by a stabilized power unit, the circuit of which is given in Fig. 10. The mains supply, stepped down and rectified, drives a series stabilizer, of which the series element is a 'Darlington' pair. The reference voltage is obtained from the zener diode MR5, which maintains constant the emitter voltage of the amplifying stage TR1, and the base of this stage is supplied from a potential divider connected across the output circuit. A variation of output voltage is therefore amplified by TR1 and transmitted to the series transistor TR3, in a sense tending to correct the variation. The resistance R9 is adjusted to give the required output voltage of 24·5 volts for mean conditions of input voltage and load current.

Protection against switching surges is given by R1, C1. The unit is designed to work from any value of mains input voltage at 50 c/s between 190 and 260 volts R.M.S. without adjustment and the circuit conditions, particularly the values of R7 and R10, are arranged to enable it to accept a short-circuit of the output without damage. The fuses FS1 and FS2 are delayed-action types and are for the purpose of protecting the transformer from overheating if a short-circuit condition persists. A low-consumption electromagnetic cross-indicator takes the place of the conventional pilot lamp.

The maximum output of the unit is 500 mA at 24·5 volts. The extra half-volt is provided to allow this amount to be dropped in wiring and light-current fuses while retaining 24 volts at the apparatus terminals.

The output voltage does not depart by more than 0·4 volts from the nominal value in any conditions of load or input voltage within the stated ranges. The output impedance is less than 0·75 ohms and the maximum hum voltage is less than 1 millivolt peak.

6. Mechanical Construction

Printed-wiring cards are used for all the units except the power supply, and most of the components are mounted

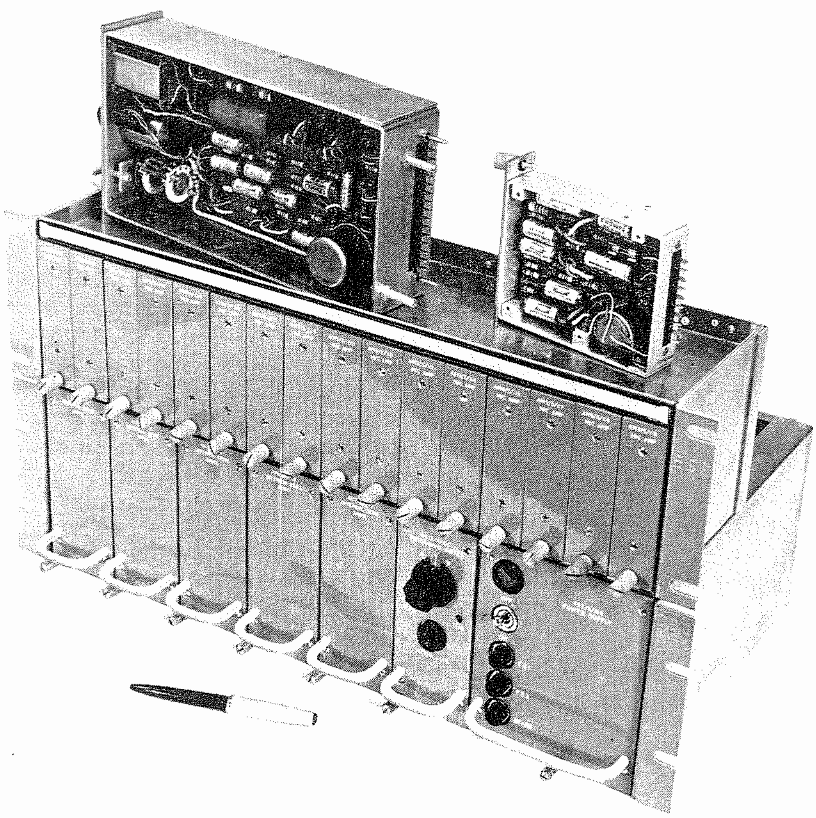

Fig. 11 — *Groups of amplifiers in their mounting panels together with a Line-receiving Amplifier and a Microphone Amplifier, showing internal construction and plug arrangements*

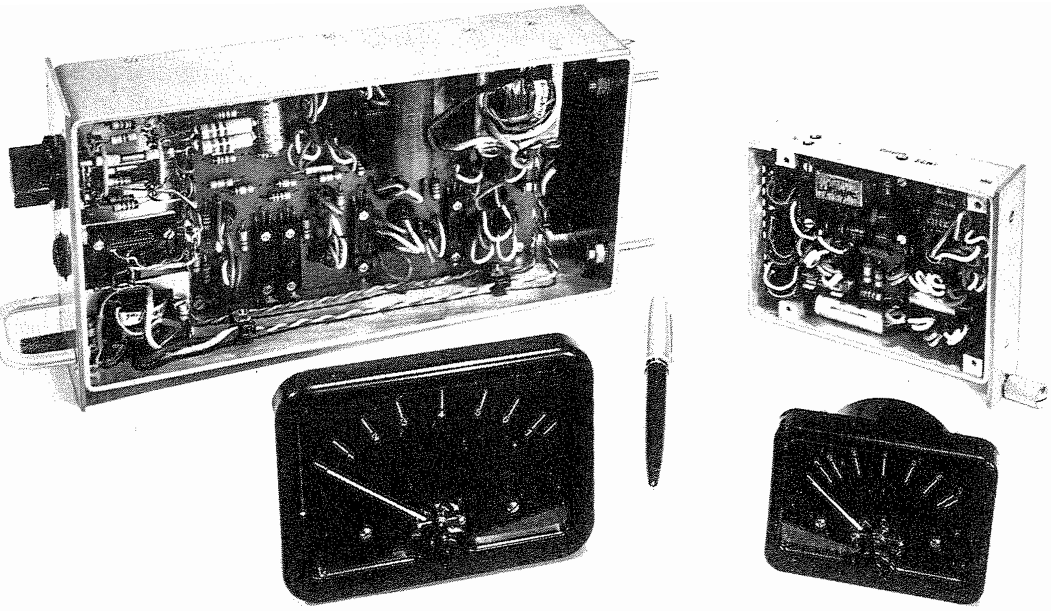

Fig. 12 — The Peak Programme Meters ME 12/4 and ME 12/5 with their associated meters

on them. The card is mounted in a chassis of small width so that both sides are readily accessible when a cover plate is removed. The construction of the individual units and the method of mounting them in groups are illustrated in Figs. 11 and 12.

There are two basic sizes of amplifier chassis; the larger has a front panel $2\frac{1}{16}$ in. by 5 in. and is $8\frac{5}{8}$ in. deep; the smaller has a front panel $1\frac{1}{16}$ in. by $3\frac{3}{4}$ in. and is $4\frac{1}{8}$ in. deep. The larger basic size is used for most of the units described and a double-width variety of it for the power supply. The use of the smaller size is confined to the microphone amplifier and the smaller of the two peak programme meters, the ME 12/5.

Each chassis has a multi-pin plug which engages with a corresponding socket at the back of the mounting panel into which the amplifiers are assembled in groups, each unit held in place by a screw fixing. The screw has a knurled knob for finger operation and is held in place so that when unscrewed it assists in withdrawing the plug from its socket.

There are also at the rear of the chassis two projecting spigots arranged in one of a number of coded positions; they engage with correspondingly located holes at the back of the mounting panel if the type of amplifier is correct for the particular position but prevent the insertion of a wrong type.

The larger mounting panel has the front dimensions of a standard $5\frac{1}{4}$-in. panel for 19-in. bay mounting and will contain eight amplifiers of the larger dimensions or six amplifiers and one mains unit. The mounting panel for the smaller amplifiers has the same width but is $4\frac{3}{8}$ in. deep; it will contain up to sixteen amplifiers.

7. Conclusion

A description has been given of some newly developed transistor apparatus together with relevant details of the system with which they are to be used and an outline of some of the design principles on which they are based. The larger programme of work, of which this apparatus forms part, is directed towards the utilization of transistors and other semiconductors to their best advantage in the whole of the audio-frequency apparatus used by the BBC. The examples dealt with were chosen because they constitute the electronics of the main chain from microphone to transmitter, and are therefore most important and will be used in large numbers. There is much auxiliary apparatus to which transistors can usefully be applied, portable equipment particularly. Some of this has already been done and the work continues.

8. Acknowledgments

The author would like to thank his colleagues of the BBC Designs Department for their valuable co-operation in the work described in this monograph.

9. References

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APPENDIX 1

RESISTANCE RELATIONSHIPS FOR THE PHASE-SPLITTER OF FIG. 2

When the current gain, β, of TR1 is large the voltages developed across R1 and R2 are proportional to the total resistances in the collector and emitter circuits, i.e. to $\frac{R_1 R_p}{R_1+R_p}$ and $\frac{R_2(R_3+R_p)}{R_2+R_3+R_p}$ respectively, where R_p is the parallel combination of R_4 and Z_{in}.

But the voltage applied to TR2 is reduced by the factor $\frac{R_p}{R_3+R_p}$ and, therefore, for equal currents into TR2 and TR3, the relationship should be:

$$\frac{R_1 R_p}{R_1+R_p}=\frac{R_2(R_3+R_p)}{R_2+R_3+R_p}\times\frac{R_p}{R_3+R_p}$$

from which $$R_2=\frac{R_1 R_3}{R_p}+R_1$$

putting $\frac{R_4 Z_{in}}{R_4+Z_{in}}$ for R_p and rearranging

$$R_2=R_1\left(1+\frac{R_3}{R_4}+\frac{R_3}{Z_{in}}\right) \qquad (1)$$

Also, the source impedance at the emitter of TR1 is, very nearly, $\frac{R_0}{\beta}$. The impedances presented to the inputs of TR2 and TR3 will then be equal if

$$R_1=\left[\frac{R_0 R_2}{\beta}\Big/\left(\frac{R_0}{\beta}+R_2\right)\right]+R_3$$

and therefore $$R_3=R_1-\frac{R_0 R_2}{R_0+\beta R_2}$$

Where $\beta R_2 \gg R_0$ this closely approximates to

$$R_3=R_1-\frac{R_0}{\beta} \qquad (2)$$

Equations (1) and (2) must therefore be satisfied if equal currents are to be fed to TR1 and TR2 from equal source impedances.

APPENDIX 2

MICROPHONE AMPLIFIER: EFFECT OF INPUT CIRCUIT RESISTANCES ON NOISE FACTOR

The relevant parts of the microphone amplifier input circuit are shown in Fig. 13(a). The noise factor given by an amplifier with such an input circuit is to be compared to one with the simple input circuit of Fig. 13(b).

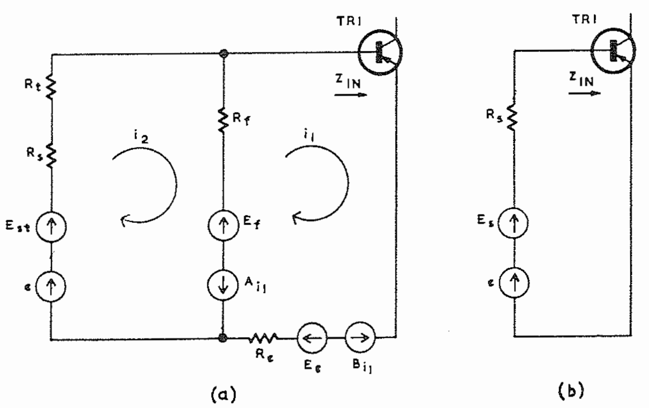

Fig. 13(*a*) *and* (*b*)

The resistances R_s and R_t are respectively the source resistance and the transformer winding resistance, both referred to the secondary of an otherwise ideal transformer; they have associated with them a signal e.m.f., e, and a total thermal noise e.m.f., E_{st}. The resistances R_e and R_f are those associated with the series- and shunt, connected feedback e.m.f.s, both of which are proportional to the input current i_1 and are shown as Bi_1 and Ai_1 respectively. R_e and R_f also have associated with them thermal-noise e.m.f.s E_e and E_f. Z_{in} is the input impedance of the transistor TR1. For the present the transistor is considered to be noise-free and the e.m.f.s are regarded as instantaneous voltages.

The following mesh equations may be set up:

$$e+E_{st}-E_f=-(A+R_f)i_1+(R_s+R_t+R_f)i_2 \qquad (1)$$

$$E_e+E_f=(A+B+R_e+R_f+Z_{in})i_1-R_f i_2 \qquad (2)$$

Eliminating i_2 from (1) and (2) results in:

$$E_e+E_f+\frac{R_f(e+E_{st}-E_f)}{R_s+R_t+R_f}$$

$$=\left\{(A+B+R_e+R_f+Z_{in})-\frac{(A+R_f)R_f}{R_s+R_t+R_f}\right\}i_1 \qquad (3)$$

Putting $\frac{R_f}{R_s+R_t+R_f}=P$ and $(A+B+R_e+R_f+Z_{in})=Q$

equation (3) may be written:

$$Pe+E_e+(1-P)E_f+PE_{st}=\{Q-P(A+R_f)\}i_1$$

Transferring attention now to R.M.S. quantities, the total transistor input power, W_1, can be considered to be made up of two parts, that due to the signal e.m.f.

$$Z_{in}\left[\frac{Pe}{Q-P(A+R_f)}\right]^2=W_{1a}$$

and the other due to the several thermal-noise e.m.f.s

$$Z_{in}\left[\frac{E_e^2+(1-P)^2E_f^2+P^2E_{st}^2}{\{Q-P(A+R_f)\}^2}\right]=W_{1b}$$

and the ratio of these is:

$$\frac{W_{1b}}{W_{1a}}=\frac{E_e^2+(1-P)^2E_f^2+P^2E_{st}^2}{P^2e^2}$$

The above condition is to be compared with the ideal case where e is acting in association only with the source resistance R_s and the corresponding thermal-noise generator E_s as shown in Fig. 13(b). It is assumed that other conditions, such as optimum source impedance, remain unaltered. In this event the ratio thermal-noise power to signal-noise power is E_s^2/e^2 and, therefore, for the same value of signal power the thermal-noise power is increased by the insertion of the other circuit components by the factor:

$$\frac{E_e^2+(1-P)^2E_f^2+P^2E_{st}^2}{P^2E_s^2}=\frac{W_{t2}}{W_{t1}} \tag{4}$$

where W_{t1} is the input thermal-noise power of the simple ideal circuit, Fig. 13(b), and W_{t2} is the total input thermal-noise power of Fig. 13(a), for the same value of input signal power in both cases.

But the noise e.m.f.s are proportional to the square root of the resistances by which they are generated, e.g. $E_e=K\sqrt{R_e}$ and, therefore, (4) becomes:

$$\frac{W_{t2}}{W_{t1}}=\frac{R_e+(1-P)^2R_f+P^2(R_s+R_t)}{P^2R_s}$$

$$=\left\{\left(1+\frac{R_s+R_t}{R_f}\right)^2R_e+\frac{(R_s+R_t)^2}{R_f}+R_s+R_t\right\}\Big/R_s=N \tag{5}$$

when P is replaced by its value $\dfrac{R_f}{R_s+R_t+R_f}$

Now the noise factor of the amplifier is, from the definition, given by the ratio (the sum of the total thermal-noise power due to the input circuit and the noise power generated by the remainder of the amplifier)/(the thermal-noise power due to the source impedance). The input thermal-noise power due to the source impedance R_s is W_{t1} and let the equivalent input-noise power generated by the remainder of the amplifier be denoted by W_{tr}. Then, if the noise factor for the simple circuit of Fig. 13(b) is 10 ($\log N_o$) dB,

$$N_o=\frac{W_{t1}+W_{tr}}{W_{t1}}=1+\frac{W_{tr}}{W_{t1}} \tag{6}$$

and for the circuit of Fig. 13(a) the noise factor is:

$$\frac{W_{t2}+W_{tr}}{W_{t1}}=\frac{W_{t2}}{W_{t1}}+\frac{W_{tr}}{W_{t1}}$$

but, from equation (5), $\dfrac{W_{t2}}{W_{t1}}=N$ and, from (6), $\dfrac{W_{tr}}{W_{t1}}=N_0-1$

The noise factor of Fig. 13(a) then becomes:

$$N_T=10\log[N+N_o-1]\text{ dB}$$

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