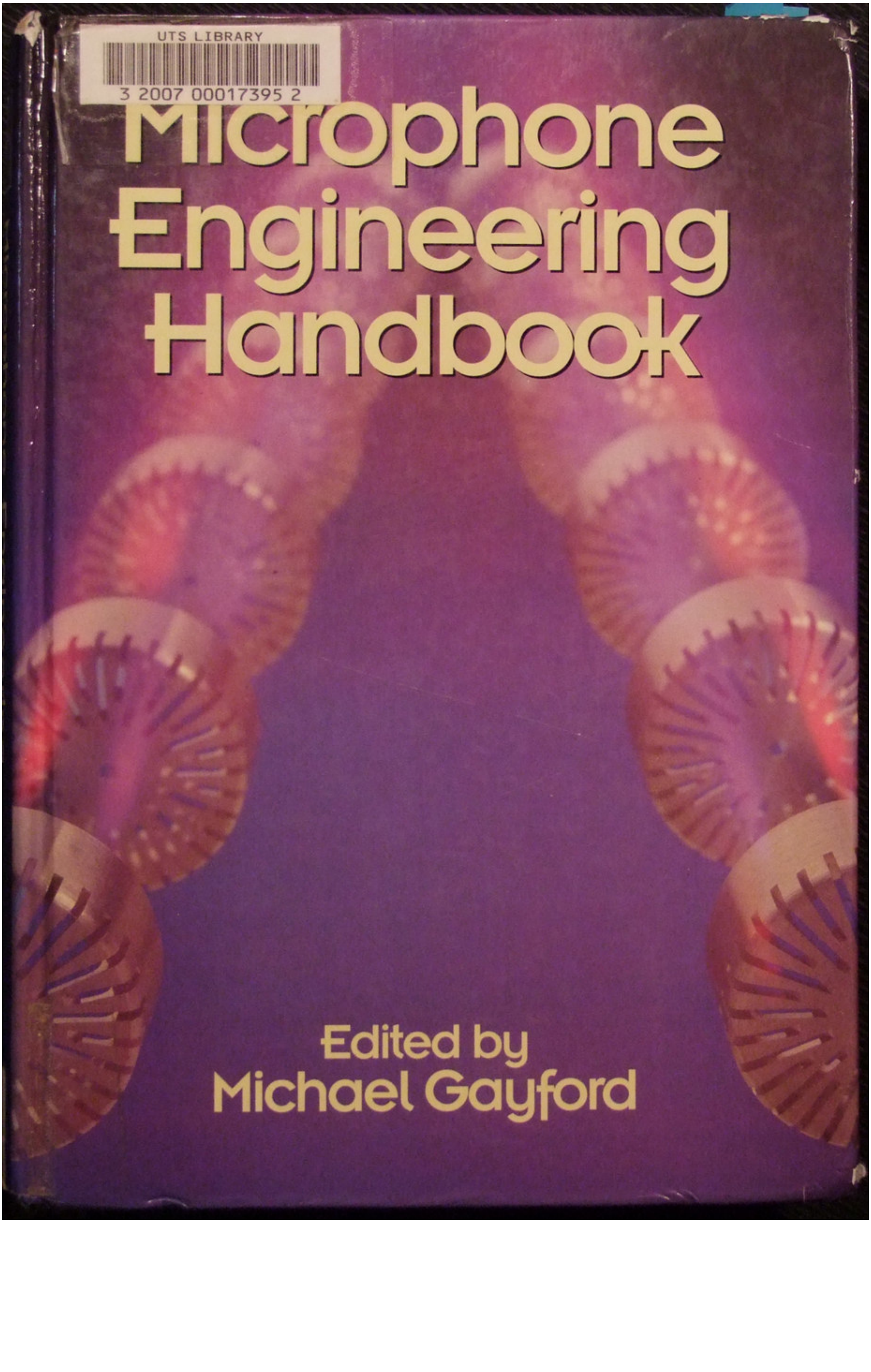

Microphone Engineering Handbook

Edited by
Michael Gayford

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Focal Press
An imprint of Butterworth-Heinemann Ltd
Linacre House, Jordan Hill, Oxford OX2 8DP

A member of the Reed Elsevier group

OXFORD LONDON BOSTON
MUNICH NEW DELHI SINGAPORE SYDNEY
TOKYO TORONTO WELLINGTON

First published 1994

British Library Cataloguing in Publication Data
Microphone Engineering Handbook
I. Gayford, Michael
621.38284

ISBN 0 7506 1199 5

Library of Congress Cataloguing in Publication Data
Microphone engineering handbook/edited by Michael Gayford.
p. cm.
Includes bibliographical references and index.
ISBN 0 7506 1199 5
1. Microphone – Handbooks, manuals, etc. I. Gayford, M. L.
TK6478.M53 93–33681
621.382'84–dc20 CIP

Cover photograph kindly supplied by Bruel & Kjaer

Composition by Genesis Typesetting, Laser Quay, Rochester, Kent
Printed in Great Britain by Redwood Books, Trowbridge, Wiltshire

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Preface

Today more than ever before the microphone represents a vital interface between the sound field and the multitude of electrical circuits now in use.

A number of excellent books have appeared which give a good grounding in the general theory and applications of microphones. However, it is apparent that there is an advantage in producing an up-to-date work in which the various specialized aspects of microphone design and use are covered in detail by acknowledged experts involved today. Special emphasis is placed on inclusion of comprehensive references to other work. The aim is to give clear and detailed treatment of microphone design and performance in fields such as communications, sound measurements, sound reinforcement and public address, broadcasting and sound recording etc., including the latest stereophonic techniques. It is hoped that all concerned with the engineering design and applications of microphones in the diverse fields of use today, and many students, will find interest and enlightenment from this book.

Michael Gayford
Harlow Essex

Acknowledgements

In compiling a technical handbook of this nature a considerable number of manufacturers are involved world-wide in providing technical information both directly and through reviews, catalogues and technical articles. The following are thanked in particular:

AKG Acoustics, AMS (Calrec), Bruel & Kjaer, beyerdynamic, Cannon, Coles Electroacoustics, ACO Pacific Inc, Crown International, Gotham Cables, Neumann, F.O Bauch Ltd, Magnetic Developments Ltd, Neutric, Jensen Transformers Inc, Sanken Microphones, Shure Inc, Sowter Audio Transformers, Stirling Sound Distributors, Switchcraft, Sennheiser, Electrovoice Inc.

Illustrations provided are acknowledged on the page. These are all gratefully acknowledged.

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8 Microphone amplifiers and transformers

Peter Baxandall

8.1 Introduction

In this chapter the primary aim is to present the *fundamental principles* that underlie the design of microphone amplifiers and transformers as lucidly as possible. However, attention will also be given to engineering aspects which practical experience has shown to be important.

Section 8.2 has been included largely because it is felt that some appreciation of how the present state of the art has come about helps to make work in this field more interesting and satisfying. The Section also provides a convenient way to introduce some topics that are dealt with in greater detail in later parts of the chapter.

8.2 History

A basic microphone amplifier is nowadays a diminutive device of great reliability, tiny power consumption and low cost – but how very different things once were!

The recording industry had no requirement at all for microphone amplifiers until about 1925 when electrical disc recording came in, and the significant history of such amplifiers really starts at about the same time as that of regular broadcasting, say 1920 or just after – although Wente's capacitor microphone and amplifier for sound-intensity measurement,[1] based on work at Bell Telephone Laboratories, appeared in the literature as early as 1917. This latter date is in fact only about five years after the vital notion first began to dawn that the triode valve, invented in 1907 by Lee de Forest, and regarded at first merely as an improved radio detector, had potent amplifying properties and could be cascaded to make high-gain amplifiers.

The BBC (British Broadcasting Company, as it then was) took over responsibility, in November 1922, for running three broadcasting stations that had previously been owned and operated for a short period by independent electrical firms.

One of these firms was Western Electric, whose 0.5 kW London station was soon moved (by steam lorry!) to Birmingham. It had equipment of American design, including a Western Electric Type 373 double-button stretched-diaphragm carbon microphone, which gave much better quality than ordinary telephone microphones, and a Western Electric Type 8A microphone amplifier. This amplifier, which was powered by storage batteries, is shown in Figure 8.1, and cost £160, equivalent to several thousand pounds today. It had three choke–capacitance coupled stages with sizable input and output transformers.

Surprising as it may now seem, a microphone amplifier as such was not used at all in other BBC installations in the very earliest days, the output of the studio

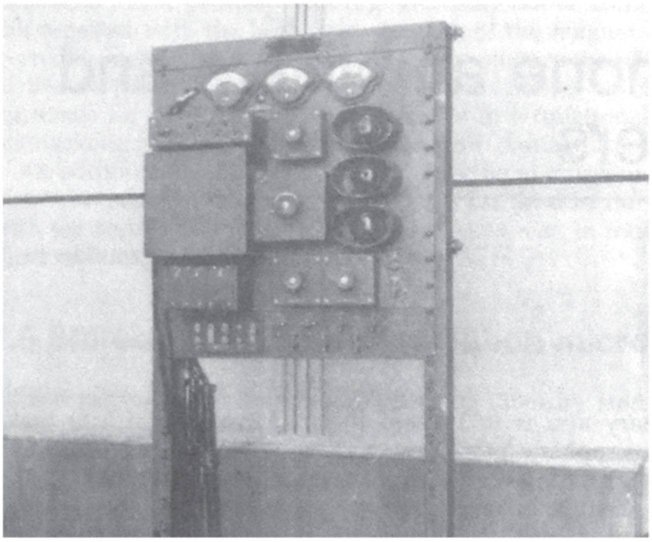


Figure 8.1 *Western Electric Type 8A three-valve microphone amplifier as installed, with added feed-current milliammeters, in Birmingham control room, 1923. There was a separate filament rheostat for each valve and choke – capacitance interstage coupling was employed. Storage batteries were used for the power supply.*

carbon microphones – of the telephone variety – being fed directly to the input of the modulation amplifier in the transmitter. This was the initial state of affairs at the original London station 2LO as installed by the Marconi Company at Marconi House in the Strand, though it did not last long.

The much superior quality of the Birmingham transmissions was noticed by listeners – perhaps the first faint glimmering of a hi-fi outlook! – and before long Western Electric carbon microphones and amplifiers were introduced at other stations. Nevertheless it became evident that even these had fairly serious performance shortcomings, and this, together with their very high cost and the fact that microphones and amplifiers were obviously going to be needed in much larger numbers as the broadcasting service expanded, resulted in a good deal of effort being put into finding other solutions.

Captain H. J. Round at Marconi's, who was a key figure in the early days of broadcasting in Britain (and with whom I once spent a most enjoyable day when he came to see some work we were doing at the Royal Signals and Radar Establishment about 1948), soon evolved the Marconi–Sykes magnetophone, the essential notion coming from a 1920 patent of Adrian Sykes. This was a form of moving-coil microphone whose diaphragm was a flat annular coil of aluminium wire suspended on cotton wool in the field of a large pot-type electromagnet consuming 4 A at 8 V from a battery.

An early prototype of the magnetophone, and the amplifier which Round had designed to go with it, were brought into regular service in May 1923, just after the BBC London studios at Savoy Hill were occupied. The designs were finalized within a few months and became the standard equipment in most BBC studios for several years. The microphone was widely known as the meat-safe!

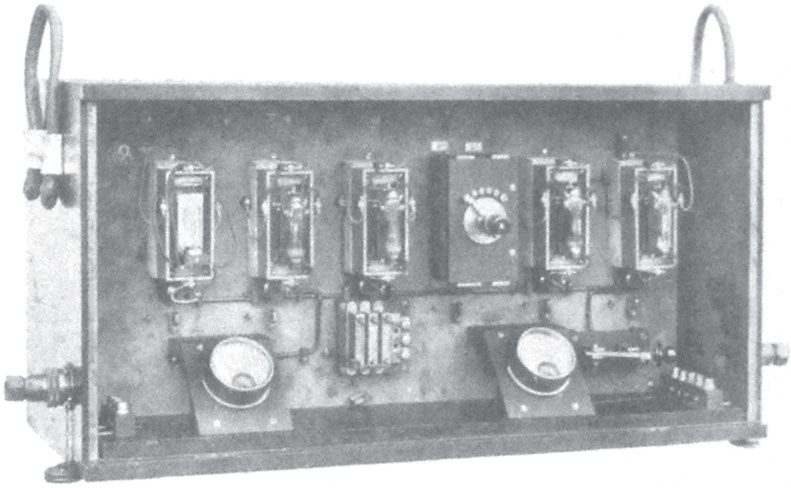


Figure 8.2 *Marconi Type GA1 single-channel five-valve microphone amplifier, having resistance – capacitance interstage coupling. The metal screening boxes over the valves have been removed, but can be seen in Figure 8.3. This was the standard BBC microphone amplifier for several years. It operated from a 6 V filament supply and a 300 V HT supply.*

The production version of the microphone amplifier, Marconi Type GA1, is shown in Figure 8.2, with the metal screening covers over the valves removed. The amplifier weighed over 45 kg, but was considerably smaller than the original experimental prototype. There were five R–C coupled stages of triode amplification, with input and output transformers mounted behind the valve panel. The price was only about half that of the Western Electric Type 8A.

Each valve, connected by gold-plated contacts, was mounted in a sub-assembly suspended by rubber bands, with flexible leads to the main circuit. This was done to reduce microphony, or 'ponging' as it was often called, which occurred if the valves were subjected to the slightest vibration – this was indeed a major problem with directly heated valves of the types then employed.

The GA1 amplifier had provision for adjusting the high-frequency response 'to suit the acoustic properties of the hall in which the microphone is used', to quote from the leaflet supplied.[2]

The magnetophone and GA1 amplifier were used not only in studios but also for outside broadcasts, as also were the Western Electric items previously mentioned. Figure 8.3, taken in 1925, shows two of the Marconi amplifiers rigged for one of the famous early OBs of the song of the nightingale from a wood in Oxted, Surrey.[2] Notice the boxes of batteries and the checking radio receiver.

Though the above Marconi equipment was in widespread BBC use by 1924, Birmingham continued to use Western Electric products, soon changing from the double-button carbon microphone to a capacitor microphone, which was retained there until 1926. It was the only capacitor microphone in BBC service. The microphone and its amplifier 'were both very susceptible to the least trace of dampness and so were apt to be "temperamental" '. Though capacitor microphones were employed to a very limited extent after 1926, it was many years before their early BBC reputation for making frying noises and exhibiting general unreliability was largely overcome.

Figure 8.3 *Two Marconi GA1 microphone amplifiers installed for a 1925 outside broadcast of the song of the nightingale from a Surrey wood. A 6 V battery and six wooden boxes containing 50 V batteries may be seen.*

A very significant event in the history of microphones and microphone amplifiers in the BBC was the introduction of the Marconi–Reisz transverse-current carbon microphone – the familiar marble-block device of octagonal shape that is seen in so many early broadcasting photographs. It was invented in Germany by Georg Neumann, founder of the present-day firm of that name, while employed by the Reisz company.[3] It was brought into service towards the end of 1925, and was in almost universal use throughout the BBC by 1927. Many of these microphones, together with a BTH version of the same type but round in shape, were to be seen in action until 1935, and to a limited extent even later.[2,4]

The Reisz microphone was robust, extremely reliable, acceptably small by the standards of those days, it had a convenient (resistive) impedance of about 300 Ω, and, above all, it was highly sensitive, giving well over 10 mV r.m.s. on the grid of the input valve even on speech. The bass response extended without loss down to 30 Hz and below. The axial high-frequency response exhibited a very broad peak of over 10 dB centred around 4 kHz, much of this being due to diffraction. However, with suitable equalization, which was incorporated in the later amplifier designs, and with appropriately positioned artists, quite good sound quality was obtainable.

The microphone took about 20 mA at 6 V DC to polarize it, and generated considerable noise, of $1/f$ spectrum. The noise, expressed as an equivalent sound pressure level (SPL), though a good deal higher than for a normal modern microphone, was not intolerably great, but bearing in mind the high sensitivity, it corresponded to quite a high electrical output noise level, which over-rode the thermal noise in the amplifier input circuit by a substantial factor. This greatly eased the problems of amplifier design from a noise and microphony viewpoint and it also reduced the audibility of interference picked up by microphone wiring. The microphone unfortunately gave considerable non-linearity distortion at very high SPLs, but its limitations were well understood by the engineers of the day and by-and-large it gave very satisfactory results on medium-wave radio.

Figure 8.4 shows, in redrawn and slightly simplified form, the circuit of a Reisz microphone amplifier as given in Reference 5 of 1928. The first three valves are 2 V filament valves of fairly low gain as then used in ordinary radio sets, the output valve being a small power valve with 6 V filament. Note the arrangement of three flashlamp bulbs whose relative brightness, in association with the meter readings, gave an immediate indication, in the event of a filament failure, of which valve was faulty – such failures were considerably less rare than in later times.

Note too the way the 6 V supply for the filaments serves also for energizing the microphone, and that the microphone circuit is not a balanced one. This latter feature would be frowned upon today, but apparently it was then found good enough, largely because of the high output level of the Reisz microphones.

Obviously gain adjustment effected right at the input as in Figure 8.4 leaves the amplifier output due to valve noise and microphony at its full level when the gain is reduced, but because of the high microphone sensitivity, and consequently only moderate maximum-gain requirement, this scheme was found to be quite satisfactory, and it had the great virtue of keeping the amplifier distortion to a minimum even at the low gain settings required with loud programme sources – negative feedback was not known about when this amplifier was designed.

The purpose of the input gain control, as normally used, was to set the gain to suit the type of programme involved, so that the level at the microphone amplifier output would be suitable for easy handling on the faders at the control desk.

Before the advent of Reisz carbon microphones, when magnetophones and GA1 amplifiers were employed, these amplifiers and their batteries were usually located close to the associated studios, but when Reisz microphones came in, the amplifiers were installed in the control room, allowing all batteries to be located nearby and shared between amplifiers. This new scheme, unlike the old one, involved long cable runs at microphone level, and these, as already mentioned, were operated in an unbalanced mode.

Another feature of the Figure 8.4 circuit is the use of grid stoppers in the first two stages. These markedly reduced the tendency of the circuit to respond, by rectification, to radio-frequency interference, and their use for this purpose has been credited to H. J. Round. This dodge, of course, is still used in the semiconductor era, though sometimes primarily for preventing VHF oscillation.

After 1928, the design of BBC microphone amplifiers, otherwise called 'A' amplifiers, evolved for several years along lines fairly closely related to the Figure 8.4 type of circuit, but by 1933 indirectly heated triodes, which became available about 1930 with the advent of mains-operated radio receivers, had displaced the filament type and were less microphonic.[6]

Though Reisz microphones were still in very widespread use in 1933, they were beginning to be replaced by more modern and sensitive forms of moving-coil microphone, and, to a much lesser extent, by capacitor microphones which had individual preamplifiers.[7,8]

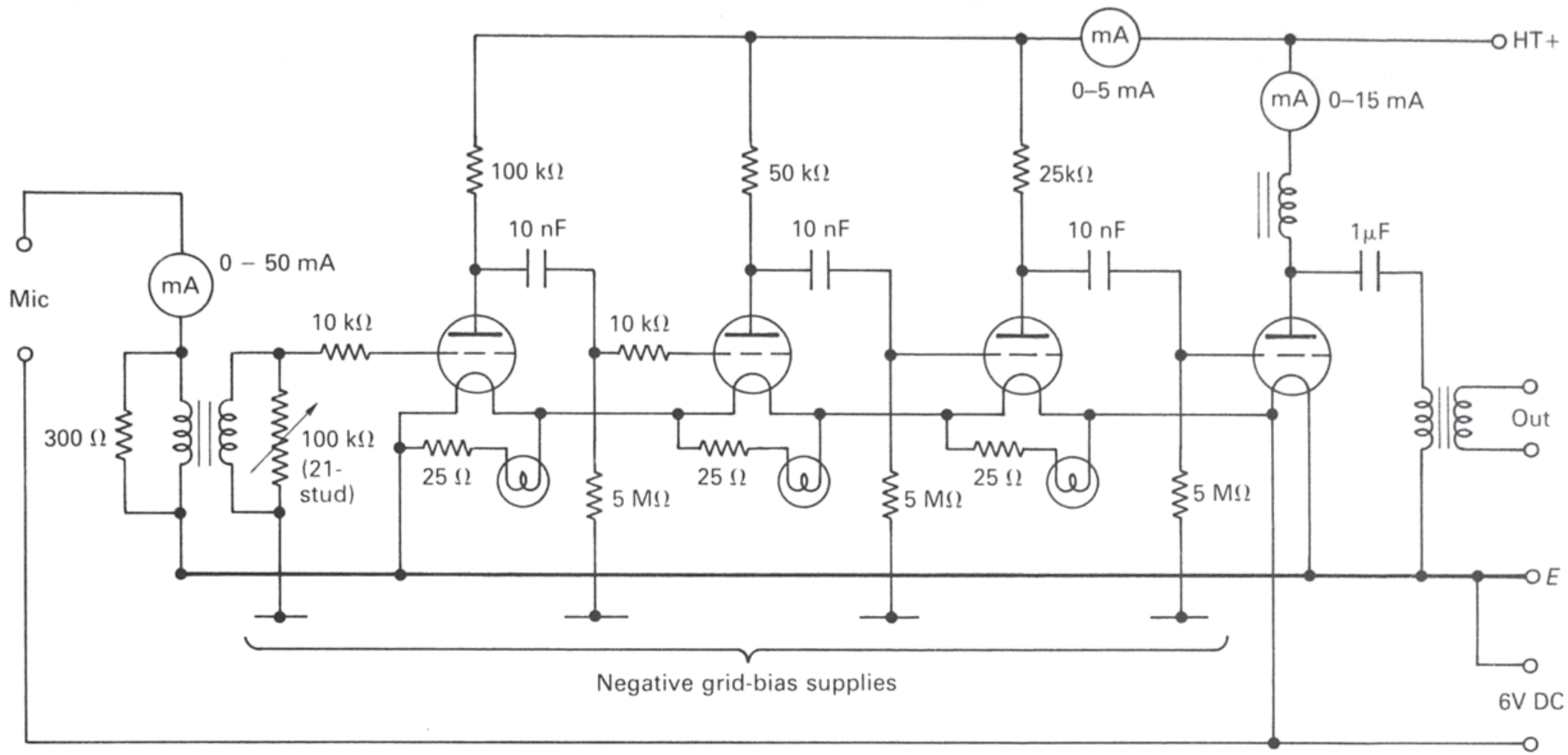


Figure 8.4 *BBC microphone amplifier as used with Marconi-Reisz transverse-current carbon microphones in 1928. Sometimes two microphones in parallel were connected to the input. The low internal resistance of the storage-battery HT supply made early-stage decoupling unnecessary.*

With the new valve types and related transformer designs, the new 'A' amplifiers required only three stages. To compensate for the over-emphasized high-frequency response of the Reisz microphones, a low-*Q* series tuned circuit, resonating at just under 4 kHz and having adjustable series resistance, could be switched across the output of the first stage.

A shelving bass loss below about 300 Hz was also switched in, because of a belief that the Reisz microphones had a lifting bass response of at least +6 dB at 50 Hz. This belief must, I think, have been mistaken; a recent measurement on a microphone of this type, using a B & K Type 4133 measuring microphone, showing it to have a low frequency response flat down to 30 Hz within ±1 dB, as would be expected from its principle of operation. (It is noticeable that around 1932/1933 published BBC microphone response curves,[7,9] not only for carbon microphones but also for stretched-diaphragm and Voigt slack-diaphragm capacitor microphones, all exhibited a fairly similar bass rise, suggesting that the measuring set-up then in use probably had an unsuspected bass loss. This is perhaps less surprising when one bears in mind that acoustic measurement techniques were in their infancy at the time.)

The gain adjustment, for accommodating different types of programme source, came, as before, at the input, though the change had been made to a normally connected potentiometer rather than a shunt 'rheostat', and the primary of the input transformer was floating so that it could be used in either a balanced or an unbalanced manner.

Despite the use of indirectly heated valves in the new amplifiers, the heater, high-tension and grid-bias supplies were all still obtained from central batteries – there seems to have been a profound distrust of mains supplies for amplifiers by broadcasting engineers in those days, presumably on grounds of less reliability and propensity to mains hum, the latter being a characteristic of much badly designed public-address equipment at the time!

The next major change was the bringing into service in 1935 of the BBC/Marconi Type A ribbon microphone, after a short trial period. By 1936 the majority of BBC microphones were of this type. The ratio of the internal transformer was chosen to give an output impedance of 300 Ω as for the Reisz microphones, which was convenient. Though the new microphones were of much better fidelity and had polar characteristics of a usually beneficial type, the lower output level nevertheless caused acute problems at first.

The cable runs from studios to control room not then being of the balanced variety, studio preamplifiers were initially found necessary with ribbon microphones in order to avoid troubles from switch-click interference etc. However, by changing to balanced twin-core lead-covered cable, plus balanced input transformers with screens, long microphone cable runs, even of 300 m, were found to be quite satisfactory.

As already mentioned, outside broadcasts in the earliest days were mostly done using cumbersome amplifiers designed for studio or control-room purposes, accompanied by massive boxes of batteries, but in later years amplifiers designed specifically for OB use were introduced, of which the OBA/8, shown in Figure 8.5, which appeared in 1938, is the one of by far the greatest technical significance.[4,10,11]

The OBA/8 amplifier was an outstandingly successful design which, for several reasons, had a great influence on later developments in both OB and studio equipment. It remained in service for several decades and was manufactured in rack-mounting form (Type APM1) as well as in the box form used for normal OBs. Large numbers of these amplifiers were used as the basis of temporary wartime studio installations. The main new features, apart from reduced size and weight, were:

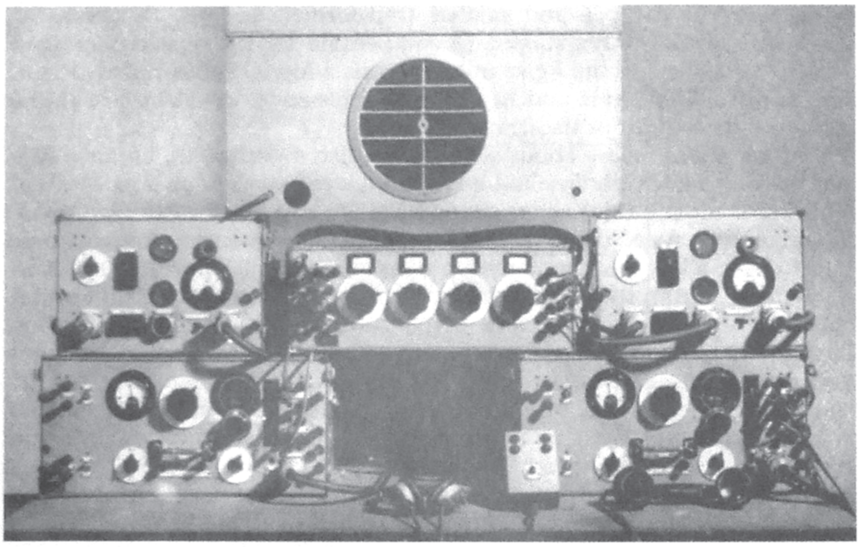


Figure 8.5 *OBA/8 outside-broadcast equipment. Two amplifiers are on the table, surmounted by a pair of power units between which is a four-channel microphone mixer. The monitoring loudspeaker, in a case unfolding to form a baffle board, contained a power amplifier consisting simply of a pair of ACSP/3 valves in push-pull – 'extensive tests indicated that the maximum output of about 1 ½ watts was adequate'.*

(1) Only two valves for signal amplification, these being AC/SP3 high-slope television-type pentodes. The 4 V 1 A heaters were run on 50 Hz AC and the mains-derived HT was at 250 V.
(2) Total HT current only 25 mA, allowing ordinary radio-receiver type dry HT batteries to be used for emergency operation in the event of mains failure, plus a storage battery to provide the total heater current of 3.5 A at 4 V.
(3) Negative feedback employed, with combined feedback and passive gain control.[10,11]
(4) Equalization incorporated to correct for falling response of ribbon microphones at high frequencies.
(5) Built-in peak programme meter provided.

The amplifier was capable of supplying a power output of 25 mW, with not more than about 1% harmonic distortion, to any resistive land-line impedance between 75 Ω and 600 Ω , and had sufficient available gain to be able to do this, using ribbon microphones, for any likely kind of programme material. At reduced settings of the single-knob gain control, the amplifier could handle inputs corresponding to ribbon microphone SPLs of about 133 dB* with similarly low distortion and without the sacrifice of signal-to-noise ratio which the simple input gain-control scheme of Figure 8.4 would have produced. The distortion under most operating conditions was substantially less than 1% and subjectively negligible.

*The later Type AXBT version of the Marconi ribbon microphone had a Ticonal magnet and was several dB more sensitive,[12] giving a lower maximum SPL for 1% distortion.

On outside broadcasts and in studios it is, of course, frequently required to mix the outputs of several microphones. Nowadays, since microphone amplifiers are small and relatively cheap, the almost universal practice is to have a separate amplifier for each microphone and do the mixing at a moderately high signal level. But in earlier days the mixing was often done at low level, with an amplifier shared between several microphones. Thus a standard adjunct to the OBA/8 amplifier was a four-channel mixer or series fader. Each control knob on this unit varied the amount of series resistance inserted between the associated 300 Ω microphone and the OBA/8 input, the input impedance also being 300 Ω.

If one microphone was fully faded up, with the others faded out, there was no sacrifice of signal-to-noise ratio, nor was there if more than one was fully faded up, but with all knobs at less than the maximum setting, the signal-to-noise ratio became significantly inferior to what can be achieved by using a separate amplifier for each microphone and doing the mixing at a higher level. However, by judicious use of the main gain control on the OBA/8 amplifier in conjunction with the mixer controls, excellent results could be obtained, though this procedure was less than ideally convenient for the operator.

A further feature of this low-level mixing scheme is that the level of the contribution from one microphone is reduced when further microphones are faded up.

Before the advent of the OBA/8 equipment, rather surprisingly in retrospect, the control of the dynamic range of outside broadcasts had normally been done in the main control room, the gain of the OB amplifier being kept constant. This method of working was far from optimum with respect to OB amplifier distortion and land-line signal-to-noise ratio, and the much sounder policy of effecting the dynamic-range control at the OB point was introduced when the OBA/8 amplifier came into use. The built-in peak programme meter enabled this to be done in an optimum manner.

The development of mains-operated equipment of the OBA/8 type, plus the very reassuring experience gained by using it in numerous temporary war-time studios, resulted in the abandonment after the war of the centralized control-room location of microphone or 'A' amplifiers and the post-fader 'B' amplifiers. The studio then became a self-contained unit so far as amplifier equipment was concerned. In the earlier post-war years of austerity, OBA/8 and APM1 amplifiers continued to be widely used, pending the introduction of new and more versatile equipment specifically intended for studio installations. Thought had in fact been given to the design of such apparatus during the later part of the war, leading to what became known as Type A studio equipment.[13,14] Though this began its service trials in December 1944, it was not until the mid 1950s that it had replaced virtually all earlier equipment in studios.

The Type A equipment, which was still based on the pre-war ACSP/3 pentode valve, had relatively small amplifier modules, mounted on sliding shelves in lockable cabinets in such a way as to be easily removable for servicing.

A new feature was that a separate 'A' amplifier was used for each microphone or gramophone source, low-level mixing being abandoned. A jack field, accessible without unlocking the main cabinet doors, enabled the output of each 'A' amplifier to be fed to any chosen fader, to suit operational requirements.

With the amplifiers at a distance of some metres from the control desk, the feeds to and from the faders had to be at low impedance, balanced stud-type constant-impedance bridged-T faders being employed, with 600 Ω input and output transformers in all amplifiers – a clean technique, although rather expensive.

The use of constant-impedance faders fed in this way from 600 Ω source impedances allowed passive mixing to be achieved simply by parallel-connecting the fader outputs, and the scheme avoided the interaction between controls that was a characteristic of the simple series-fader arrangement as carried out with OBA/8 amplifiers.

To secure a good output signal-to-noise ratio, with a considerable number of inputs and allowing some control range to spare – so that none of the faders would normally need to be set at maximum – the individual microphone or 'A' amplifiers had to have sufficient gain to supply a fairly high input level to the faders.

The power-supply switching for the amplifiers was controlled by push-buttons on the control desk and standby 'A' and 'B' amplifiers were provided which, by operating the appropriate desk keys, could be instantly substituted for any amplifier unit that had gone faulty.

Many changes have occurred in the design of microphone amplifiers and associated equipment since the first post-war designs mentioned above came into service. These changes have resulted not only from far-reaching developments in the general field of electronic technology but also from changing attitudes concerning the style of presentation of many kinds of sound programme. The introduction of stereo broadcasting, which began on a regular programme basis in July 1966, has also of course had a major influence on equipment design.

The general trend, for better or for worse, has been in the direction of far greater system complexity, coupled with a requirement that the equipment should be of more flexible design than in the past, allowing it to be more readily adapted to changing operational needs.

The first of the new and more economical equipment introduced into BBC studios from about 1954 to replace Type A was known as Type B,[15] whose main new features were:

(1) Miniature components used throughout, including CV455 (inter-services equivalent of ECC81 or 12 AT 7) 9-pin glass-based double-triode valves in all stages. Each pair of triodes required less than half the heater power of a single AC/SP3 pentode, permitting economical push–pull output stages to be employed. This eliminated DC polarization of the cores of the transformers used to provide balanced 600 Ω outputs, leading to a substantial reduction in the size and cost of these transformers.
(2) The amplifiers were built as small plug-in rack-mounting units of standardized size, so easily removable for servicing that the jack sockets or meters for anode-current checking incorporated in earlier amplifiers were deemed unnecessary and were omitted.
(3) Plugging of specific microphone lines to specific faders carried out on a jack field preceding the microphone amplifiers instead of coming after them as in Type A equipment.
(4) Control-desk modules built on standard-sized sub-panels, permitting a variety of functionally different control desks to be quickly assembled to suit particular operational requirements, and altered if these requirements changed.

This Type B equipment still followed the philosophy of accommodating the electronics on racks or cabinets separate from the control desk.

The pace of development in electronic technology towards the end of the 1950s was such that both the above studio equipment and also a fairly new valve design of outside broadcast equipment, the OBA/9,[16] began to seem obsolescent fairly soon after going into service. This was mainly because, by about 1955, germanium PNP junction transistors had reached the stage at which their characteristics were good enough to enable microphone and other amplifiers to be designed with a performance that would be acceptable for high-quality broadcasting purposes.

The BBC reacted quite quickly to this development and by May 1957 prototype circuits for transistor outside broadcast amplifiers and peak programme meters were undergoing service trials. The same circuit design work also led to what became known as Type C studio equipment,[17,18] which underwent trials in two drama studios in Broadcasting House Extension, opened in 1961. The units operated off a single mains-derived stabilized supply voltage of 24 V, positive earthed. The microphone amplifier noise figure was initially about 5 dB, which is at least 3 dB worse than for a good valve amplifier, but this situation soon improved as lower-noise transistors became available.

A further most significant feature of this new Type C studio equipment was that it adopted the now almost universal practice of accommodating the amplifiers within the control desk. One of the arguments against doing this in the valve era was that control desks are subjected to a certain amount of mechanical vibration when in use, which may include accidental kicking, and this would be liable to produce audible sounds due to valve microphony – transistors, however, are not appreciably microphonic.

Incorporating the amplifiers in the desk, in addition to saving floor space, had the advantage of eliminating the long cable runs to faders that were previously necessary, and this made it possible to use good-quality carbon potentiometers in place of expensive stud-type constant-impedance balanced faders and associated transformers.

The Type C equipment included 'means of manipulating the frequency response of a microphone channel to obtain the effects required in the production of some kinds of light music'.[18] A bass and treble negative-feedback tone-control circuit, which I published in 1952,[19] was employed, plus a tuned circuit to give an optional 'presence boost' of 3 or 6 dB centred on 2.8 kHz.

The provision of the above controls was indicative of a new trend in attitudes, largely instigated by the pop recording industry, and it has continued to the present day. Such frequency-response manipulation is often called equalization or 'EQ'. Logically, however, the word 'equalization' implies correction of a non-flat frequency response, for example of a land-line, or a tape replay head, or the bass rise given by a directional microphone on close speech, so that its use in relation to tone controls, presence circuits etc., whose function is artistic (sometimes inartistic!) seems rather unfortunate.

Another trend in control desk design, for both mono (largely TV) and stereo, has been to have much larger numbers of microphone channels than in the past – often well in excess of thirty – and this has been facilitated by the development of small channel modules, using either discrete transistors or integrated circuits and each including the basic microphone amplifier, plus the frequency-response manipulating controls, the routing switches etc. and the slider-type fader.

The BBC Type C equipment, which was monophonic and had rotary faders, came before the above-mentioned modern type of channel-module format had been evolved, but it represented a most praiseworthy pioneering stage in the development of simple yet elegant circuit and constructional techniques for transistor sound equipment.

Figure 8.6 shows a Type C microphone amplifier unit and its circuit diagram as given in Reference 18. The diagram is drawn on a 'collectors upwards' basis, even though this necessitates having the negative supply line at the top and conventional current flowing upwards. An issue is here involved that was highly controversial at the time – I strongly favoured the 'positive upwards' convention.[20] However, this issue largely resolved itself later when the NPN silicon planar transistor became the dominant active device – diagrams could then be both 'positive upwards' and 'collectors upwards' and yet have the earth line at the bottom.

The 10 kΩ potentiometer shown in Figure 8.6(b) was mounted externally to the unit and connected by pins on the plug-in connector at the back. In some applications it might be the channel fader, whereas in others it might be replaced by an external pair of resistors to give a fixed gain, or by a switched network of gain-setting resistors.

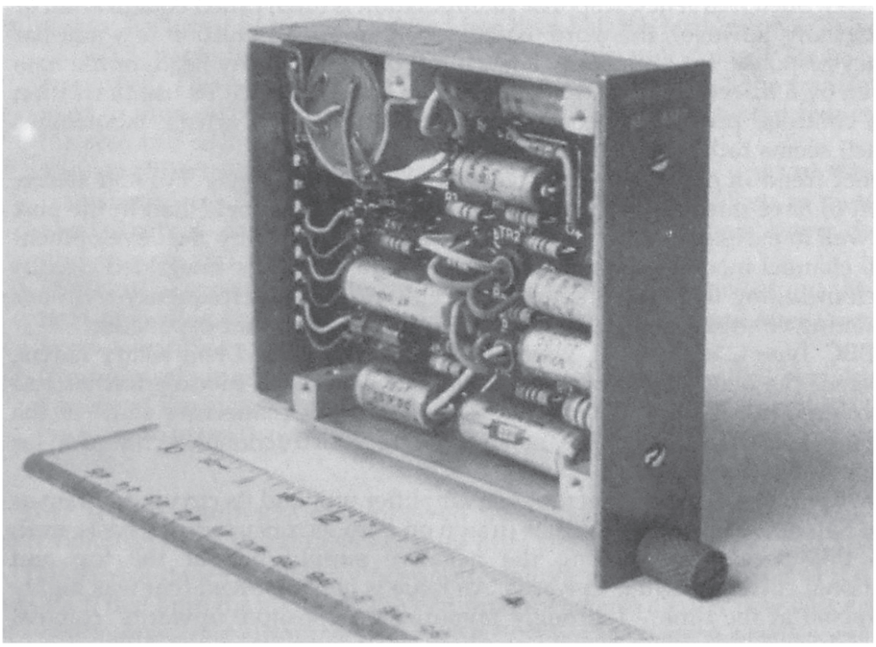
The work on the Type C equipment was followed by further BBC work which led to new transistor studio equipment with even better performance, known as Type D,

The advent of stereo during this period had a profound effect on BBC studio equipment activities, as might be expected. At first the fairly small number of stereo programme contributions were handled by making a few special stereo desk units for use in association with Type B valve amplifiers. These had quadrant-type faders mounted as 'outrigger units' on the desk surface in front of the other controls.

Though the above-mentioned Type D equipment was initially conceived on a monophonic basis, it was nevertheless adapted to full stereophonic operation and was installed, for example, in the sound radio studios at Pebble Mill, Birmingham, which opened in 1971.

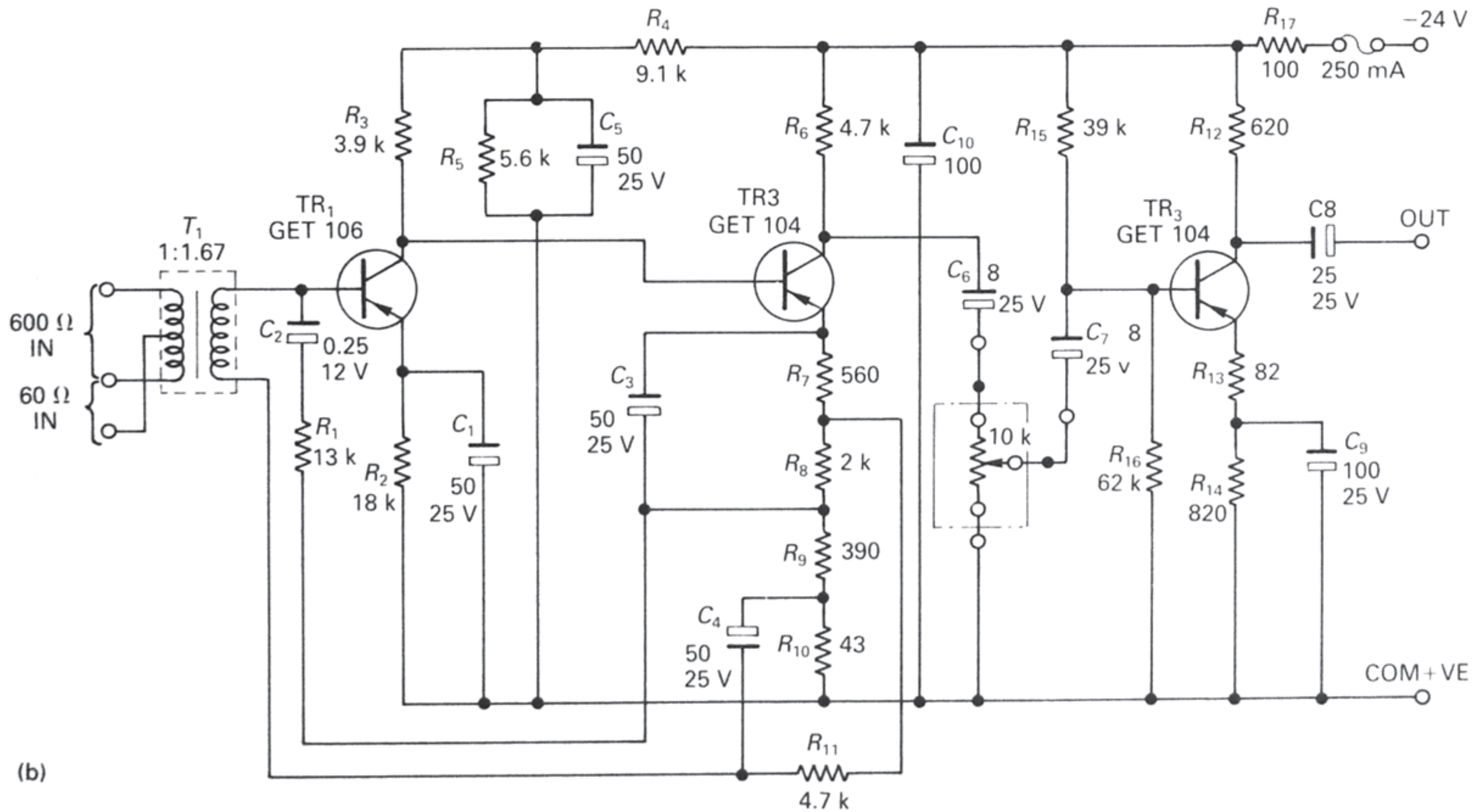
The arrival of stereo, in combination with other influences, both technical and political, has tended to shift most of the detailed design work on new sound equipment in recent years from the BBC to industry.

The proliferation of independent recording studios employing multi-track recording techniques, allied with the great increase in the number and complexity of the sound facilities connected with TV and sound broadcasting, both national and local, has brought into being a large and flourishing industry for the



(a)

Figure 8.6 (a) *Type C microphone amplifier unit.*



(b)

Figure 8.6 (b) *Circuit of Type C microphone amplifier unit as given in Reference 18 of 1963.*

manufacture of control desks, or mixer desks as they are now more frequently called.

The faders in these commercial desks normally have conductive-plastic elements, giving smooth and continuous variation of gain, and are of the linear-motion rather than quadrant type. The mixing is done on a virtual-earth basis,[21] which effectively eliminates any interaction between controls.

The BBC, of course, still has complex and exacting requirements for new sound equipment, both for studios and for outside broadcasts, but its role in fulfilling these needs has tended to become more one of drawing up specifications and carrying out approval testing on the resultant commercial products than in doing the major part of the detailed design work.

Though the above account has been closely related to BBC practice – largely because this is so well documented – the trends outlined are broadly representative of developments throughout world broadcasting and recording organizations.

8.3 Some facts about random noise

Though man-made interference is sometimes included within the meaning of the term noise, only truly random noise of natural origin – usually perceived as a gentle hiss or mush in the background – is considered in this Section, which presents some basic noise concepts underlying the design and use of microphone amplifiers.

8.3.1 Noise waveforms

Random noise waveforms normally have Gaussian statistical characteristics, and one consequence of this is that for 99.7% of the time the peak instantaneous value does not exceed 3 times the r.m.s. value.

Another fact about Gaussian noise is that if it is measured on an ordinary mean-rectifier a.c. voltmeter, the true r.m.s. voltage is $2/\sqrt{\pi}$ or 1.128 times the scale reading. (If this voltmeter is of the electronic variety, care must be taken to ensure that its circuitry is not overloaded by instantaneous inputs up to at least 3 times the r.m.s. input.)

An entirely separate attribute of random noise is its frequency spectrum, which shows how the mean squared noise voltage or current per unit bandwidth varies with frequency. If this spectrum is flat over the frequency band of concern, then the noise is called white noise, by analogy with white light.

The digitally generated white noise tracks on test CDs are liable to be non-Gaussian, the probability density function being somewhere between Gaussian and rectangular.

When two random noise voltages V_{N_1} and V_{N_2} are produced by totally independent systems or components, they are said to be uncorrelated, and when they are added electrically the total r.m.s. value V_{N_T} is given by

$$V_{N_T} = \sqrt{(V_{N_1}^2 + V_{N_2}^2)} \tag{8.1}$$

If, on the other hand, V_{N_1} and V_{N_2} are derived from the same original source, via equal time delays if these are appreciable, then they are said to be 100% correlated and their sum is obtained by simple algebraic addition, as for any other pair of voltages with identical waveforms.

8.3.2 Johnson noise

One cause of random noise in microphone amplifier equipment is thermal agitation or Johnson noise, produced in every resistor or other resistive element. It may be represented by a noise voltage V_N acting in series with the resistance and given by

$$V_N = \sqrt{(4kTBR)} \tag{8.2}$$

where V_N = r.m.s. noise voltage (V)
k = Boltzmann's constant, 1.38×10^{-23} (J/°C)
T = absolute temperature (K)
B = effective bandwidth over which the noise is measured (Hz)
R = resistance (Ω)

It is helpful to remember one specific result, such as that $V_N = 1.0\,\mu$V r.m.s. for $R = 3\,\text{k}\Omega$, $B = 20\,\text{kHz}$ and $T = 300\text{K}$ (27°C).

The Johnson noise current that flows when a pure resistance R is short-circuited is given by

$$I_N = \sqrt{\left(\frac{4kTB}{R}\right)} \tag{8.3}$$

where I_N = r.m.s. noise current (A), and the other quantities are as before.

8.3.3 Shot noise

Another cause of random noise in valves and semiconductor devices is shot noise, which arises because the currents flowing within them involve the rapid* and mutually independent passage of very large numbers of identical electronic charges, each producing a tiny output pulse of very short duration. Like Johnson noise, shot noise is normally both Gaussian and white, and its magnitude is given by

$$I_N = \sqrt{(2qI_{dc}B)} \tag{8.4}$$

where I_N = r.m.s. shot-noise current (A)
q = electronic charge, 1.60×10^{-19} (C)
I_{dc} = d.c. current (A)
B = noise bandwidth (Hz)

8.3.4 Flicker noise

In addition to Johnson noise and shot noise, practical amplifying devices also produce flicker noise, whose spectral density is inversely proportional to frequency – in other words, the mean squared noise voltage or current per unit bandwidth, instead of being constant as for white noise, varies as $1/f$, rising, therefore, at 10 dB/decade or 3 dB/octave with falling frequency. Flicker noise is otherwise known as $1/f$ noise, pink noise, or excess noise, and for such noise it is the mean-squared noise voltage per octave that is independent of frequency.

*In great contrast, the charges in copper conductors normally move at less than 1 mm/s.

Flicker noise is greatly influenced by slight imperfections in manufacturing techniques and varies much more from sample to sample than does shot noise. Advances in technology, however, have brought flicker noise under much better control than was once the case.

The noise output component from a microphone amplifier due to flicker noise, though not white, is normally Gaussian.

8.3.5 Popcorn noise

Another type of noise sometimes exhibited by semiconductor devices is so-called popcorn or burst noise, whose waveform is far from being Gaussian – as the name suggests, it is of a sporadic or pulse-like nature. With earlier transistors and op. amps. it was fairly frequently encountered, but improved manufacturing techniques have made it comparatively rare nowadays.

8.3.6. Representation of total amplifier noise

Ignoring popcorn noise, the random noise generated inside a microphone amplifier is due to a combination of Johnson, shot and flicker noise, but may always be correctly represented by equivalent voltage and current noise generators placed right at the input of an imaginary noiseless amplifier with other characteristics identical to those of the actual amplifier, as shown in Figure 8.7.

With the input terminals short-circuited, only V_N produces any noise output from the amplifier; with the input open-circuited, V_N has no effect but I_N now gives rise to a noise output by feeding its current into the amplifier input impedance Z_{in}, thus producing an input voltage.

With a signal source of finite impedance connected, both V_N and I_N produce noise output, and their fixed values, plus the degree of correlation between them, can always be chosen to give perfect simulation of the manner in which the noise output due to the amplifier's internal noise mechanisms is influenced by the impedance of the signal source. Additional Johnson noise is, of course, produced by the signal source itself – unless this is purely reactive, or at zero absolute temperature. The correlation between the V_N and I_N noise generators is normally very slight in audio-frequency amplifiers and can be ignored.

V_N and I_N at the input of a well-designed microphone amplifier are mainly due to the input transistors themselves, augmented as little as possible by Johnson noise from associated resistors and later amplifying stages.

Just as the noise produced by a complete amplifier may be accurately represented by voltage and current noise generators right at the input, so also may be the noise produced by an isolated transistor or op. amp.

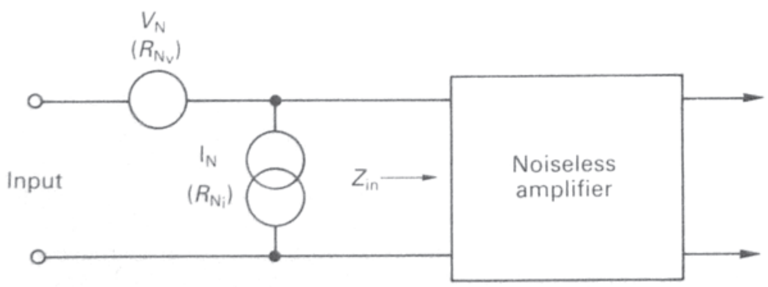


Figure 8.7 *Equivalent voltage and current noise generators representing the total noise produced within an amplifier.*

8.3.7 The R_{N_v} and R_{N_i} concept[22,23]

In data sheets for semiconductor devices, information on voltage and current noise is often given in nV/√Hz and pA/√Hz respectively. These data-sheet quantities are the values of $V_N/\sqrt{B}$ and $I_N/\sqrt{B}$ at the device input, and are sometimes denoted by the symbols e_n and i_n.

As will become apparent, there are very real advantages to be gained by expressing device noise, and also total amplifier noise, in the alternative form of equivalent room-temperature Johnson-noise-producing resistance values R_{N_v} and R_{N_i}, and from equations (8.2) and (8.3) it follows that

$$R_{N_v} = \frac{[V_N/\sqrt{B}]^2}{4kT} \tag{8.5}$$

$$R_{N_i} = \frac{4kT}{[I_N/\sqrt{B}]^2} \tag{8.6}$$

Figure 8.8 presents the relationships (8.5) and (8.6) in a convenient form, enabling the R_{N_v} and R_{N_i} values, corresponding to given data sheet nV/√Hz and pA/√Hz figures for $V_N/\sqrt{B}$ and $I_N/\sqrt{B}$, to be quickly read off.

R_{N_v} and R_{N_i} do not represent actual resistances appearing in the circuit – they merely constitute a convenient method for expressing the magnitudes of the noise voltage and noise current generators, as shown in Figure 8.7. V_N is the room-temperature open-circuit Johnson-noise voltage produced by R_{N_v} and I_N is the short-circuit Johnson-noise current produced by R_{N_i}. Representing voltage noise by an equivalent resistance R_{N_v} was a well-established notion in the valve era, but the use of R_{N_i} to represent semiconductor current noise seems to have been first proposed in 1966 by Dr E. A. Faulkner of Reading University,[23] and its more widespread adoption is much to be recommended.

If only Johnson noise and shot noise were present within the devices, then R_{N_v} and R_{N_i} would be independent of frequency throughout the whole audio band,

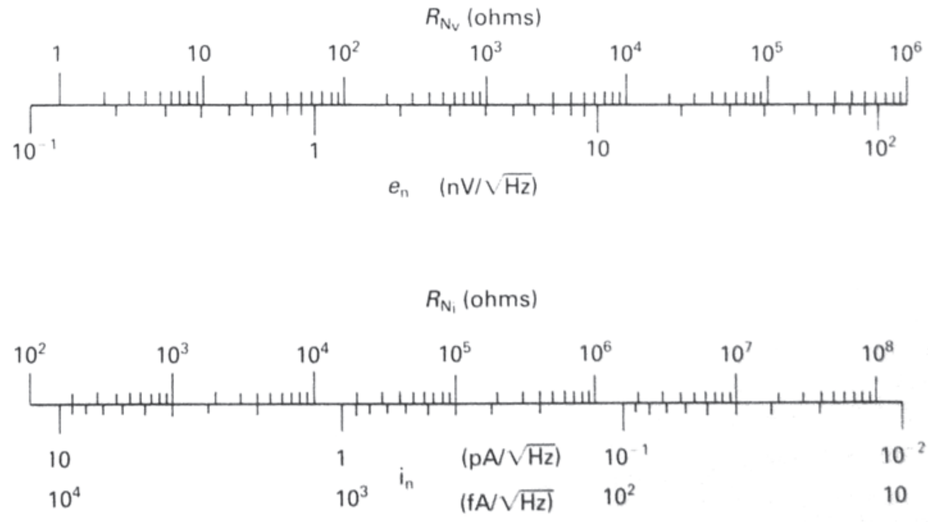


Figure 8.8 *Conversion scales (20°C).*

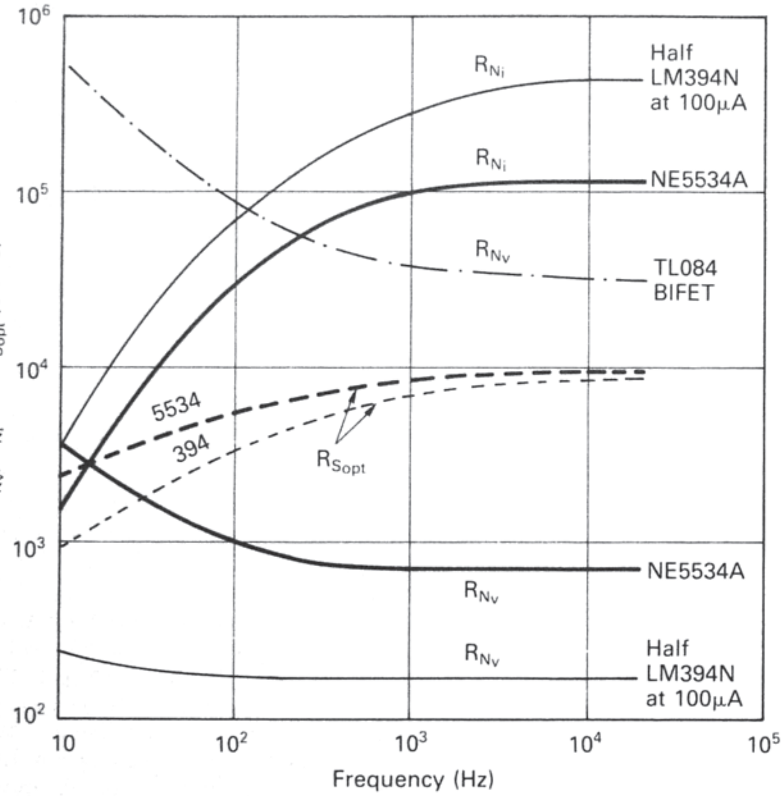


Figure 8.9 *R_{N_v}, R_{N_i} and $R_{S_{opt}}$ curves for representative devices.*

but, as shown in Figure 8.9, flicker noise in ordinary bipolar transistors causes a reduction in R_{N_i}, corresponding to an increase in the pA/√Hz figure, below a certain frequency, and at a normally much lower frequency still, a rise in R_{N_v}.

With field-effect transistors (FETs) on the other hand, flicker noise shows up mainly on the R_{N_v} curve, with negligible reduction, even at 20 Hz, in the very high R_{N_i} value. See TL084 curve in Figure 8.9.

8.3.8 Noise figure

An important concept when amplifiers are fed from passive resistive signal sources, to which moving-coil and ribbon microphones approximate, is that of noise figure, which is usually expressed in decibels and defined as in equation (8.7):

$$\mathrm{NF} = 10\log_{10}\left[\frac{\text{Total noise output power from amplifier}}{\text{Noise output power due to source Johnson noise alone}}\right] \tag{8.7}$$

Since the output power is proportional to the mean-squared output voltage, an alternative definition is

$$\mathrm{NF} = 10\log_{10}\left[\frac{\text{Total mean-squared noise output voltage}}{\text{Mean-squared noise output voltage due to source Johnson noise alone}}\right] \tag{8.8}$$

Neglecting any slight correlation between V_N and I_N, it may be shown that*

$$\mathrm{NF} = 10\log_{10}\left[1 + \frac{R_{N_v}}{R_S} + \frac{R_S}{R_{N_i}}\right] \tag{8.9}$$

In this equation R_S is the source resistance, and R_{N_v} and R_{N_i} are the values applying to the complete amplifier – they may or may not be approximately equal to the values for the input device in isolation, depending on the circuit design details.

It is evident from the form of equation (8.9) that there must be an optimum value of R_S which will give a minimum noise figure, and this is given by

$$R_{S_{opt}} = \sqrt{(R_{N_v} R_{N_i})} \tag{8.10}$$

(Alternatively the value of $R_{S_{opt}}$ in kilohms may be obtained by dividing the data-sheet nV/√Hz figure by the pA/√Hz figure.)

It is also apparent from equation (8.9) that for this optimized noise figure to be low, as normally desired, R_{N_i} must be much larger than R_{N_v}, the actual relationship being

$$\mathrm{NF}_{opt} = 10\log_{10}\left[1 + 2\sqrt{\frac{R_{N_v}}{R_{N_i}}}\right] \tag{8.11}$$

These facts may be related to features of Figure 8.9 as follows:

(1) The broken-line curves representing $R_{S_{opt}}$ come exactly midway between the corresponding R_{N_v} and R_{N_i} curves.
(2) Wide spacing of the R_{N_v} and R_{N_i} curves makes the noise figure very low when $R_S = R_{S_{opt}}$, and tolerably low over a very wide range of R_S values.

For the NE5534A, NF_{opt} at 1 kHz, from equation (8.11), is 0.7 dB, corresponding to $R_{S_{opt}} = 8.7\,\mathrm{k}\Omega$, and from equation (8.9) the noise figure is 2 dB or better for any value of R_S from 1.3 kΩ to 57 kΩ. For comparison, the values for one half of an LM394N super-matched pair, at $I_{dc} = 100\,\mu\mathrm{A}$, are $\mathrm{NF}_{opt} = 0.2$ dB for $R_{S_{opt}} = 6.8\,\mathrm{k}\Omega$, and NF = 2 dB or better for R_S values from 310 Ω to 150 kΩ.

For early germanium transistors, the R_{N_i} and R_{N_v} curves were much less widely spaced, making it essential, for a tolerably good noise figure, to operate with R_S values not departing very far from $R_{S_{opt}}$. One reason for the relatively poor performance of these transistors is that the collector-to-base leakage current, I_{c_o},

*The presence of Z_{in} (noiseless) in Figure 8.7 affects signal and noise equally, so has no influence on the noise figure. It is convenient to assume either $Z_{in} = 0$ or $Z_{in} = \infty$ in deriving equation (8.9).

often gave a substantial increase in current noise. With silicon transistors, however, such leakage currents are normally quite negligible.

With ordinary op. amps. the operating current of each input transistor is, of course, predetermined (it is about 160 µA for the NE5534A), but when discrete transistors are used, a suitable current must be decided upon by the circuit designer, and this has a large influence on the values of R_{N_v} and R_{N_i} obtained.

8.3.9 Theoretical prediction of transistor noise

Noise information, sometimes only very scanty, and unfortunately often rather erroneous, is given in various forms in manufacturers' data sheets, but it is helpful to appreciate that at frequencies high enough for flicker noise to be neglected, the approximate values of R_{N_v} and R_{N_i}, and the manner in which they vary with operating current, can often be satisfactorily estimated on a theoretical basis.

Although the expressions involved are a little cumbersome at radio frequencies, where operating currents are liable to be quite high, nevertheless at the relatively low currents and frequencies involved in microphone amplifiers, the following very simple equations are applicable with adequate accuracy:

$$R_{N_v} = r_{bb'} + \frac{1}{2g_m} \tag{8.12}$$

$$R_{N_i} = \frac{2\beta_{dc}}{g_m} \tag{8.13}$$

where R_{N_v} = equivalent resistance representing voltage noise (kΩ)
R_{N_i} = equivalent resistance representing current noise (kΩ)
$r_{bb'}$ = base spreading resistance (kΩ)
g_m = mutual conductance of the intrinsic transistor (mA/V)
β_{dc} = d.c. current gain, I_c/I_b.

Figure 8.10 illustrates the significance of equations (8.12) and (8.13). The base spreading resistance $r_{bb'}$ comes between the base lead b and the effective base terminal b′ of the 'inner ideal transistor' or 'intrinsic transistor' as it is usually called. The mutual conductance of the intrinsic transistor is given by

$$g_m = \frac{qI_c}{kT} \tag{8.14}$$

where I_c = collector current (A), the other quantities being as given under equations (8.2) and (8.4). g_m is 40 mA/V at I_c = 1 mA.

The terms $1/2g_m$ and $2\beta_{dc}/g_m$ in equations (8.12) and (8.13) account for the collector-current and base-current shot noise respectively. (Though shot noise is given basically by equation (8.4), which involves q, q has been eliminated from equations (8.12) and (8.13) by exploiting equation (8.14), thus allowing (8.12) and (8.13) to be given in a form often more convenient for practical purposes.)

The Johnson noise produced by $r_{bb'}$ has been allowed for in Figure 8.10 by increasing the value of R_{N_v} correspondingly, $r_{bb'}$ then being regarded as a noiseless resistance.

R_{N_v} and $r_{bb'}$ can obviously be shown in the opposite sequence if preferred, with R_{N_v} immediately next to the *b* terminal, but the R_{N_i} noise generator, unlike that shown in Figure 8.7, inevitably does not come right at the input. However, the

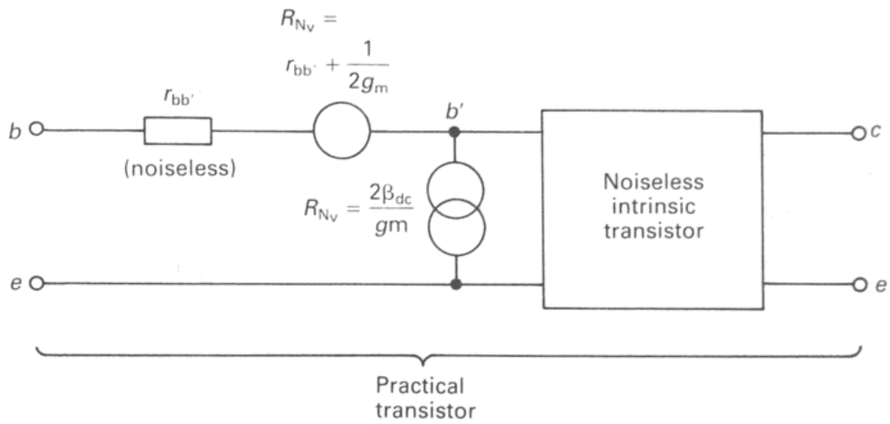


Figure 8.10 *Voltage and current noise generators as given by equations (8.12) and (8.13).*

presence of the noiseless resistance $r_{bb'}$, whose value seldom exceeds about 300 Ω, in series with the input, has a quite negligible effect on the noise performance under all normal microphone amplifier conditions, since the noise voltage-drop across it due to current from R_{N_i} is insignificant compared with the R_{N_v} noise voltage. This is true even at very low frequencies, where flicker noise reduces the value of R_{N_i} and therefore increases the current it produces.

Thus both R_{N_v} and R_{N_i}, as given by the simple equations (8.12) and (8.13), can in practice be taken as coming right at the input as shown in Figure 8.7, without significant error, the noiseless resistance $r_{bb'}$ being simply ignored.

The curves in Figure 8.11 are based on equations (8.12), (8.13) and (8.14), taking $r_{bb'}$ = 300 Ω and β_{dc} (or h_{FE}) = 300, these values being fairly typical for an input transistor such as the BC 109.

An alternative equation to (8.13) for R_{N_i}, which may sometimes be more useful in practice, may be obtained by putting I_{dc} in equation (8.4) equal to the d.c. base current I_b, or bias current as it is usually called with op. amps. This gives I_N, which may then be expressed as an equivalent R_{N_i} value using equation (8.3), the basic result being equation (8.15):

$$R_{N_i} = \frac{2kT}{qI_b} \tag{8.15}$$

A convenient version of this equation, for T = 293 K or 20°C, is

$$R_{N_i} = \frac{50.5}{I_b} \tag{8.16}$$

where R_{N_i} is in kilohms and I_b is in microamperes.

The NE5534A provides an illustrative example of the application of equation (8.16). The data sheets give a bias current value of 0.5 μA, and substituting this in equation (8.16) gives R_{N_i} = 101 kΩ, which agrees well with the high-frequency end of the R_{N_i} curve in Figure 8.9, where the effect of flicker noise is negligible. This curve was derived from the data sheet curve for input noise current density.

A warning is called for in relation to equations (8.15) and (8.16), because some op. amps. employing bipolar input transistors achieve low bias current values, not by operating these transistors at very low base current, but by means of additional bias-current-neutralizing circuitry. This scheme, however, does not reduce the base-current noise – indeed it adds further uncorrelated noise current – so that the value of R_{N_i} for such op. amps. is very much less than is given by the above equations. Op. amps. of this type are not ideally suited to microphone amplifier applications.

8.3.10 'Low-noise transistors'

The term 'low-noise transistor' is used in the literature with two distinct meanings.

According to the first meaning, a low-noise transistor is one whose voltage noise is low, i.e. which has a low value of R_{N_v}, implying that $r_{bb'}$ is also low. Such a transistor might be used without an input transformer to amplify the output of a 30 Ω microphone, for example, and yet give an acceptable noise figure. Under such conditions the source resistance is usually much less than $R_{S_{opt}}$, so that only R_{N_v} rather than R_{N_i} is of real significance.

The requirement for low $r_{bb'}$ can be met by choosing a fairly large-area type of transistor, as used in the driver stages of audio power amplifiers. For example, types BFX85 (NPN) and BC461 (PNP) have an effective noise $r_{bb'}$ value in the

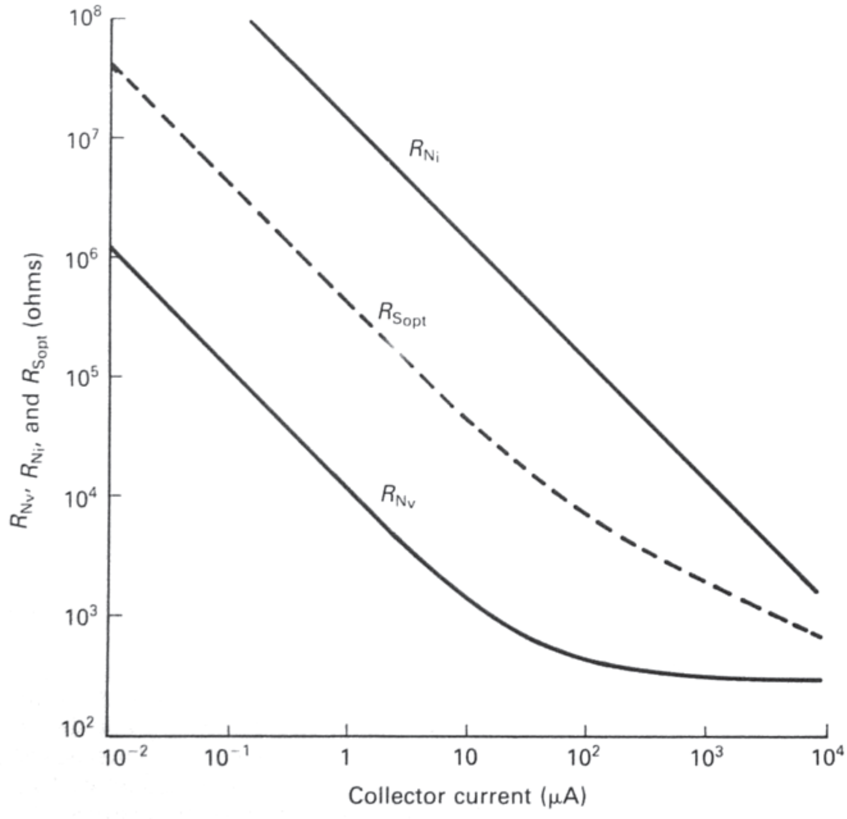


Figure 8.11 *Theoretical R_{N_v}, R_{N_i} and $R_{S_{opt}}$ curves for a transistor assumed to have $r_{bb'}$ = 300 Ω and a constant β_{dc} (or h_{FE}) of 300.*

region of 15 Ω, compared with maybe 200 Ω for small transistors such as the BC109.

An alternative is to connect n small transistors in parallel, which divides the effective values of $r_{bb'}$, R_{N_v} and R_{N_i} by n if the transistors are identical and each is operated at the same current as the original single transistor. Some transistors, such as the National Semiconductors LM394N supermatched pair, exploit integrated-circuit technology and have a large number of paralleled transistors built in – 50 per section in the example just quoted.

For any single transistor, flicker noise causes the value of R_{N_i} to fall off below a certain corner frequency – see Figure 8.9 – and increasing the working current of such a transistor in order to reduce R_{N_v} always raises this corner frequency, thus degrading the overall noise performance for $R_s = R_{s_{opt}}$. However, if the total working current is increased by adding more transistors in parallel, each passing the same current as the original single transistor, then R_{N_v} and R_{N_i} are reduced without adversely affecting the flicker noise corner frequency.

Flicker noise is not normally much of a problem in the design of microphone amplifiers, largely because the ear has reduced sensitivity at the low frequencies involved, and because the flicker noise is usually well below the studio background noise, whose spectrum also rises at low frequencies. However, liberties should not be taken and it is wise to avoid unnecessarily high current densities.

According to the second meaning of the term a 'low-noise transistor' is one which, when operated at an appropriate current and fed from a source of optimum resistance, gives an exceptionally low noise figure. It is evident from equation (8.11) that the criterion in this case is the ratio R_{N_i}/R_{N_v}, which should be as high as possible. This is equivalent to the requirement that the product (nV/$\sqrt{\text{Hz}}$) × (pA/$\sqrt{\text{Hz}}$) should be as small as possible. Table 8.1 gives some typical values for a selection of transistors and op. amps. R_{N_i} often varies much more

Table 8.1 *Parameters for a range of transistors and op. amps.*

Type	*Pol*	*Case*	I_{dc} (mA)	R_{N_v} (Ω)	R_{N_i} (kΩ)	$2\beta_{dc}/g_m$ (kΩ)	R_{N_i}/R_{N_v}	$R_{S_{opt}}$ (Ω)	NF_{opt} (dB)
BC109	NPN	T018	0.1	300	100	150	330	5480	0.45
LM394	NPN	8-pin DIL	0.1	180	260	275	1440	6840	0.22
BFX37	PNP	T018	1.0	90	8.0	10	89	850	0.84
GET106*	PNP	–	0.3	190	3.8	8.0	20	850	1.6
2N4401	NPN	T092	1.0	31	5.5	8.0	177	413	0.61
2N4403	PNP	T092	1.0	28	8.0	11	286	473	0.49
BFX85	NPN	T039	2.5	25	1.5	2.2	60	194	1.0
BC143	PNP	T039	2.5	13	1.6	2.4	123	144	0.72
BC143	PNP	T039	1.0	21	3.2	6.0	152	259	0.65
NE5534	NPN	8-pin DIL		760	100		132	8720	0.70
LT1028	NPN	8-pin DIL		45	16		356	850	0.44

*Input transistor used in 1962 in BBC microphone amplifiers.

from sample to sample than R_{N_v} and is sometimes much lower than the ideal β_{dc}/g_m value because of the influence of $1/f$ noise. Noise below 1 kHz and above 10 kHz was attenuated in making the noise measurements.

From equations (8.12) and (8.13) it is evident that to give the desired high R_{N_i}/R_{N_v} ratio, β_{dc} should be as high as possible and well maintained down to low working currents. The base spreading resistance $r_{bb'}$ should be reasonably low, but at a typical current of 100 µA there is not much advantage in having a value of $r_{bb'}$ below, say, 50 Ω.

An important point to appreciate is that when the source impedance can be correctly optimized, by means of a transformer or otherwise, no improvement whatever in noise figure can be obtained by adding further transistors in parallel – a point of which the designer of at least one commercial microphone amplifier appears to have been unaware!

An optimized noise figure of 1 dB or better is quite readily obtainable without resort to costly input devices.

The description 'low noise' in device data sheets, in addition to the above meanings, usually implies that the devices have been subjected to production tests to eliminate specimens with high flicker or popcorn noise.

8.3.11 Noise criteria for moving-coil, ribbon and capacitor microphones

Modern moving-coil microphones usually have a nominal impedance within the range 150 to 300 Ω, with a typical open-circuit voltage sensitivity of about 0.13 mV/µbar or 1.3 mV/Pa. (1 Pa = 1 N m^{-2}, which corresponds to a sound-pressure level or SPL or 94 dB.)

The passive resistance of the coil constitutes the main component of the impedance over most of the audio spectrum, though this is augmented in mild degree by motional impedance and, at the very highest audio frequencies, by passive inductive reactance.

The motional impedance above about 100 Hz, for good omnidirectional moving-coil microphones, is fairly frequency-independent and resistive, and typically represents about 20% of the total impedance – see curve (a) in Figure 8.12.*

For cardioid moving-coil microphones, the motional component of the impedance varies much more with frequency, and the total impedance usually rises to a peak of at least twice the d.c. resistance at some frequency in the region of 100 Hz – see curve (b) in Figure 8.12.

Despite the above detailed effects, no error that is of much consequence in normal circumstances is introduced if it is assumed that the noise output of a moving-coil microphone, in truly silent surroundings, has a white spectrum with a magnitude corresponding to Johnson noise in the nominal resistive impedance, and the same applies to ribbon microphones – see curve (c) in Figure 8.12.

Assuming a resistive microphone impedance of 200 Ω and a noise bandwidth of 12.5 kHz (this relates to an A-weighting network with unity gain at 1 kHz and for

*A level acoustic frequency response requires that the diaphragm should be resistively controlled over most of the audio-frequency band, so that frequency-independent acoustic pressure gives frequency-independent velocity and hence output voltage. If, now, a frequency-independent current is fed to the coil, this will produce frequency-independent force, velocity, and hence motional voltage – in other words a frequency-independent resistive motional impedance. At the highest frequencies the effects of diffraction have to be allowed for in the design, and the use of a Helmholtz resonator technique to augment the bass output by applying pressure to the rear of the diaphragm also complicates matters.

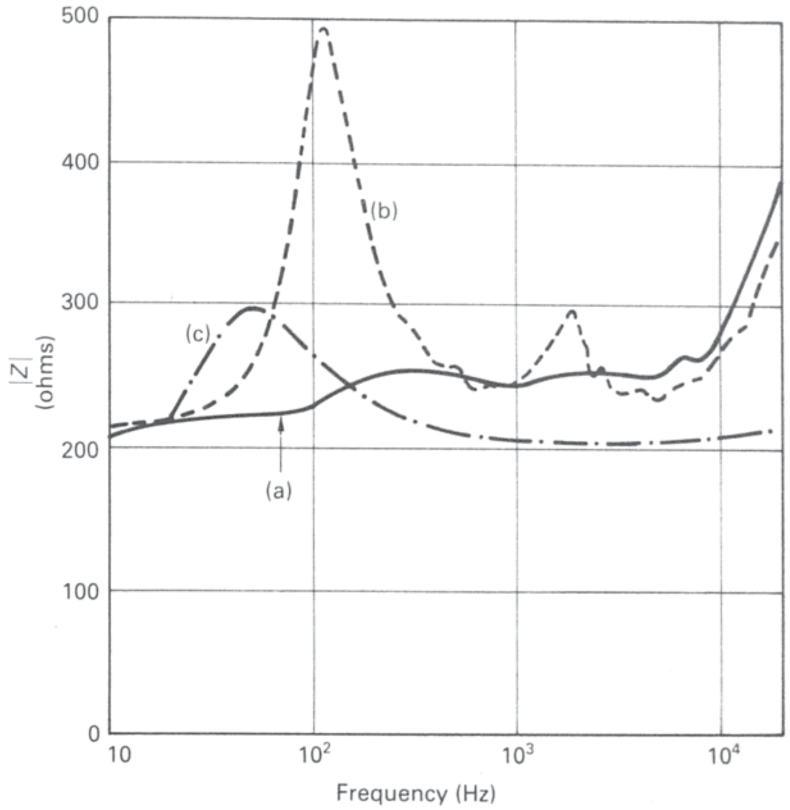


Figure 8.12 *Impedance characteristics for three types of microphone: (a) omnidirectional moving-coil; (b) cardioid moving-coil; (c) ribbon.*

frequencies up to 20 kHz*), then the A-weighted noise voltage at 293 K (20°C) is 0.20 µV r.m.s. With the above-mentioned typical microphone sensitivity of 0.13 mV/µbar, this noise level corresponds to an equivalent noise SPL of 17.7 dBA. Hence if used with an amplifier which achieves a 1 dB noise figure, the actual noise SPL obtained would be 18.7 dBA.

For capacitor microphones the much higher voltage sensitivity of 1 mV/µbar is about average, with a typical output impedance from the internal active circuitry of around 150 Ω. However, the output noise level in silent surroundings is much higher than that of Johnson noise in 150 Ω, and the noise spectrum, unlike that for moving-coil and ribbon microphones, is far from white. The noise output arises mainly from the following causes:

(1) The action on the rear surface of the diaphragm of a frequency-independent thermal agitation noise pressure originating within the viscous air-damping resistance between the diaphragm and the fixed electrode.
(2) Shot and flicker noise associated with the FET drain current, as represented by R_{N_v} in series with the gate lead.
(3) Gate current shot noise as represented by the current generator R_{N_i} between gate and source.

*The 12.5 kHz equivalent noise bandwidth figure is an accurately computed value supplied by John Vanderkooy.

(4) Noise current fed to the gate due to the Johnson noise voltages acting in series with the high-value leak resistors, if present.

A further small amount of electrical noise may in practice be produced by additional transistors or op. amps. and their associated resistors when these are incorporated.

Before considering the quantitative noise performance of practical capacitor microphones, it is perhaps worth pausing to contemplate the fact that the ultimate and irreducible noise level of an imaginary ideal capacitor microphone is determined by the thermal agitation noise pressure associated with the radiation resistance seen by the diaphragm. If this were the only source of noise, the noise performance would be very much better than that of any existing practical microphone and, moreover, it would be independent of the microphone diameter provided this was small compared with the wavelength at all frequencies.* The achievable signal-to-noise ratio with such an ideal microphone would be a characteristic of the acoustical field itself rather than the microphone.[24] The calculated A-weighted noise SPL for frequencies up to 20 kHz is then about –4 dBA.

The above result emphasizes the point that in microphone capsules, just as in low-level electrical circuits, the presence of resistance always degrades the noise performance, and that the amount of degradation present in even the best commercial microphone designs is quite large, since these seldom have a noise SPL as good even as +14 dBA.

In practical omnidirectional capacitor microphones, cause (1) in the above list results in a noise output component that has the same spectrum as the acoustical pressure response, which rolls off at high frequencies if the axial free-field response is flat. This source of noise is normally the dominant one at the higher audio frequencies in a good design, but is liable to be overridden at lower frequencies by noise from the other causes listed. However, with a good enough FET and a leak resistance value of some thousands of megohms – or no leak at all as in many electret-polarized microphones – the noise due to cause (1) may predominate down to frequencies not very much higher than 100 Hz.

Turning now to cardioid and figure-of-eight capacitor microphones, the actuating acoustic pressure difference between the two sides of the diaphragm now becomes progressively smaller as the frequency falls, so that to maintain a level frequency response the diaphragm stiffness has to be made very much less than in an omnidirectional microphone – by a factor well in excess of a hundred in practice. The damping resistance, which has about the same value as in omnidirectional versions, then exerts a dominant control over the diaphragm motion down to quite low audio frequencies, though there is usually an appreciable reduction in response at the extreme bass end.

Though the total signal force is much less at low frequencies than in an omnidirectional capacitor microphone, the noise force due to the viscous damping is about the same and therefore produces a much larger diaphragm amplitude, and corresponding noise output voltage, than in the omnidirectional version. The spectrum of the output noise due to cause (1) in the list now rises with falling frequency at 20 dB/decade, maintaining its dominance over the other causes listed at all frequencies in a good design.

*For practical microphones of small size, the equivalent noise SPL, if due predominantly to the internal viscous damping resistance, is reduced by 3 dB on doubling the diameter, assuming the Q-value is kept constant. The noise performance can also be improved by under-damping the diaphragm and correcting the resultant high-frequency peak by electrical equalization.

Hence it is evident that in directional capacitor microphones with the usual type of capsule design, a substantial rise in the noise spectrum with falling frequency is quite unavoidable no matter how good the associated circuit design may be.

The inherently lower low-frequency noise given by omnidirectional capacitor microphones may be reduced even further by employing RF circuitry of enlightened design in place of the ordinary FET scheme.

It is interesting to notice that whereas a rising low-frequency noise spectrum is inherent in practical cardioid and figure-of-eight microphones of the capacitor type, because their diaphragms are resistively controlled, this is not an inherent feature with ribbon microphones, which are basically mass-controlled.

To put the above matters in their proper perspective, it should be added that although obtaining extremely low noise levels at low frequencies is undoubtedly a matter of some academic interest, it is nevertheless of no real practical concern in an ordinary music context, for even the relatively high low-frequency noise level of a good commercial cardioid or figure-of-eight capacitor microphone is normally quite swamped by low-frequency studio background noise, and may be below the threshold of hearing at normal listening levels.

The noise performance requirements for a microphone amplifier are much less exacting if it is to be used only with capacitor microphones than if it is also required to give first-class results, with an input transformer if necessary, when moving-coil or ribbon microphones are employed. That this is so becomes evident from the following example.

As already mentioned, a very good present-day capacitor microphone may possibly have a noise level, expressed as an equivalent SPL, as low as 14 dBA. At a sensitivity of 1 mV/μbar this corresponds to a noise output voltage of 1 μV r.m.s., which is the Johnson noise voltage, in the A-weighting bandwidth of 12.5 kHz, for a resistance value of 5.2 kΩ.

In the above situation, amplifier noise will worsen the overall noise performance by 3 dB if the amplifier R_{N_v} value is 5.2 kΩ, and by 1 dB if R_{N_v} is 1.35 kΩ. The R_{N_i} noise, for any reasonable value of R_{N_i}, will certainly be fairly negligible if the microphone output impedance has the typical value of 150 Ω mentioned previously.

For a good noise performance with a 200 Ω moving-coil microphone used without an input transformer, R_{N_v} would have to be much less than 200 Ω and R_{N_i} would have to be much greater than 200 Ω. For a 1 dB noise figure, possible values would be $R_{N_v} = 26\ \Omega$ and $R_{N_i} = 1.5\ \text{k}\Omega$. If, as might well turn out to be the case, R_{N_v} was not as low as this, then a suitable input transformer would be required for achieving the desired low noise figure – but with or without the transformer, the above 58:1 ratio of R_{N_i} to R_{N_v} is a minimum requirement for achieving a 1 dB noise figure. A very much smaller ratio would suffice with the capacitor microphone.

Thus with capacitor microphones only the voltage noise of the amplifier is likely to be of any real significance, whereas with other types of microphone both voltage and current noise must be taken into account if the best possible performance is to be obtained.

Though microphone amplifier voltage noise can be quoted in nV/√Hz, or as an equivalent R_{N_v} value, a widely adopted practice is to express it in dBu, i.e. in decibels relative to 0.775 V r.m.s.* for some stated or assumed bandwidth such as 20 kHz.

*This voltage across 600 Ω gives 1 mW. The 600 Ω figure is a hangover from the early days of telephony, being the characteristic impedance of a typical open-wire telephone line as carried on old-style telephone poles. In America dBv has the same meaning as the European dBu. Another unit is the dBV, or dB(V), which means decibels relative to 1 V r.m.s.

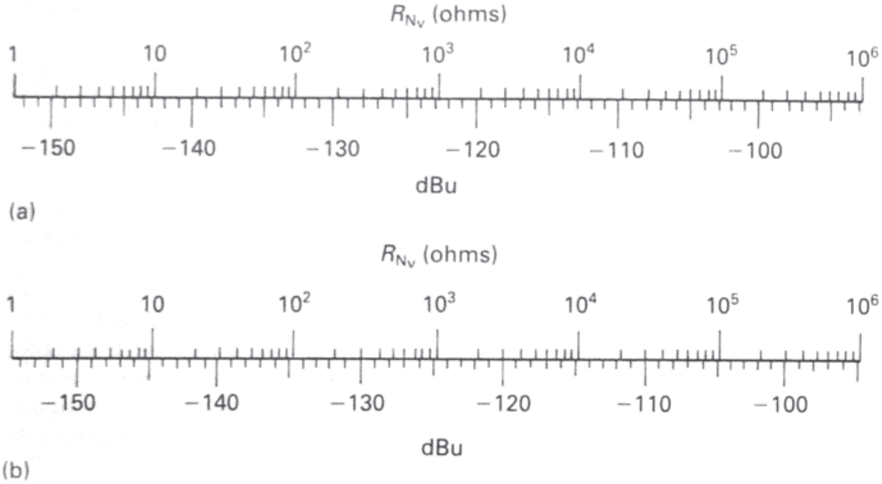


Figure 8.13 *Conversion scales (20°C): (a) 20 kHz, (b) 12.5 kHz noise bandwidth.*

Figure 8.13 provides a convenient means for converting R_{N_v} values to dBu or vice versa, and makes it evident, in relation to the above capacitor microphone example, that if the amplifier voltage noise is –125 dBu or less, it is unlikely to degrade the noise performance of any capacitor microphone by more than 1 dB. Most commercial mixer desks achieve a noise performance at least as good as this.

Though dBus are here being used to express just the voltage noise of the amplifier itself, they are also frequently used to express the total noise in a microphone input circuit with a passive microphone resistive impedance present – this usage is fully explained in Section 8.3.13.

8.3.12 Input impedance, negative feedback and noise

It is important to appreciate the following two points.

Firstly, the application of overall negative feedback to an amplifier does not in itself affect the noise figure at any specific frequency, though the resistors introduced for applying the feedback will generate some extra Johnson noise, which is kept to a minimum in good designs.

Secondly, though the input impedance of an amplifier is often greatly influenced by the application of negative feedback – feedback applied in shunt with the input lowering it and feedback applied in series raising it – nevertheless the value of source resistance, $R_{S_{opt}}$, for optimum noise performance is not affected by the feedback as such.

8.3.13 The 'equivalent input noise' (EIN) concept

As already expounded, the noise performance of a microphone amplifier may be specified very satisfactorily by quoting its R_{N_v} and R_{N_i} values, or the corresponding nV/√Hz and pA/√Hz figures.

Given such information, equation (8.9) then enables the noise figure for any passive resistive source impedance to be easily calculated, and equation (8.10) gives the source impedance required for optimum noise performance. This optimum performance is then as indicated by equation (8.11).

For complete information the frequency dependence of R_{N_v} and R_{N_i} should ideally be included, though the variation in R_{N_v} is usually negligible. In practice, however, single broad-band or spot-frequency values are normally quite adequate.

An alternative and widely used method for presenting the noise performance of a microphone amplifier, or amplifying device, is based on the notion that with some specific value of passive source resistance R_S connected to the input, the total noise output voltage magnitude may be accounted for in terms of a single equivalent input noise (EIN) voltage of white spectrum acting in series with the source.

A change in the value of R_S alters the magnitude of the EIN voltage, both because of the different Johnson noise voltage generated by R_S and also because the value of R_S affects the magnitude of the contribution from the amplifier's current-noise generator – see Figure 8.7. However, the total output noise may still be regarded as being due solely to a single EIN voltage, now of modified value, acting in series with R_S.

The EIN voltage itself may be expressed in several different ways.

It may be given in nV/√Hz, or as a noise voltage in a stated bandwidth. Figure 8.14(a) and (b), based on the data sheets for the NE5534 op. amp., embody these two methods, the descriptions above them being those used by the manufacturer.

Figure 8.14(c) is taken from the data sheets for the Trans-Amp balanced-input microphone amplifier module made by Valley People Inc., the EIN voltage here being expressed in dBu, though labelled 'dBv (dB re 0.775 V)' on the American data sheet.

It is evident from the definition of EIN voltage that if the amplifier itself was noiseless, then the EIN voltage would be just the Johnson noise voltage in the source resistance R_S, and it is helpful to include graphs, shown in broken-line, depicting this ideal performance. The vertical spacing between the device curve and the corresponding broken-line graph, at any given value of R_S, then gives the noise figure for that R_S value.* The excellent noise performance of the Trans-Amp when fed directly from a 200 Ω moving-coil microphone will be noticed, this unit having a considerable number of transistors in parallel in each half of its input stage.

8.4 Electronically balanced input circuits and noise

Though microphone amplifier circuits with transformerless balanced inputs differ considerably in design details, the circuit shown in Figure 8.15(a) is representative of good contemporary practice.[25] The value of R_G is varied to provide gain adjustment, sometimes in 10 dB steps. Additional components, here omitted for clarity, are often included to reduce the sensitivity to r.f. interference and to control the rate of attenuation of loop gain so as to give good stability margins.

The resistors labelled R' serve to hold the input circuit at the right d.c. level, but are normally made of sufficiently high value to avoid significant shunting of the signal source.

The negative feedback from each op. amp. output, via R_{fb_1} or R_{fb_2} to the emitter of the associated transistor, ensures that the two collector voltages are at all times

*For (a) and (b) in Figure 8.14, NF = 20 log (noise voltage ratio), whereas in (c) the spacing may be related directly to the decibel scaling.

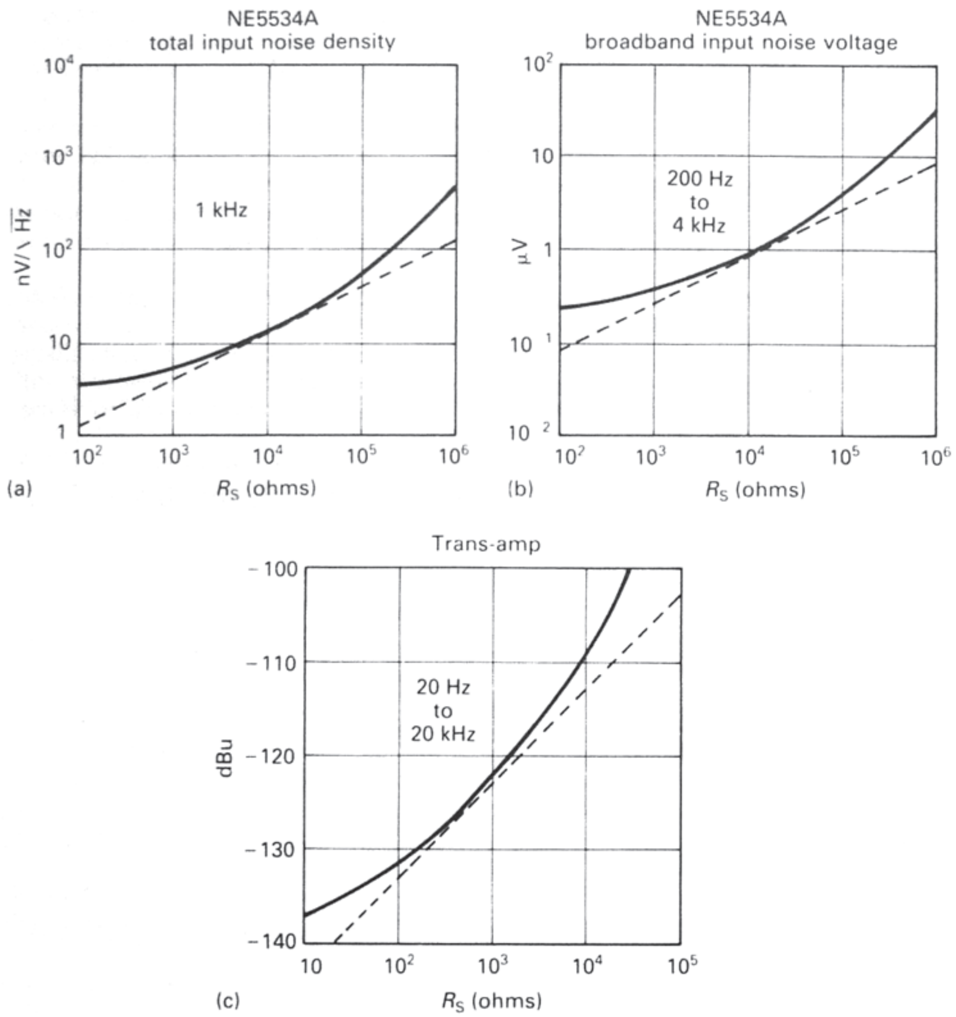


Figure 8.14 *Equivalent Input Noise (EIN) voltage data (20°C). The broken-line graphs represent the source Johnson noise alone.*

held very close to the d.c. bias voltage level $+V_B$, thus keeping the instantaneous collector currents very nearly constant.

When a common-mode input voltage V_{com} is applied to input terminals 1 and 4, voltage changes virtually equal to V_{com} are caused to appear also on points 3 and 6, thus keeping constant the voltages across R_{fb_1} and R_{fb_2} and hence the transistor currents.

The equal voltages on 3 and 6 are applied as a common-mode input to the balun stage involving op. amp. 3, giving zero output from it if the resistor values are accurately matched. Thus the complete Figure 8.15(a) circuit has ideally zero response to common-mode inputs. No voltage appears across R_G for such inputs.

When a balanced input voltage V_{bal} is applied between terminals 1 and 4, the feedback again holds the transistor currents virtually constant, and to achieve this it is now necessary for the feedback circuit to produce a voltage change between the emitters equal to V_{bal}. Hence

$$V_{3\text{-to-}6} \times \frac{R_G}{2R_{\text{fb}} + R_G} = V_{\text{bal}} \tag{8.17}$$

where $R_{\text{fb}_1} = R_{\text{fb}_2} = R_{\text{fb}}$.

The voltage between points 3 and 6 is subjected to an inverting gain of unity in the balun stage, the final output voltage therefore being given by

$$V_{\text{out}} = -V_{\text{bal}} \times \frac{2R_{\text{fb}} + R_G}{R_G} \tag{8.18}$$

The circuitry within each broken-line enclosure in Figure 8.15(a) may be regarded as constituting an op. amp. with modified characteristics, terminal 1 of the left-hand circuit being the non-inverting input, terminal 2 the inverting input and terminal 3 the output. Representing these modified op. amps. by simple broken-line triangles enables the Figure 8.15(a) circuit to be drawn as at (b).

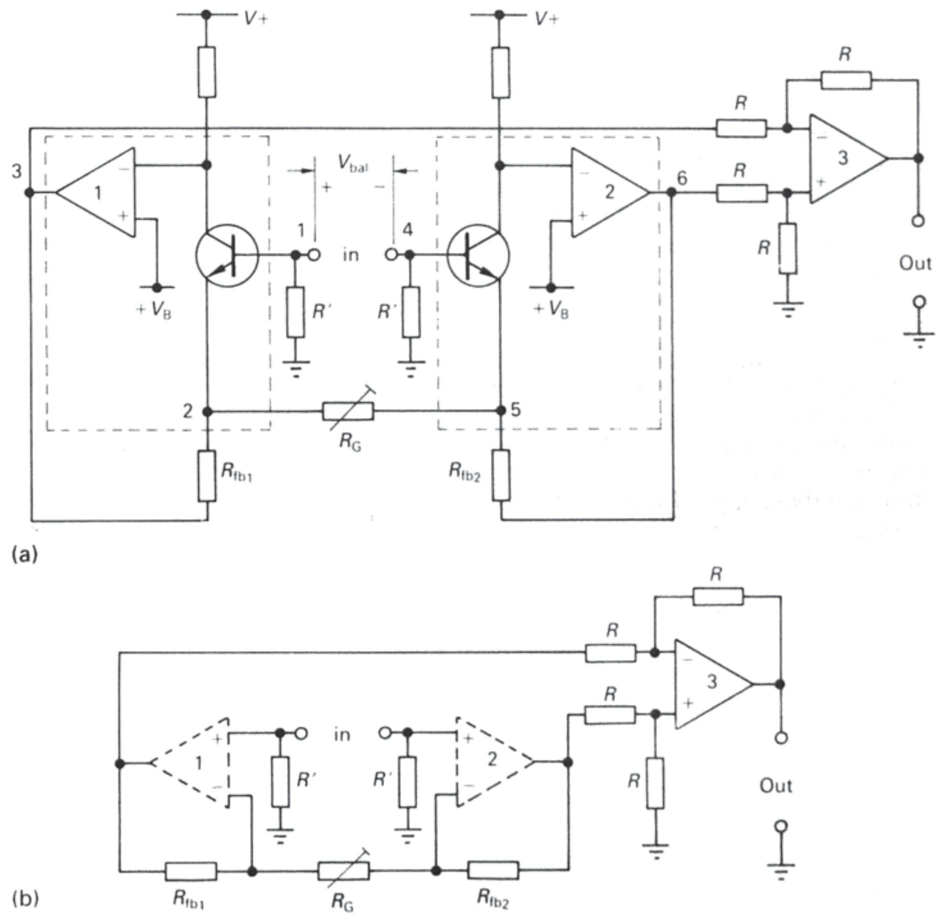


Figure 8.15 *(a) Basic circuit of a modern transformerless microphone amplifier with balanced input; (b) simplified version of (a).*

A satisfactory practical microphone amplifier, suitable for feeding straight from 200 Ω moving-coil and ribbon microphones, could indeed be directly based on Figure 8.15(b), op. amps. such as the LT1028 (see Table 1), which have low voltage-noise, being used in positions 1 and 2.

In recent times, however, a single op. amp., type SSM2015P, intended specifically as a balanced-input microphone amplifier, has become available and has internal circuitry in broad conformity with Figure 8.15(a). This product provides a neat, economical and generally highly satisfactory solution to the microphone amplifier problem, tending, indeed, to render most other solutions somewhat obsolescent.

The noise aspect of electronically balanced microphone amplifiers is interesting and demands careful consideration.

Referring again to Figure 8.15(a), it will be assumed that the input transistors provide sufficient gain to make it reasonable to ignore the noise produced by the op. amps. Thus the significant noise sources are the transistor voltage and current noise generators, Johnson noise in the feedback network, and source Johnson noise.

The high-value resistors R' exert a minor shunting effect across the source, which only very slightly degrades the noise performance, and they also produce a common-mode noise input, to which, however, the system is non-responsive. Hence the influence of these resistors on the noise output can be neglected in practice.

Figure 8.16 represents the essential elements of Figure 8.15 from a noise point of view.

Considering voltage noise first, it is evident that there will be introduced in series with the input circuit, in a balanced manner, a total Johnson noise voltage corresponding to a resistance value of $R_{N_{v1}} + R_{N_{v2}} + R_S + R'_G$, where R'_G is the parallel value of R_G and $(R_{fb_1} + R_{fb_2})$.

R_{fb_1} and R_{fb_2} are made as low in value as is practicable, consistently with the ability of the op. amps. to drive them to a sufficient output level when R_G is set to a low value. Thus the Johnson noise contribution from the feedback network is kept minimal.

The total Johnson noise voltage from the above-mentioned four resistances in series is subjected to a gain of $(R_{fb_1} + R_{fb_2} + R_G)/R_G$.

The noise current generators I_{N_1} and I_{N_2} together produce both a common-mode component of current noise, to which the complete circuit is non-responsive, and also a balanced component flowing round the input circuit. This latter component produces voltages across R_S and R'_G which give rise to a noise

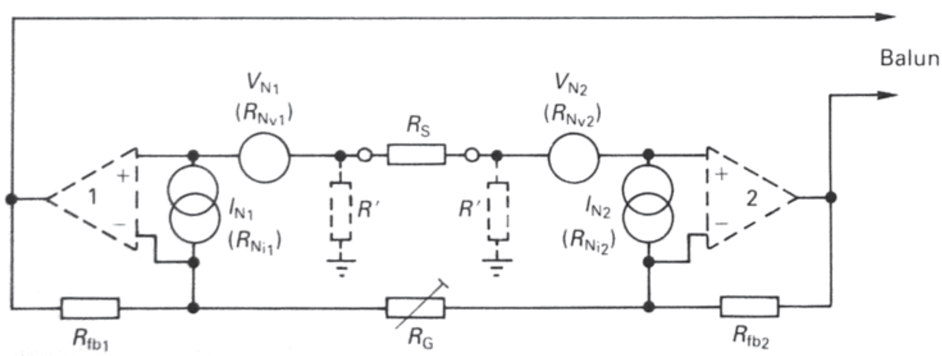


Figure 8.16 *Diagram for explanation of noise performance of Figure 8.15*

output. It is therefore necessary to consider the derivation of these separate noise current components.

Let the instantaneous values of I_{N_1} and I_{N_2} be i_{n_1} and i_{n_2} respectively, both being taken as positive when flowing upwards, say. Whatever the values of i_{n_1} and i_{n_2} may be, another pair of current components i_a and i_b can always be chosen so that, at the time instant concerned, the following relationships are satisfied:

$$i_{n_1} = i_a + i_b \tag{8.19}$$

$$i_{n_2} = i_a - i_b \tag{8.20}$$

These are two equations with two unknowns, so that given the i_{n_1} and i_{n_2} values at any instant, i_a and i_b can be determined.

It is evident that i_a constitutes the common-mode current component, flowing in the same direction in both current generators, whereas i_b constitutes the balanced component which flows round the circuit through R_S and R'_G.

Adding and subtracting equations (8.19) and (8.20) gives

$$i_a = \frac{i_{n_1} + i_{n_2}}{2} \tag{8.21}$$

$$i_b = \frac{i_{n_1} - i_{n_2}}{2} \tag{8.22}$$

Now the noise waveforms of which i_{n_1} and i_{n_2} in equation (8.22) are instantaneous values each have an r.m.s. value of I_N, it being assumed that $I_{N_1} = I_{N_2} = I_N$ in Figure 8.16. However, since I_{N_1} and I_{N_2} are uncorrelated, the minus sign in equation (8.22) is now of no consequence and the r.m.s. value of the right-hand side of the equation is $\sqrt{2}I_N/2$, i.e. the r.m.s. value of the balanced noise current flowing round the circuit due to the noise current generators is $I_N/\sqrt{2}$. The effective value of equivalent current-noise resistance for the complete circuit is therefore $2R_{N_i}$, where R_{N_i} is the value for a single transistor.

From the above, assuming equal R_{N_v} and R_{N_i} values for the two transistors and ignoring the small noise effect of the R' resistors, the noise performance of the Figure 8.16 balanced circuit is the same as that of the unbalanced circuit shown in Figure 8.17(a). Comparison with Figure 8.7 and reference to equations (8.9), (8.10) and (8.11) shows that, for the Figure 8.17(a) circuit

$$\mathrm{NF} = 10 \log_{10}\left[1 + \frac{2R_{N_v} + R'_G}{R_S} + \frac{R_S}{2R_{N_i}}\right] \tag{8.23}$$

$$R_{S_{opt}} = \sqrt{[(2R_{N_v} + R'_G)(2R_{N_i})]} \tag{8.24}$$

$$\mathrm{NF}_{opt} = 10 \log_{10}\left[1 + 2\sqrt{\frac{2R_{N_v} + R'_G}{2R_{N_i}}}\right] \tag{8.25}$$

When, as is often the case, the feedback network resistance R'_G is negligibly small, we then have

$$\mathrm{NF} \approx 10 \log_{10}\left[1 + \frac{2R_{N_v}}{R_S} + \frac{R_S}{2R_{N_i}}\right] \tag{8.26}$$

$$R_{S_{opt}} \approx 2\sqrt{(R_{N_v}R_{N_i})} \tag{8.27}$$

$$NF_{opt} \approx 10 \log_{10} \left[1 + 2 \sqrt{\frac{R_{N_v}}{R_{N_i}}} \right] \tag{8.28}$$

It is seen from equations (8.27) and (8.28) that though $R_{S_{opt}}$ is twice as high as it would be for a single transistor of the same type, nevertheless use of this higher R_S value gives the same noise figure as for the single transistor.

It is of interest to determine what happens to the noise performance of the Figure 8.15 type of circuit if, instead of operating with a balanced signal source, one end of the source is earthed to give unbalanced operation.

For simplicity it will again be assumed that R'_G is negligibly small. In other words it is being assumed that the currents in the bottom leads of the two current generators in Figure 8.16 flowing into the feedback network produce negligible noise output, the significant effect in normal balanced operation being produced by the balanced component of current in their top leads flowing through R_S. However, when the left-hand end, say, of R_S is earthed, I_{N_1} is virtually prevented from contributing anything at all to the noise output, the observed current-generator-originated noise output now being due almost entirely to the voltage across R_S produced by the full value of I_{N_2} flowing through it.

Voltage noise from $R_{N_{v1}}$, R_S and $R_{N_{v2}}$ is of course still applied between the non-inverting inputs as before. The noise performance of the system is therefore approximately as for the circuit of Figure 8.17(b), and we now have

$$\mathrm{NF} \approx 10 \log_{10} \left[1 + \frac{2R_{N_v}}{R_S} + \frac{R_S}{R_{N_i}} \right] \tag{8.29}$$

$$R_{S_{opt}} \approx \sqrt{(2R_{N_v} R_{N_i})} \tag{8.30}$$

$$NF_{opt} \approx 10 \log_{10} \left[1 + 2 \sqrt{\frac{2R_{N_v}}{R_{N_i}}} \right] \tag{8.31}$$

Comparisons of equations (8.29), (8.30) and (8.31) with (8.26), (8.27) and (8.28) shows that earthing one end of R_S to give unbalanced operation has the following effects:

(1) If the value of R_S remains unchanged, the noise figure is worsened.
(2) The value of R_S required for optimum noise performance is reduced by a factor of $\sqrt{2}$.
(3) Even with R_S reduced as in (2), NF_{opt} is still not as good as with balanced operation.

These effects are borne out experimentally, but if the value of R_S used is well below $R_{S_{opt}}$, as is frequently the case in practice, the differences become quite negligible since voltage noise then tends to become dominant and is the same for balanced and unbalanced operation.

An interesting sidelight on the above relates to the ordinary use of op. amps., for example in the virtual-earth manner shown in Figure 8.18(a). The long-tailed-pair input stage of the op. amp. is here being operated in the unbalanced manner referred to above, and the noise performance obtainable is commensurate with the device's R_{N_v} and R_{N_i} (or nV/$\sqrt{}$Hz and pA/$\sqrt{}$Hz) values as normally quoted. However, by changing to the (b) scheme, op. amp. 1 input stage is now operated under proper balanced conditions and a noise performance somewhat better than given by the data sheet figures is obtained.

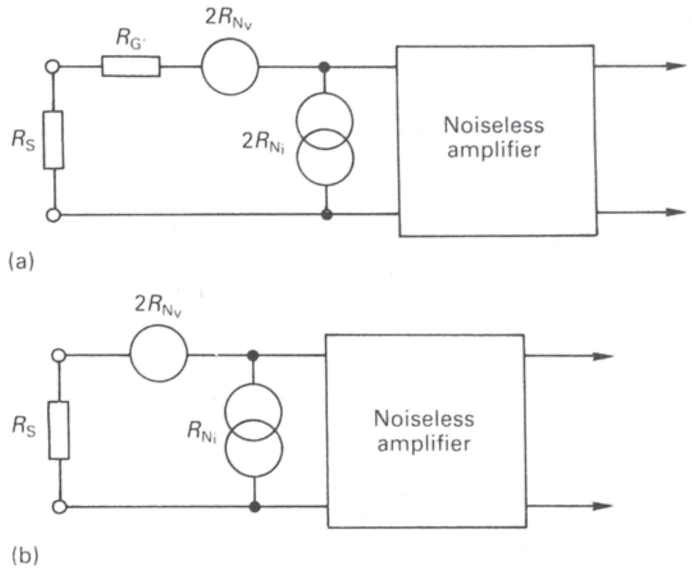


Figure 8.17 *(a) Circuit having the same noise performance as the circuit of Fig 8.16; (b) as for (a) but with one end of R_S in Figure 8.16 earthed. $R_{G'}$ in (a) is the small parallel value of R_G and (R_{fb1} + R_{fb2}), and was neglected in deriving (b).*

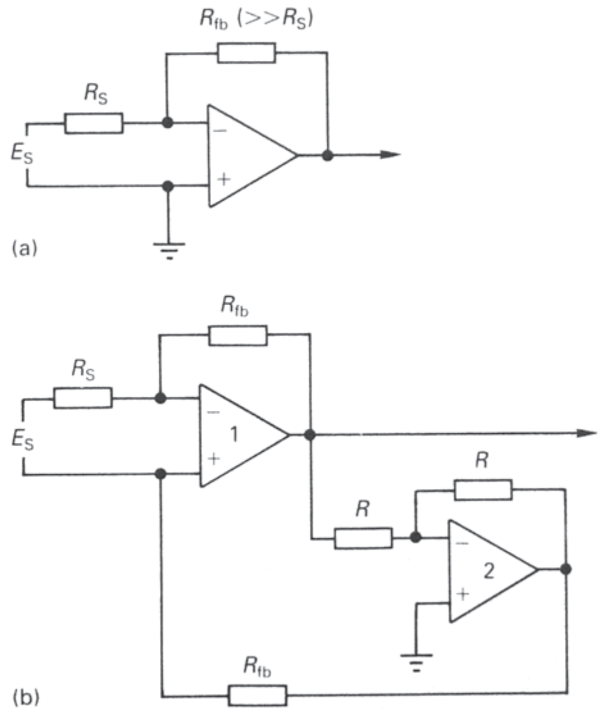


Figure 8.18 *Diagrams illustrating a point about op. amp. noise characteristics.*

8.5 Transformers

8.5.1 Introduction

Very few present day professional audio engineers have any detailed understanding of the theoretical principles and practical design problems relating to audio transformers, and this often creates difficulties in the liaison between users and manufacturers.

While the properties of a textbook 'ideal transformer' are generally well appreciated, difficulties arise in trying to understand the numerous ways in which practical transformers depart from this simple ideal.

When looked into in detail, the behaviour of an audio transformer is actually quite complex – and fascinatingly interesting – but the skilled designer knows, largely from experience, which complications are likely to be significant in a specific context and which can be ignored. Though the complications really constitute shortcomings, which no manufacturer can completely avoid, nevertheless in a well-designed transformer, used in the intended manner, none of them produce any audible quality degradation, even under the most searching conditions of subjective assessment. A good transformer is a virtually immaculate device.

Microphone transformers for use with transistor amplifiers do not normally have to step the source impedance up to nearly such a high value, for good noise performance, as was necessary in the valve era, and this has greatly eased the design problem with regard to the effects of leakage inductance, secondary shunt capacitance and inter-winding capacitance.

Indeed, the conditions under which a modern microphone transformer is required to operate are sometimes such that a first-rate performance can be achieved using only the very simplest design procedure, and a design falling into this category will first be considered.

8.5.2 A simple microphone transformer design

Suppose that it is desired to use a pair of 30 Ω 4038 A ribbon microphones – the commercial equivalent of the BBC 'PGS' design[12,32] – to feed via balanced lines to the unbalanced inputs of a Sony PCM-F1 digital recorder system.

A simple noise test may first be made by connecting resistors of various values to one of the microphone inputs of the PCM-F1 and plotting, on the vertical axis, the square of the amplified noise output voltage against the resistor values horizontally. (The lower audio frequencies, say below 500 Hz, should be attenuated by a simple output C–R filter to subdue possible hum and/or flicker-noise contributions. Alternatively an A-weighting or CCIR filter may be employed.)

The graph obtained is found to be a good straight line below about 2 kΩ, with an intercept on the horizontal axis at about –400 Ω, so that the value of R_{N_v} for this microphone amplifier is approximately 400 Ω.

Above 2 kΩ, however, the graph starts to bend over towards the horizontal, Because internal 10 kΩ resistors have unfortunately been incorporated across the inputs, considerably worsening the noise performance potentially obtainable from these otherwise excellent microphone amplifiers, which have a junction-FET long-tailed-pair input stage in combination with a separate op. amp. The 10 kΩ resistors, R_{103} and R_{203} – incidentally of a type looking more like a ceramic capacitor – should preferably be changed to 330 kΩ, to give the amplifiers a respectably high R_{N_i} value. I have made such a modification to several PCM-F1 units. The modification is quite pointless, of course, if only capacitor microphones are to be used – see Section 8.3.11.

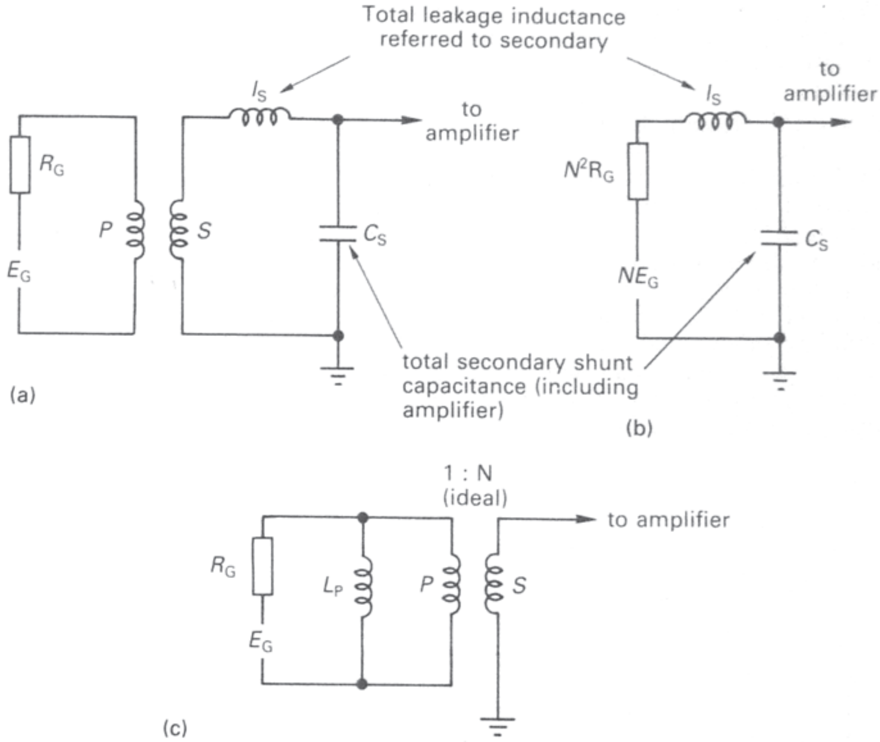


Figure 8.19 *Approximate equivalent circuits for microphone transformer, omitting copper and core losses; (a) and (b) apply at high frequencies, (c) at low frequencies. The ratio N of the ideal transformer is preferably made precisely equal to the actual turns ratio, though some authors depart slightly from this usage.*

If the input transformer has a ratio 1:N, then so far as medium- and high-frequency behaviour is concerned, the effective circuit is approximately as shown in Figure 8.19(a).* When the primary circuit impedance is low, as in the present instance, the effect of primary shunt capacitance may be completely ignored.

By transferring all quantities to the secondary side, the equivalent circuit of Figure 8.19(b) is arrived at. Now the number of turns required on the primary side is determined by low-frequency considerations, as discussed later, so that increasing the step-up ratio N can be achieved only by increasing the number of turns on the secondary winding. This increases the value of the leakage inductance l_S shown in the diagram, which, for a given winding geometry, is proportional to the square of the number of secondary turns.† The value of C_S, on the other hand, is almost unaffected by the number of turns.

*The source resistance and e.m.f. are here given the suffix G, for 'generator', since in a transformer context the suffix S, used for 'source' in the earlier part of this chapter, is now more conveniently employed to denote 'secondary'.

†The concept of leakage inductance is a convenient way to represent the fact that the magnetic coupling between the windings of a practical transformer is always slightly less than 100%, i.e. not quite all the flux that links with the turns of one winding also links with those of the other.

Thus, as the ratio N is increased, the resonance frequency of l_S and C_S falls, and would ultimately come right down into the audio band if N was made excessively high. But even if the resonance occurs at a frequency somewhat higher than the top of the audio band, it may still have an adverse effect on the levelness of response at high audio frequencies, to an extent dependent on the Q-value of the series-tuned circuit formed by the elements shown in Figure 8.19(b).

By careful design with regard to the values of l_S and C_S, and maybe with the addition of further damping in shunt with the secondary, N^2R_G may be given quite a high value without sacrificing uniformity of audio response up to about 20 kHz. Values of N^2R_G up to at least 100 kΩ can sometimes be achieved if C_S can be kept low enough. Such designs are considered in more detail later.

However, if N^2R_G is made not more than several kilohms, even very limited practical experience will make it obvious to the designer that the resonance will then occur at a frequency so far above the top of the audio band that its influence on the response at audio frequencies will probably be negligible. The above complications may then be ignored and the values of l_S and C_S allowed to look after themselves while other aspects of the design are attended to.

First, in the present design, the type and size of core to be employed must be decided upon, and Mumetal, Permalloy C or other commercially equivalent nickel–iron laminations, having about 78% nickel content, are almost universally used in such transformers, because of their very high value of initial permeability.[26] A lamination thickness of 0.38 mm or 0.015 in is normally adopted; thinner laminations would give the transformer a higher value of shunt eddy-current loss resistance (see Section 8.5.6), but the advantage of this in a normal audio transformer context is very small, so that the extra cost is seldom justified.

With regard to the influence of core size on performance, this is considered more fully in Section 8.5.12. Obviously the larger the core size selected, the higher will be the cost of the transformer and its associated screening can. The Type 187 size of lamination shown in Figure 8.20 is economical and enables a good performance to be obtained, especially if a stack thickness of twice the width of the middle limb is adopted, as for the present design.

The core dimensions l_m and A have the meanings shown in Figure 8.21, but in determining the effective cross-sectional area A_{eff}, it is usual to multiply the geometrical area shown by a stacking factor, often taken as 0.9, to allow for surface oxidation of the laminations – paint insulation is now much less used than was once the case. For the present '2 × square' design, the values taken are A_{eff} = 0.72 cm^2 and l_m = 5.1 cm.

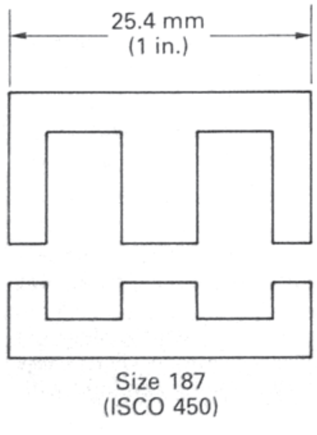


Figure 8.20 *A widely-adopted lamination size for microphone transformers. Other sizes in use are illustrated in Reference 1.*

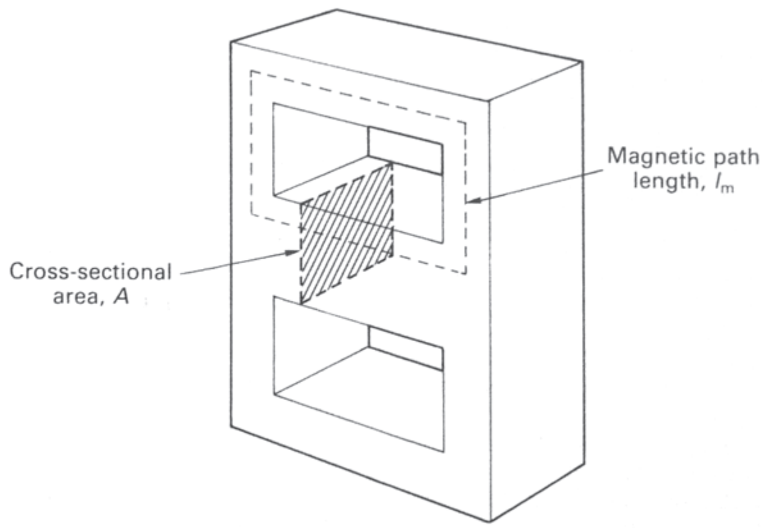


Figure 8.21 *Significance of the quantities A and l_m in a normal transformer core. In some sizes, e.g. Size 21 (ISCO 421), which is sometimes used for larger microphone transformers, the width of the centre limb is not quite double that of the other parts, requiring 'corrected' values of A and l_m.*

The equivalent circuit, at its simplest, for determining the low-frequency performance, is shown in Figure 8.19(c).

If R_G is 30 Ω and the response is to be not more than 1 dB down at 20 Hz, this corresponds fairly closely to –3 dB at 10 Hz, so that the reactance of L_P should be at least 30 Ω at 10 Hz, requiring L_P = 0.48 H or more.

Adopting the rationalized MKS, or SI (Système International) system of units, as is now the normal practice, the inductance is given by

$$L = \frac{4\pi n^2 A_{\text{eff}}\ \mu}{10^7\ l_m} \tag{8.32}$$

where n = number of turns
A_{eff} = effective core cross-sectional area (m^2)
l_m = magnetic path length (m)
μ = effective relative or specific permeability.*

To use equation (8.32) to determine the number of turns required, a value for μ must be assumed – that applicable at low frequencies and low-signal levels to a small transformer core of the normal type using interleaved laminations of 0.38 mm Mumetal. This is called the effective initial permeability, μ_i.

Several decades ago it was customary to assume a value of 7000 for the μ_i of an ordinary Mumetal core, but much progress has since been made in improving these nickel–iron alloys by the addition of small amounts of other constituents

*It is here convenient to use μ without a suffix, though in more fundamental work μ_s is often used for specific permeability. The quantity $4\pi/10^7$ in equation (8.32) is the permeability of free space, often denoted by μ_0. The product $\mu_0\mu_s$ is then called the absolute permeability and denoted by μ without a suffix. But since only specific or relative permeability is normally referred to in practical transformer design work, it is widespread engineering practice to use μ for this. Other suffixes of a more practical nature may then be added, e.g. μ_i for initial (specific) permeability.

and by improvements in the production technology. Effective values of μ_i well in excess of 20 000 are now regularly found.

In the present design μ_i = 15 000 was assumed. Then substituting the above mentioned values L_P = 0.48 H, A_{eff} = 0.72 cm^2 and l_m = 5.1 cm in equation (8.32) gives just over n = 134 turns. A winding of 136 turns of 32 s.w.g. (0.27 mm) self-fluxing enamelled wire was in fact adopted, and for experimental purposes this was put on as a centre-tapped bifilar winding requiring 68 bobbin revolutions – see Section 8.5.15.

The secondary winding, outside the primary, and similarly occupying the full bobbin width, was made 1150 turns of 42 s.w.g. (0.10 mm) enamelled wire, giving a ratio of 1:8.46 and thus ideally stepping up the nominal 30 Ω primary source resistance to 2.14 kΩ.

A copper-foil screen was placed between the two windings, being separated from each by a layer of approximately 0.1 mm insulation tape. Care was taken to avoid forming a shorted turn – a vital point. For comments on the effects of such a screen, reference should again be made to Section 8.5.15.

Another practical point is that when assembling a laminated Mumetal core, the laminations should be treated gently, rather as if they were made of glass. Stressing them beyond the elastic limit, so as to leave a permanent bend, inflicts serious magnetic damage and can cause a large reduction in permeability.

8.5.3 Simple tests on the above transformer

Resistance of value 31.6 Ω was shunted across the 600 Ω output of a Levell *R–C* oscillator, to provide a 30 Ω source for feeding the transformer primary. The secondary was fed to an oscilloscope on 10 mV/cm sensitivity, with the oscillator and oscilloscope earth terminals joined. The *outside* of the secondary was taken to the HI input of the oscilloscope, whose input capacitance is approximately 100 pF.

The frequency response was then determined in terms of oscillator attenuator readings for a constant output level of 10 mV peak-to-peak, and was found to be within the limits ±0.25 dB from 40 Hz to 20 kHz, being –1.0 dB at 20 Hz and –2.9 dB at 10 Hz. Thus all seemed well with regard to audio-frequency response.

The d.c. resistances of the windings were measured and found to be r_P = 1.9 Ω and r_s = 151 Ω.* Hence, assuming a 30 Ω resistive source and referring all quantities to the secondary side, gives the equivalent circuit of Figure 8.22 when the transformer is connected to the PCM-F1 amplifier. The noise figure is given by

$$NF = 10 \log_{10} \left[\frac{2150 + 687}{2150} \right] = 1.2\ \mathrm{dB}†$$

*If such a resistance measurement is made on a transformer after inserting the core, it can leave the core in a sufficiently polarized state to give some reduction in low-level inductance and a noticeable increase in microphony. Ideally it is then desirable to depolarize the transformer by applying a signal current, at say 30 Hz, of sufficient level to give core saturation, indicated by a highly distorted voltage waveform, and then to turn the level smoothly down to zero over a period of many seconds.

†This simplified calculation ignores the noise contribution (≈0.1 dB at significant frequencies) from the eddy-current loss and the PCM-F1 shunt input resistance.

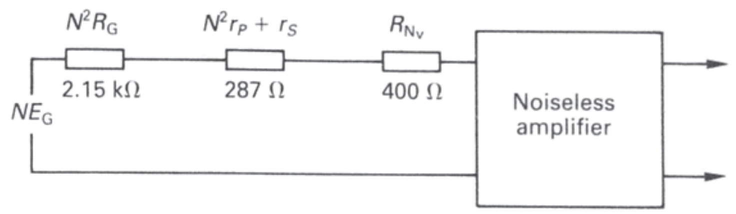


Figure 8.22 *Resistance values for simplified noise calculation of Section 8.5.3. The very small noise contributions from the shunt eddy-current loss, and the amplifier R_{N_i}, are here ignored.*

8.5.4 Tests of a more probing nature

Extending the above frequency-response measurement to higher frequencies gives curve *A* of Figure 8.23.

A much flatter response for frequencies well above the audio band may be obtained if a 10 kΩ resistor is shunted across the secondary, as for curve *B*, which was obtained with the same input level as for curve *A*. There is also a small improvement in the low-frequency response, the loss at 10 Hz being reduced from 2.9 dB to 2.2 dB.

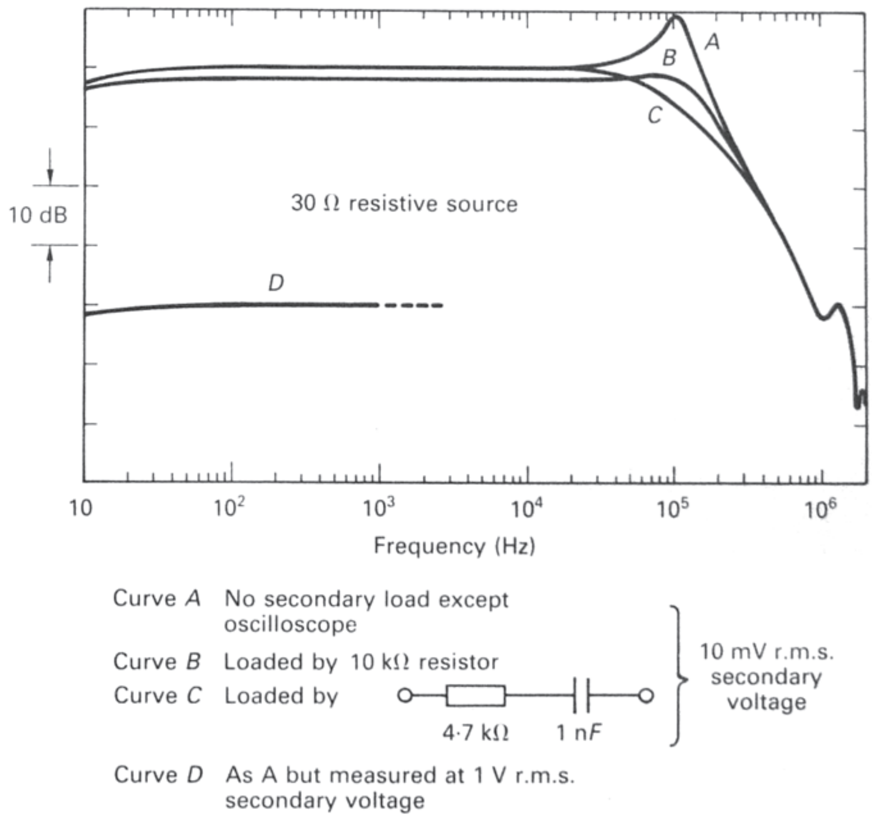


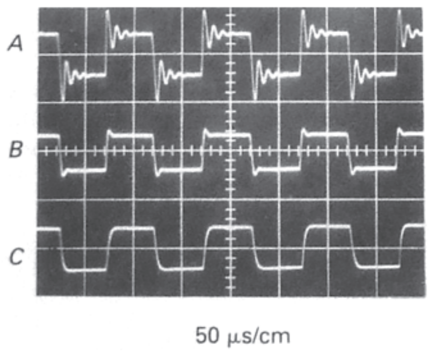
Figure 8.23 *Microphone transformer frequency responses with 30Ω resistive source.*

A disadvantage of such a shunt resistor, which appears between the 287 Ω/400 Ω junction and earth in Figure 8.22, is that it reduces the signal by more than it reduces the noise, increasing the calculated noise figure from 1.2 dB to 2.3 dB. However, this worsening of the noise performance can be avoided if the *equivalent* of a 10 kΩ resistor is obtained by shunt negative feedback via a resistor of much higher value – see Section 8.5.16.

An alternative technique for levelling the high-frequency response, which is widely used, is to shunt a series combination of *C* and *R* across the secondary, and curve C was obtained with 1 nF and 4.7 kΩ. This produces negligible worsening of the audio-frequency noise performance with the values here involved, since it is the parallel equivalent resistance value of this series combination that must be compared with the above 10 kΩ value from a noise viewpoint – the parallel resistance is 18 kΩ at 20 kHz, 59 kΩ at 10 kHz, and increases rapidly as the frequency falls further.

The response curves in Figure 8.23 conform closely to what would be expected from the simple equivalent circuits of Figure 8.19 – with damping additions in the case of curves *B* and *C* – for frequencies up to about 600 kHz, but above this frequency the rate of attenuation begins to depart from the simple theoretical 40 dB/decade, and above 1 MHz complex behaviour with multiple resonances sets in, as occurs with all transformers if the frequency is made high enough. These resonances involve a combination of leakage inductances between one part of a winding and another, and distributed capacitance. In higher-impedance transformers, complex behaviour is liable to start at a lower frequency.

Figure 8.24 shows 10 kHz square-wave responses corresponding to the three response curves of Figure 8.23. Though waveform *A* may not be thought to 'look nice', I do not believe that ringing at such a high frequency is ever of any direct subjective significance. Reports, which are not infrequent, that such effects can be perceived, are usually explicable in terms of the amazing power of the human imagination!



A No secondary load except oscilloscope

B Loaded by 10 kΩ resistor

C Loaded by

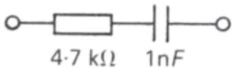


Figure 8.24 *Microphone transformer 10 kHz square-wave responses with 30Ω resistive source.*

However, the addition of secondary damping is sometimes beneficial in ensuring an adequate negative-feedback stability margin in the associated amplifier, and should be regarded as good general practice. See Section 8.5.16.

The response curves *A*, *B* and *C* of Figure 8.23, as already mentioned, were measured at a secondary voltage of only 10 mV peak-to-peak. If the measurement is repeated at a higher level, it is found that the shape of the response curve at low frequencies is altered, curve *D* having been obtained at a constant level of 1 V r.m.s. This happens because the mean slope of the *B–H* loop for the core material is amplitude dependent, as considered in more detail in Section 8.5.7.

The equation from which the peak flux density under sine-wave-voltage conditions may be obtained is

$$\hat{B} = \frac{E_{\text{rms}}}{4.44 A_{\text{eff}}\ nf} \tag{8.33}$$

where $\hat{B}$ = peak flux density (T)
E_{rms} = induced e.m.f. (V)
A_{eff} = effective core cross-sectional area (m^2)
n = number of turns
f = frequency (Hz)

For 1 V r.m.s. at 10 Hz, with $A_{\text{eff}} = 0.72\,\text{cm}^2$ or $0.72 \times 10^{-4}\,\text{m}^2$ and $n = 1150$, this equation gives $\hat{B} = 0.272$ T or 2720 G. Waveform distortion is visible below about 13 Hz with a 30 Ω source; 1 V r.m.s. secondary voltage corresponds to a sound pressure level, using a 4038A microphone, of approximately 140 dB.

So far, in accordance with widespread practice, the response curves presented have been obtained using a resistive signal source to simulate the microphone. Microphone impedances, however, as shown in Figure 8.12, are liable to depart fairly markedly, in both magnitude and phase angle, from the nominal resistive value, so that it is really more meaningful to make measurements by injecting the test voltage at very low impedance, preferably in a balanced manner, in series with actual microphones. Section 8.5.18 gives details of a 1000:1 transformer for conveniently carrying out such tests. It is suggested that regular use of such a device would now and then reveal frequency responses by no means as good as had been supposed!

Figure 8.25 shows response curves obtained as above for ribbon and moving-coil microphones, of nominally 30 Ω impedance, feeding the 1:8.46 microphone transformer.

The moving-coil microphone impedance at frequencies below 100 Hz is an almost pure resistance of about 22 Ω, giving an impressively level measured response. The ribbon microphone, on the other hand, rather similarly to that in Figure 8.12, has a broad impedance peak at about 50 Hz, whereas at the lowest frequencies its impedance becomes that of a fairly pure inductance because of the internal transformer; by 10 Hz a simple inductive potential divider effect has become established between this and the primary inductance of the 1:8.46 transformer.

Comparing Figures 8.23 and 8.25, it is seen that the high-frequency resonance comes at a lower frequency when measured with microphones than when measured with a simple resistance source, and this is because of the passive series inductance of these electromagnetic microphones.

Also shown in Figure 8.25 are low-level response curves obtained with a 100 Ω resistive source, under the same three secondary loading conditions as for the microphones; the source e.m.f. was the same for all three curves.

The 100Ω curves A and C show an interesting feature in that the response from about 200 Hz to 1 kHz shelves upwards by nearly 1 dB. This is because, with a

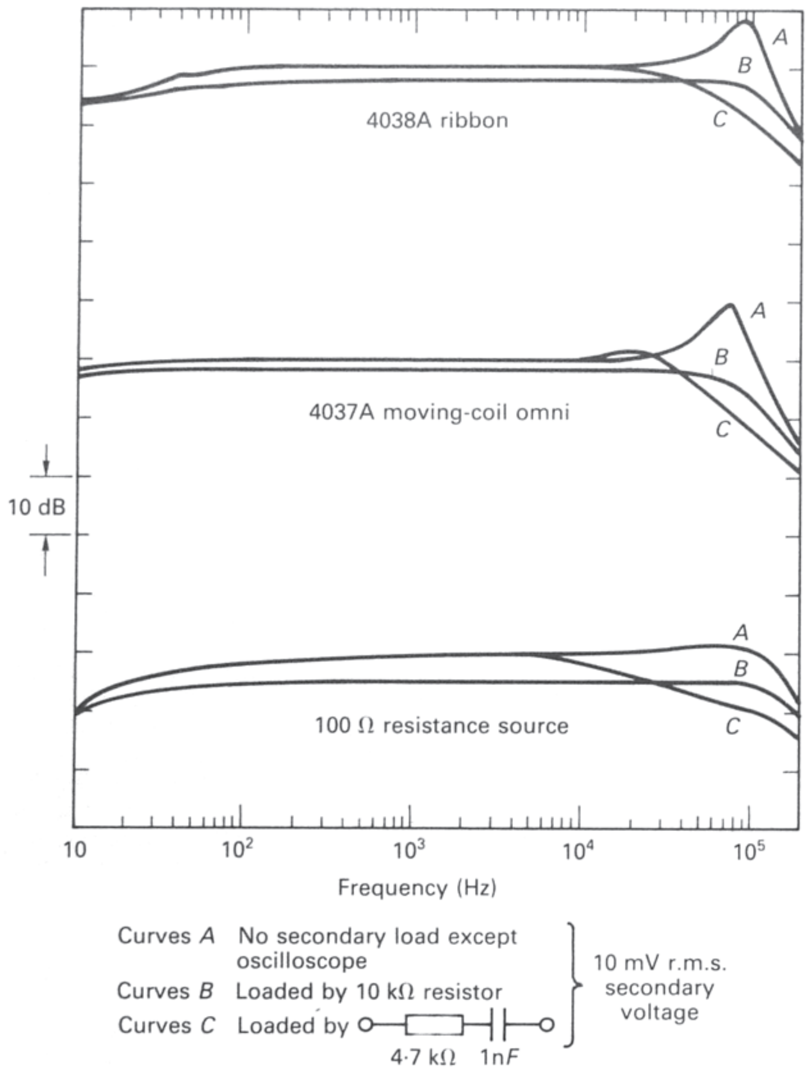


Figure 8.25 *Microphone transformer frequency responses with microphone and 100Ω resistive sources.*

high source resistance, significant attenuation is produced, at the lower frequencies only, by the shunt eddy-current loss resistance. As discussed in Section 8.5.6, this resistance starts to rise, at a rate proportional to the square root of frequency, above a corner frequency which is in the 200 Hz region for the type of core here used. Such a step is always evident with source resistances high enough to give several decibels of loss at 20 Hz, unless laminations thinner than the normal 0.38 mm are used.

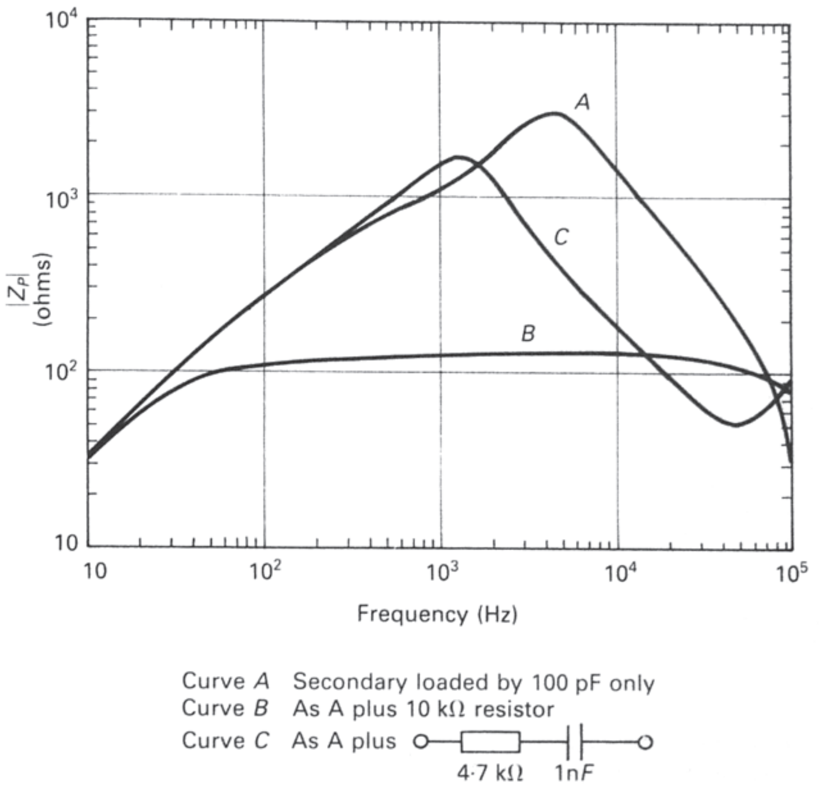


Figure 8.26 *Microphone transformer input impedance characteristics.*

Figure 8.26 shows the results of measuring the primary impedance of the 1:8.46 microphone transformer under the same three secondary loading conditions as before – a 100 pF capacitor was added across the secondary in place of the oscilloscope. The primary was fed with a current of 30 µA r.m.s. via a 100 kΩ resistor, the voltage across the winding being measured via a laboratory amplifier with a flat frequency response.

Curve *A* actually dips, just off the diagram, to 11 Ω at 110 kHz, due to resonance involving the leakage inductance and the total secondary shunt capacitance. This resonance is brought down to about 50 kHz when the 1 nF plus 4.7 kΩ loading is used.

If a frequency-independent input impedance is desired – and it was vital in the days when low-level series faders were used – then the advantage of simple resistance loading of the secondary is very evident, and Figures 8.23 and 8.25 show that this also gives a somewhat better frequency response under practical working conditions. It is certainly the best scheme to use when the transformer and amplifier can be designed as one entity, enabling the resistance loading to be provided by shunt negative feedback via a resistor of much higher value, thus avoiding the degradation of the noise figure that is experienced when passive resistive loading is employed.*

*An additional passive shunt, consisting of *C* and *R* in series, with a very high corner frequency, may also be desirable, in the interests of achieving a good negative-feedback stability margin.

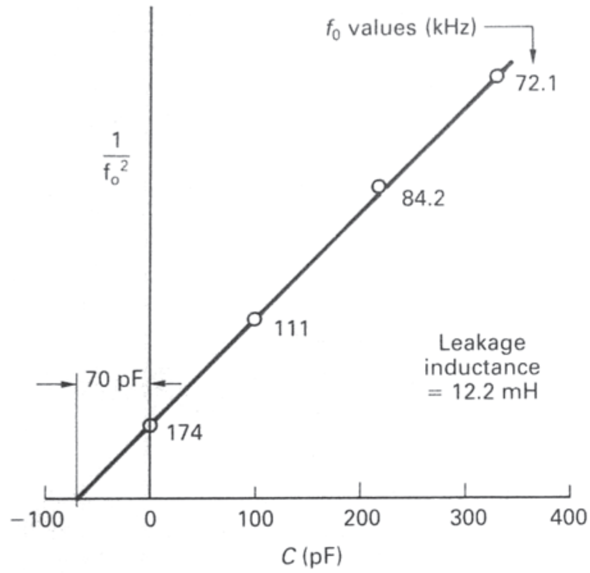


Figure 8.27 *Microphone transformer winding-capacitance and leakage determination.*

A practical method for determining the leakage inductance and effective secondary shunt winding capacitance of transformers such as that here being considered, is to connect several known values of capacitor, C, across the secondary and determine, for each value, the frequency f_0 at which the primary impedance dips to a low value. The core, the screen if fitted, and the primary, should be earthed to the end of the secondary that will be earthed in normal use.

A graph of $1/f_0^2$ (in any convenient units) against capacitor value is then plotted, that for the present transformer being given in Figure 8.27. The winding capacitance is given by the intercept shown. The leakage inductance may then be calculated from the total capacitance corresponding to a point on the graph – well up the graph for best accuracy – and the value of f_0 at that point.

In principle only two points are necessary, and a formula for leakage inductance which avoids the need to draw a graph can then, of course, be easily derived. However, plotting several points and checking that a good straight line is obtained provides reassurance that all is going straightforwardly – complications can sometimes arise with high-impedance transformers.

8.5.5 Modification to provide microphone equalization

Though microphone transformers, and their associated amplifier input circuits, are usually designed with the aim of achieving a flat frequency response at audio frequencies, this approach may sometimes be modified to obtain equalization for the non-ideal acoustic frequency response of a microphone. This turns out to be particularly appropriate with the 4038A ribbon microphone, whose frequency response, though good, falls off somewhat at very low and very high frequencies. The original BBC PGS microphones[12] have a published average axial response, for a large source distance, which is approximately −3 dB at 40 Hz, −2 dB at 10 kHz and –4 dB at 15 kHz. The 4038A commercial version[32] appears to have a high-frequency response not quite as good as this, both from published data and from my own measurements; figures of –2 dB at 10 kHz and –7 dB at 15 kHz are believed to be more nearly relevant.

Referring to the equivalent circuits of Figure 8.19 and the response curves of Figures 8.23 and 8.25, it is evident that what is required for high-frequency equalization is to modify the values so as to bring the resonance peak down to a frequency a little below 20 kHz. Merely increasing the capacitance across the secondary will not do, for the Q-value then turns out to be too low – less than unity. The need is to increase the leakage inductance so that the wanted resonance frequency can be obtained with a smaller value of shunt capacitor. In theory this could be done by increasing the spacing between the primary and secondary windings, and maybe also increasing the number of turns, but both of these changes would increase the winding resistances and hence worsen the noise performance. An alternative is to add extra series inductance externally to the transformer, in series with either the secondary or the primary. Since the latter requires a much smaller inductor value, it is more convenient, and Figure 8.28 shows a design based on this notion. The inductors were made by winding 20 turns of 40 s.w.g. enamelled wire on an R–S components anti-parasitic (ferrite) bead. The flux density in these inductors, even at very high sound levels, is less than 0.1 mT, so that the distortion introduced is negligible. The d.c. resistance is 0.5 Ω each.

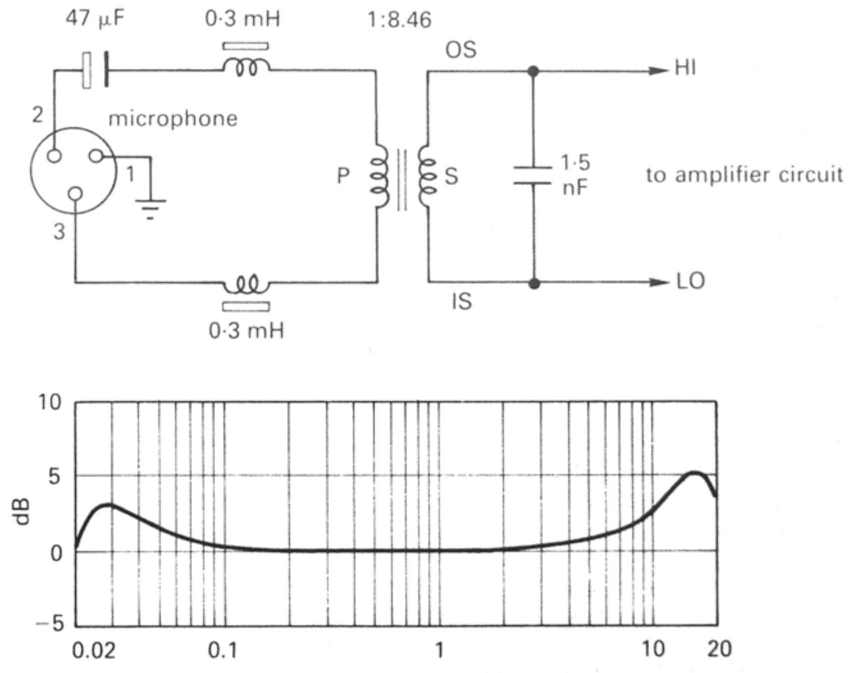


Figure 8.28 *Equalizing scheme for 4038A ribbon microphone. The response curve was measured using the 1000:1 test transformer of Section 8.5.18, at a microphone transformer secondary voltage of 2 mV r.m.s.*

Low-frequency equalization may be achieved by inserting a tantalum electrolytic capacitor of suitable value to give series resonance with the primary shunt inductance and microphone inductance at about 30 Hz. Though the inductance value is somewhat amplitude dependent, the scheme has been found very satisfactory in practice.

The specimens of 4038A microphone actually employed have a resistive impedance at medium and high frequencies which is somewhat higher than the nominal 30 Ω figure, and using the more realistic figure of 42 Ω in the noise calculation at the end of Section 8.5.3 yields a noise figure of 0.9 dB. With a 50 m microphone cable having a go-and-return resistance of 3.5 Ω, the noise figure becomes 1.2 dB. Adding 0.1 dB in accordance with the footnote in Section 8.5.3 gives a practical noise figure of 1.3 dB. With a verified microphone sensitivity of 0.07 mV/μbar, the A-weighted noise SPL is 17.5 dBA. A slightly larger transformer of higher step-up ratio would improve this to 17 dBA.

8.5.6 Eddy-current loss[27]

When a low-frequency a.c. voltage of constant magnitude is applied to a winding on a laminated transformer core, the flux in the core induces a frequency-independent voltage (back e.m.f.) not only into the winding itself, but also into the cross-sectional area of each lamination. This causes eddy currents to circulate round little loop paths within the thickness of each lamination, with accompanying I^2R power losses. The total eddy-current loss for all the laminations may be represented by an equivalent resistance R_e in shunt with the winding.

At low frequencies, where the main flux for a given applied voltage is relatively large, the flux generated by the eddy currents themselves is very small in comparison. But as the frequency is increased, the magnitude of the main flux density falls, until a frequency is reached where it is of the same order as that caused by the eddy currents. The direction of the eddy-current flux is opposed to that of the main flux. The result is that at high frequencies, near the inner region of the cross-sectional area of each lamination, which is inside most of the eddy-current paths in the lamination, the main and eddy-current fluxes nearly cancel each other, whereas no such cancellation effect occurs at the outside of each cross-sectional area. Consequently, at these higher frequencies, there is very little resultant flux in the inner regions of the laminations, the flux being concentrated near the outside – 'skin effect'. The total eddy-current loss is reduced, so that the value of the shunt resistance R_e representing it is increased.

The flux-neutralizing effect of the eddy currents also reduces the shunt inductance of the winding. Because of the distributed nature of the elements involved, the rise in R_e and fall in shunt L with increasing frequency both follow square-root or 10 dB/decade laws. Both effects come in at approximately the same critical frequency, or corner frequency, which is proportional to $\rho/\mu\delta^2$, where ρ is the specific resistance of the magnetic material, μ is the relative permeability and δ is the lamination thickness.

For 0.38 mm Mumetal, with $\mu = 20\,000$, the critical frequency is a little over 200 Hz. The effect of this on frequency response with high source impedances was mentioned in Section 8.5.4. The low-frequency value of R_e for the 1:8.46 microphone transformer, referred to the secondary, is 56 kΩ.

When the frequency is raised well above the critical frequency, the Q-value, if eddy currents were the only source of loss, would tend to unity.

With ferrite cores the specific resistance is so enormously higher than for Mumetal that eddy-current loss is normally quite negligible in transformer applications.

8.5.7 B–H loops and hysteresis loss[27]

In the rationalized MKS system, the magnetizing field strength H is expressed in ampère-turns per metre. For a core of constant permeability, the relationship between B and H is

$$B = \frac{4\pi}{10^7} \times \mu H \tag{8.34}$$

where B = flux density (T)
μ = relative permeability
H = nI/l_m, in which
n = number of turns
I = current (A)
l_m = magnetic path length (m).

At high flux densities the relationship between B and H follows a hysteresis loop of the familiar shape shown by the larger trace in Figure 8.29(a), in which the approach to saturation is evident.* In driving the material round this loop – upwards on the right and downwards on the left – an amount of energy proportional to the loop area is dissipated per cycle.

At lower peak flux densities, such as for the smaller trace in Figure 8.29(a), the shape of the loop is much simpler and follows quite closely a relationship established by Lord Rayleigh well before the advent of radio broadcasting. The Rayleigh equation is

$$B = \frac{4\pi}{10^7}\left[\underbrace{(\mu_i + \alpha\hat{H})H}_{\text{A}} \pm \underbrace{\tfrac{\alpha}{2}(\hat{H}^2 - H^2)}_{\text{B}}\right] \tag{8.35}$$

where B = flux density (T)
μ_i = initial relative permeability
H = magnetizing field strength (AT/m) ($\hat{H}$ = peak value)
α = a constant for the magnetic material

For a given peak H excursion, denoted by $\hat{H}$, term A by itself gives a linear relationship between B and H, with a slope that increases as $\hat{H}$ is made larger. It is term B that opens the line out to form a pointed loop; the – sign is taken while H is increasing positively, and the + sign while it is changing in the opposite direction.

Consideration of equation (8.35) makes it evident that as $\hat{H}$ is reduced, the width of the loop diminishes more rapidly than its length, so that the loop tends more and more closely towards being a simple straight line at very low signal levels. The slanting loop in Figure 8.29(b) has about a quarter of the $\hat{B}$ value that applies to the small loop in (a) and its slimmer proportions are obvious.

To enable the true shape of this slim loop to be more satisfactorily discerned, a signal proportional to H was mixed into the vertical deflection circuit so as to cancel out the mean slope, and the vertical gain was then increased by a factor of 10 to reveal the true shape more effectively. It is seen that the shape of the loop is in very good accord with the Rayleigh representation as 'two pieces of parabola'.

*The vertical voltage for these displays was obtained by using a Blumlein integrator to integrate the voltage, proportional to dB/dt, induced in a secondary winding.

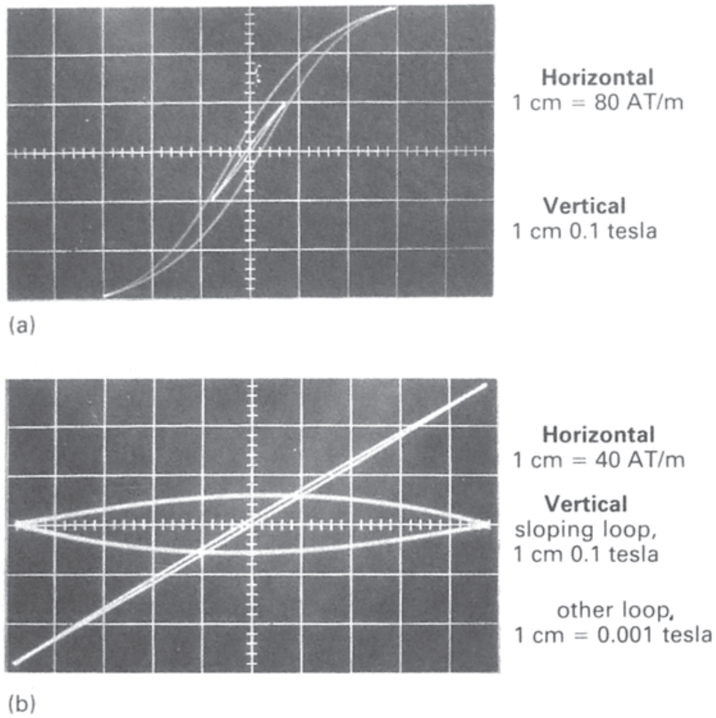


Figure 8.29 *These B–H loops were obtained using a Grade A13 (3H1) ferrite transformer core, to avoid misleading effects caused by eddy currents.*

A consequence of this parabolic shape is that the area of the loop is proportional to $\hat{H}^3$, so that the energy lost per cycle is also proportional to $\hat{H}^3$. But for a constant voltage across the winding, both B and H, at low levels and low frequencies, are inversely proportional to frequency, so that the energy lost per cycle is then proportional to $1/f^3$ and the hysteresis power loss is therefore proportional to $1/f^2$. This, as stated above, is for a constant voltage V across the winding. If V is varied at a fixed frequency, then the energy or power loss is proportional to V^3. Hence when both f and V are varied, we have

$$\text{hysteresis power loss} \propto \frac{V^3}{f^2} \tag{8.36}$$

If the hysteresis loss is represented by a shunt resistance R_h, then it follows that, at low levels and low frequencies,

$$R_h \propto \frac{f^2}{V} \tag{8.37}$$

Hysteresis loss thus tends to become negligible at high frequencies and/or small voltages. It may be added that if the loops had had constant proportions, then R_h would be independent of voltage and directly proportional to frequency.

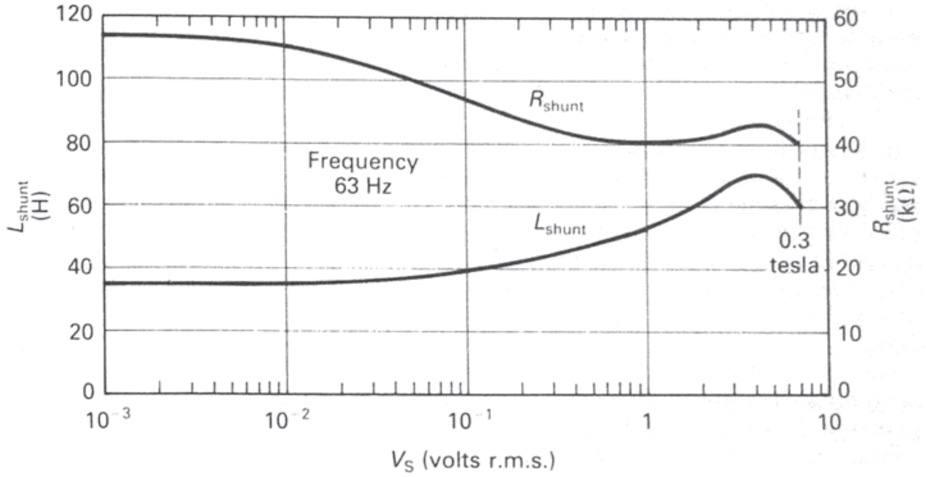


Figure 8.30 *Shunt inductance and resistance measured at the secondary of the 1:8.46 microphone transformer.*

At higher frequencies the eddy-current effects referred to in the previous Section interact with the hysteresis mechanism and matters become more complex, though this is seldom of any practical concern.

Figure 8.30 shows the result of a fixed-frequency measurement on the secondary of the 1:8.46 microphone transformer, made with a simple admittance bridge employing op. amps. The shunt resistance curve virtually represents the total core loss, though strictly speaking a small contribution to it, about 4%, comes from the copper loss.

At low levels the core loss is almost purely eddy-current loss, but as the level is increased there is an increasing hysteresis-loss contribution.

The rise in shunt inductance with measuring voltage reflects the increasing steepness of the *B–H* loops, which continues until offset by the beginning of the saturation process.

8.5.8 Barkhausen noise

Ferromagnetism is a subtle, complex and very fascinating subject[28,29,30]. Normal polycrystalline magnetic materials, such as Mumetal, are believed to contain a large number of magnetic domains, of many different sizes, which are effectively little magnets, each involving very large numbers of atoms. The magnetic effects are basically produced by moving electrons associated with these atoms.

In unmagnetized material the domains are orientated in a random manner, producing zero resultant magnetic flux.

The application of a magnetizing field of smoothly increasing magnitude produces two kinds of effect:

(a) A smooth and continuous movement of the domain boundaries, resulting in a smooth increase in *B*.
(b) Sudden 180° reversals in the orientation of some domains. These are known as Barkhausen jumps, and every time one occurs a little impulse voltage is induced in the exciting winding.

If the winding is connected to an amplifier and loudspeaker, these jumps become audible as separate clicks if the d.c. is changed very slowly, but with more usual rates of change they merge together to produce a rushing sound, called Barkhausen noise.

The effects may be convincingly demonstrated by the following experiment, which is well worth actually carrying out. Put a winding of about 1000 turns on a core consisting of just two or three small Mumetal E laminations, overlapping and held together to give a closed magnetic circuit. Feed a variable d.c. voltage of ±15 V, very well smoothed, via about 10 kΩ to this winding, the winding being connected by means of a suitable value of blocking capacitor to the input of a low-noise transistor amplifier and loudspeaker. There should preferably be some bass cut in the system, to avoid overloading on the more rapid rates of current change.

The sound heard is very much like that produced by tilting a cardboard box containing dry gravel one way and the other, with the same feature that on reversing the direction of tilt there is silence at first.

From such an experiment the following facts emerge:

1. No Barkhausen noise at all is audible for $\hat{B}$ excursions of less than about ±2.5 mT (or ±25 gauss). If any jumps do occur below this level, the domains are evidently so small that their effect is drowned by amplifier noise.
2. The loudest Barkhausen noise, involving reversals of the largest domains, occurs while the operating point is traversing the steepest parts of a B–H loop such as the larger one in Figure 8.29(a).

From an audio engineering point of view, however, the significant practical conclusion is that Barkhausen noise in microphone transformers is never audible at all, with a good margin to spare. With a 4038A microphone and the transformer design already described, it turns out that an SPL of 100 dB at 50 Hz gives $\hat{B} <$ 1 mT.

8.5.9 Residual loss

In materials such as Mumetal, eddy-current and hysteresis losses, already discussed, account for very nearly all the core loss – but not quite all, and the remaining small loss is called residual loss. Residual loss, though it gives rise to a very small phase displacement between B and H under sinusoidal conditions, is different from hysteresis loss in that there is no non-linearity, so that it produces no distortion. It involves a time-dependent mechanism within the magnetic material, which may be of a thermal nature.

If residual loss is represented by a resistance R_r in shunt with the winding, the value of R_r is independent of voltage level and directly proportional to frequency. (See conclusion of paragraph under equation (8.37).)

In ferrite materials the residual loss is also small, but because the eddy-current loss is so very much less than for nickel–iron alloys, the residual loss nevertheless becomes the dominant core loss at very low signal levels, for which the hysteresis loss has become vanishingly small. Thus, whereas residual loss can be forgotten about when using Mumetal etc., it is normally given prominence in trade literature on ferrite materials.

8.5.10 Distortion

When, as is normally the case, a transformer winding is driven by a sine-wave source of much lower internal impedance than its own reactive impedance, an approximately sinusoidal voltage is forced to exist across the winding, so that the waveform of the flux density B is also nearly sinusoidal. But from the pointed shape of the B–H loops discussed in Section 8.5.7, it is evident that the waveform of H, and the magnetizing current, will be much more highly distorted than this.

It is the flow of this distorted magnetizing current in the source impedance that gives rise to the voltage distortion that appears across the windings.

Distortion data for design purposes can be given in different ways, sometimes in the form of a family of curves showing the variation in percentage distortion with flux density for a number of different ratios of source resistance to winding reactance.[31]

The unusual method here presented involves only a single curve, which is essentially a characteristic of the core material itself and is independent of the associated circuit conditions.

Referring to Figure 8.31, the procedure for determining the percentage distortion when a transformer primary is fed from a signal source of internal resistance R_G is

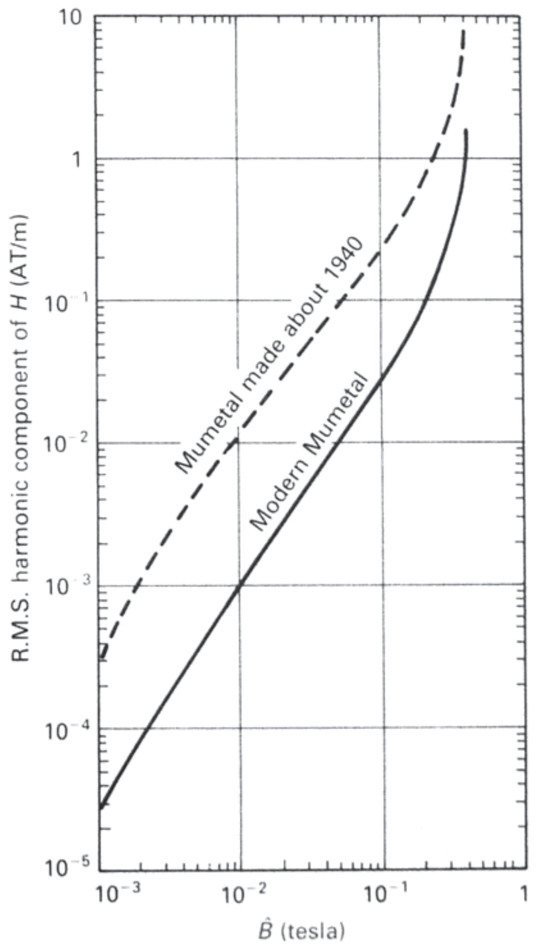


Figure 8.31 *Universal distortion curve for present-day Mumetal laminated cores, plus broken-line curve for early Mumetal.*

(a) From equation (8.33) and the transformer parameters, determine the value of $\hat{B}$ at the frequency and level of interest.
(b) From the curve determine the value of the r.m.s. harmonic component of H, and multiply this by l_m/n to obtain the corresponding r.m.s. harmonic current in amps. (l_m = magnetic path length in metres, n = number of turns in the primary winding.)
(c) Multiply the current value obtained in (b) by R_G to yield the harmonic voltage across the primary winding.
(d) Divide the figure obtained in (c) by the primary fundamental voltage, and multiply by 100 to obtain the percentage voltage distortion.

If the copper resistance of the driven winding is significant, it should be counted as part of R_G. The percentage distortion figure then obtained will apply more accurately to the secondary than to the primary voltage.

The curve of Figure 8.31 is applicable at any frequency, provided this is below the critical frequency mentioned in Section 8.5.6. However, because $\hat{B}$ falls with rising frequency in a transformer handling constant voltage, so also does the distortion; the usual requirement is therefore to know its low-frequency value. This reduction in distortion with rising frequency is an attractive feature of transformers not possessed by amplifying devices such as transitors.

If a small gap is inserted in a transformer core, for example by inserting all *T*s one way and all *U*s the other, this will markedly increase the fundamental component of the magnetizing current for a given applied voltage, but it will hardly affect the harmonic component of this current and will therefore hardly affect the percentage distortion. The Figure 8.31 curve is therefore still fully applicable, and this also applies to the effect of the small unwanted gaps that are inherent in ordinary laminated core assemblies but which are absent in toroidal cores. The curve may thus be regarded as a true characteristic of the material itself, except that slight complications arise at very high flux densities, as discussed in Section 8.5.11, and as a result of unequal magnetic properties along the rolling direction and at right angles to it as mentioned in Section 8.5.13.

The square-law slope of the Figure 8.31 experimental curve at very low levels is consistent with predictions based on the Rayleigh equation (8.35).

However, the curve is best regarded more as a guide to orders of magnitude than as highly accurate design information, since quite large variations occur in magnetic material production and between the products of different manufacturers.

Application of the above method to the 1:8.46 microphone transformer design of Section 8.5.2 gives a distortion figure of 0.14% for 1 V r.m.s. at 40 Hz on the secondary. The distortion at low and medium levels is mainly third harmonic.

8.5.11 Saturation flux-density values

The saturation flux density for Mumetal is usually given in the literature as about 0.8 T, but if a test is made on an ordinary laminated Mumetal transformer core, it will be found that saturation appears to set in at around 0.4 T.

The reason is that when the flux in one lamination approaches a slight gap, it is diverted into the two neighbouring laminations, which therefore experience a doubling of their flux density in this region. Thus a mean flux density in the core of about 0.4 T results in a flux density near the joints of around 0.8 T, and saturation therefore occurs.

This effect is shown very clearly in Figure 8.32, from which it is evident that if a large enough magnetizing field is applied, saturation can be made to occur throughout the whole core at about 0.8 T.

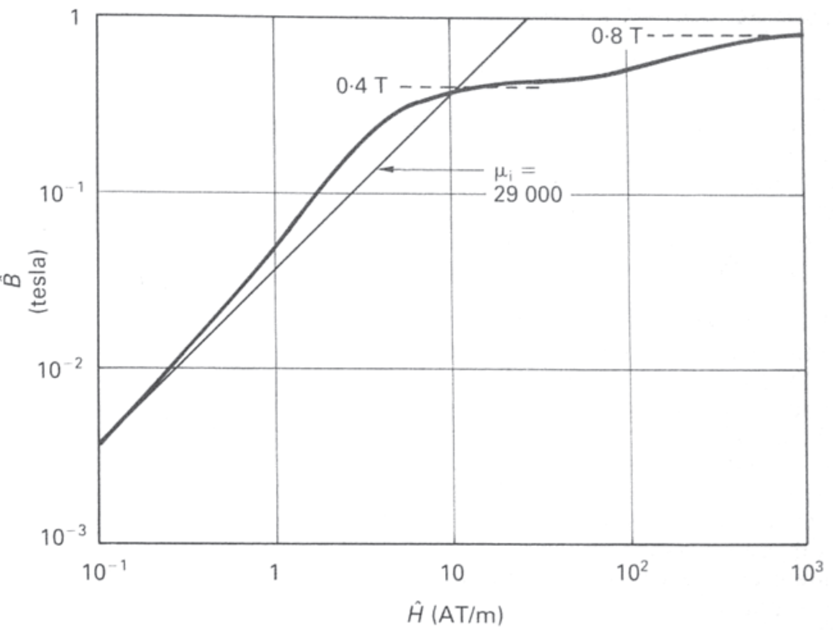


Figure 8.32 *50 Hz measurement on high-μ specimen of Mumetal core (0.38 mm laminations), showing the double-saturation effect described in Section 8.5.11.*

8.5.12 Effects of altering the input transformer size

To give a satisfactory performance with a specific value of source resistance, the number of primary turns, n_p, required, as already discussed, is normally determined by low-frequency considerations. It is of interest to find the effect on this, and on other parameters, of altering the size of the transformer. This is a straightforward problem related to the transformer equations (8.32) and (8.33) and other basic formulae for copper resistance and winding capacitances.

If all linear dimensions are multiplied by the same factor m, then it turns out that, for a constant primary inductance:

(a) $n_p \propto m^{-1/2}$
(b) $\hat{B}$ for a given voltage $\propto m^{-3/2}$
(c) Low-level harmonic distortion at constant voltage $\propto m^{-3}$
(d) Copper resistances $\propto m^{-2}$
(e) Winding capacitances $\propto m$
(f) The leakage inductance and the shunt eddy-current resistance are unaffected.

Thus doubling the linear dimensions reduces the winding resistances to a quarter of their original values and makes the distortion eight times smaller. By also using thin laminations (0.1 mm) for reduced eddy-current loss, and adopting an optimum step-up ratio, a system noise figure of less than 0.5 dB can be achieved in practice.

8.5.13 Toroidal transformers

Toroidal transformers employing spiral or 'clock-spring' cores have certain advantages.

Firstly, the direction of the flux coincides everywhere with the direction of rolling of the magnetic material in manufacture, the highest permeability always occurring in this direction.

Secondly, the absence of slight air gaps such as occur in ordinary transformer cores where the laminations butt together also contributes to enhanced effective permeability values, and avoids the saturation effect discussed in Section 8.5.11.

When these features are combined with the use of superior magnetic materials such as thin Supermumetal or Supermalloy, effective initial a.c. permeability values of 200 000 or more can be realized.

A third advantage of toroidal transformers is that the magnetic hum pick-up is much reduced. However, even if a simple winding occupies the full circumference quite uniformly, it is actually still equivalent to a single circumferential turn so far as hum pick-up is concerned. Consequently, when only a few turns are involved, the advantage of toroidal construction in this respect is not as great as might at first be expected.

A toroidal transformer enclosed in a Mumetal case is employed in the 4038A ribbon microphone.[32]

A disadvantage of toroidal construction, of course, is that winding is less easy, increasing the cost.

8.5.14 Hum pick-up and Mumetal cans

The stray magnetic field produced by mains transformers, both in theory and in practice, varies inversely as the cube of distance, except when very close.

With the normal type of single-bobbin microphone input transformer, a Mumetal screening can is always necessary, and is normally of the deep-drawn cylindrical type with an overlapping lid.

The amount of hum reduction obtained is very significantly influenced by the orientation of the transformer within the case, the least effective condition being when the axis of the transformer bobbin coincides with the axis of the can.

The appreciable reluctance of the junction between the lid and the body of the can results in the dominant magnetic flux inside the can tending to be in the axial direction, but even in the absence of the lid, the axial flux density falls off with increasing distance from the open end at a rate tending towards the theoretical piston-attenuator slope of 41.8 dB per diameter, as illustrated in Figure 8.33.

Thus the transformer should be mounted as far from the lid end as possible, with the bobbin axis at 90° to the can axis. Non-magnetic spacing material, of about 1 mm thickness, should be used to prevent the transformer core from coming into direct contact with the can.

Holes or slots of normal sizes in the can or lid exert negligible influence on the screening effectiveness when the transformer is positioned as above, but should be drilled fairly gently to avoid permanent deformation of the Mumetal.

When the above points are attended to, a properly annealed deep-drawn Mumetal screening can will be found to give at least 50 dB of hum reduction at 50 Hz.

Many capacitor microphones contain diminutive output transformers which can give rise to hum, though the problem is much less acute than with amplifier input transformers fed from electromagnetic microphones – for three reasons. Firstly, because the distance from mains transformers is normally much greater;

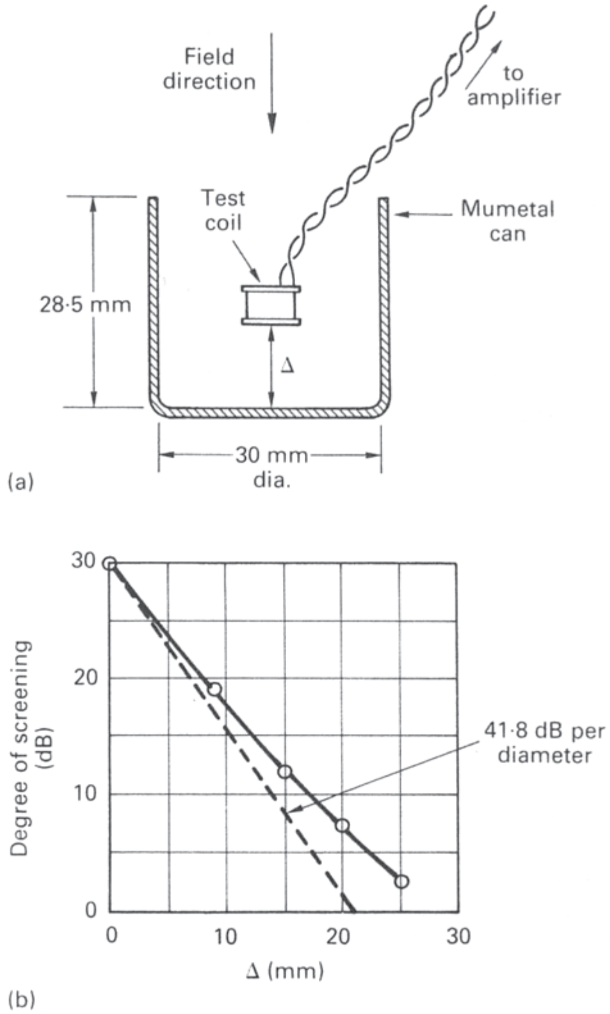


Figure 8.33 *Measurement on Mumetal screening can.*

secondly, because the signal level is usually higher than with other types of microphone; and thirdly because these internal transformers are fed from the low output impedance of an emitter-follower or other feedback circuit – an ideal transformer fed from an active circuit with zero output impedance would give no hum pick-up at all.

Mumetal screening for such internal microphone transformers is therefore frequently omitted.

8.5.15 Bifilar windings and electrostatic screens

When a microphone feeds an amplifier input transformer via a balanced cable, the set-up is as shown diagrammatically in Figure 8.34, and troubles are sometimes experienced due to both audio-frequency interference, for example

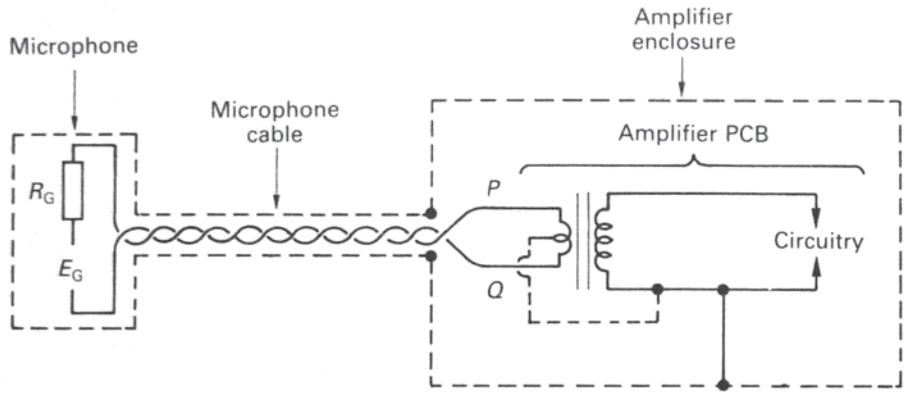


Figure 8.34 *Interconnection of microphone and amplifier.*

from nearby power cables, and RF interference. The latter is made audible by demodulation at the semiconductor junctions.

Audio-frequency balanced-mode interference can be induced magnetically into the small loop areas formed by the inner conductors, but this effect is reduced to a very low level indeed by the use of good twisted-pair or preferably quad cables.

The effects now to be considered involve longitudinal or common-mode interference appearing in the same phase at P and Q, and it is the RF components of such interference that usually constitute the main problem. For such interference the cable behaves as if it was a coaxial one.

One mechanism by which common-mode interference can get onto the inner conductors involves the fact that braided or lapped cables have tiny gaps in their screening – lapped cables, particularly those of the double-lapped or Reusen type, are generally much better than braided ones in this respect.

Another very significant mechanism involves the resistance of the cable.[33] Consider the simpler case of a length of coaxial cable, short-circuited at one end. This will behave as an aerial, in which, in the absence of any resistance, equal voltages would be developed along the lengths of both the outer and inner conductors, giving zero voltage between them at the non-short-circuited end. RF current, however, flows in this 'aerial', but is largely confined to the outer conductor because of skin effect. This current produces a voltage drop in the resistance of the outer conductor, and since almost no such effect occurs in the inner conductor, the voltage drop appears between inner and outer at the free end of the cable. So far as this effect is concerned, braided cables are better than lapped ones, having generally lower resistance.

Thus, by more than one mechanism, a length of microphone cable is always found to have a common-mode RF interference voltage between its inner conductors and the screening, sometimes of a few millivolts magnitude. The internal impedance with which any specific component of this interference voltage is presented is highly dependent on the length of the cable in relation to the wavelength of the interference.

If the cable inners are provided with a common-mode earth via a centre-tapped winding, as shown in broken-line in Figure 8.34,* a very substantial RF

*For the transformer described in Section 8.5.2, the common-mode inductance seen between the ends of the bifilar primary and its centre-tap is only 0.6 μH.

interference current may flow in this earthing lead, which, with normal constructional arrangements, would be taken to earth somewhere well inside the amplifier enclosure. An RF magnetic field is thus created which causes interference voltages to be induced into the rest of the circuitry – an effect which good designs always aim to minimize.

The general experience seems to be that it is normally better not to earth the centre-tap but to leave the primary winding floating, and this is now the nearly universal practice. Occasionally a switch has been provided so that a centre-tap may be used as an optional alternative – it is sometimes the better choice if the main interference is at high audio frequencies owing to a leaky cable screen.

With the centre-tap omitted, some interference current, usually of much smaller magnitude, will flow to earth via the transformer capacitances, and it used to be almost universal practice to include an earthed electrostatic screen between the transformer windings to prevent this current reaching the secondary winding. However, because of the low secondary impedances made possible by the advent of transistors, it is found that entirely satisfactory results can usually be obtained without involving the extra cost of a screen – many BBC microphone transformers have no screen, for example.

The low impedance allows a sizable capacitor, e.g. 1 nF, to be connected across the secondary circuit, and this capacitor can be put extremely close to the input transistor for effective VHF interference suppression. Sometimes a series inductor may be added for further filtering action. (The connection of such a capacitor directly between base and emitter of an input stage is a somewhat questionable technique, since it can cause oscillation by creating a negative input conductance.)

An overriding consideration is that the outer screen of a microphone cable should be taken solidly to earth right at the input socket and not via a lead wandering around near the amplifier circuitry.

8.5.16 Transformers and negative feedback

The benefits of negative feedback may be applied to microphone transformers, high-level line-input transformers, and output transformers.

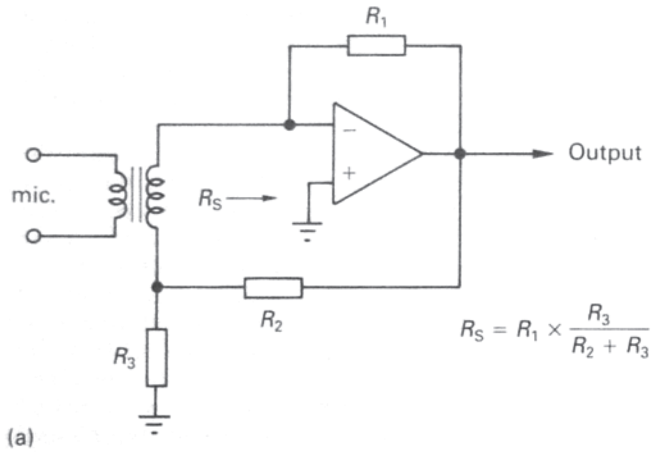
Figure 8.35 shows two ways in which a desired value of secondary loading resistance, R_S, (see Section 8.5.4) may be produced without introducing significant degradation of the noise figure. The (a) scheme is that used in the BBC microphone amplifier circuit of Figure 8.6.

The formula given for the (b) scheme is a general one, equally relevant in other contexts where the circuit may be used to produce negative resistance. For the present application, A is made negative, say –100.

Figure 8.36 shows, slightly simplified by the omission of small capacitors across the op. amp. resistors, a circuit designed for KEF Electronics Ltd for use in a studio monitor loudspeaker with self-contained amplifiers. The aim is to provide a high-impedance floating input, suitable for bridging across any available signal line, and giving a high degree of rejection of common-mode interference present on the line.

The transformer is wound highly symmetrically on a two-section bobbin, with the half-secondaries, wound in reverse directions, at the inner radius for minimum resistance – the primary resistance is of no importance whatever, and thinner wire was in fact used for this winding. The primary and secondary are well spaced apart by low-loss insulation, giving an inter-winding capacitance of less than 30 pF.

To maintain overall symmetry of behaviour, the amplifier that follows the transformer is also balanced, both inputs being virtual earths – or virtual grounds if this term is preferred!



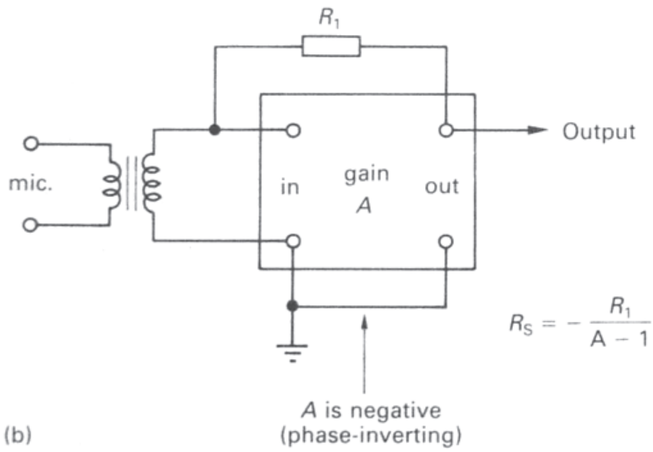


Figure 8.35 *Circuits giving effective secondary loading of value R_S, with negligible sacrifice of signal-to-noise ratio.*

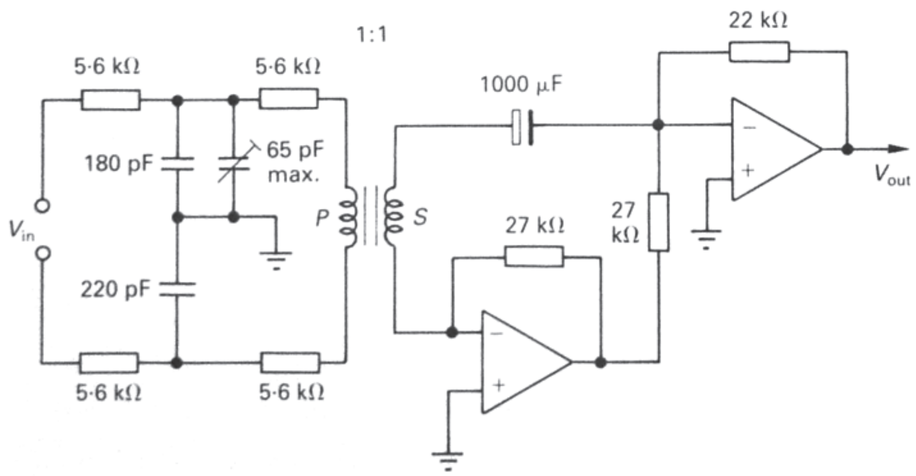


Figure 8.36 *Balanced input circuit for active studio monitoring loudspeaker.*

Feeding a transformer in this way to an almost zero impedance is enormously beneficial with regard to frequency response and distortion, since the core hardly has to develop any flux density at all.

The distortion for an input level of 0.775 V r.m.s. at 30 Hz, both as predicted using the universal curve of Figure 8.31 and as directly measured, is well under 0.001%.

The frequency response is flat within ±1 dB from 2 Hz to 100 kHz, and immaculately flat over the whole audio band. There is a peak of about 5 dB at 0.8 Hz owing to series resonance between the secondary inductance and the 1000 μF capacitor, but this may be eliminated, if desired, by replacing this capacitor by a network involving two capacitors and a resistor.

The noise level relative to 0.775 V is –108 dBA, and the unweighted hum level is also about –108 dB even with the screened input transformer about 20 cm from a large mains transformer.

With the trimmer capacitor correctly set, the common-mode rejection ratio is better than 90 dB at 1 kHz and better than 70 dB at 10 kHz.

The transformer used in the version of the scheme described above is wound on a Mumetal core of the same size as for the microphone transformer of Section 8.5.2, and has 1000 turns on both primary and secondary. A more economical version, using a ferrite transformer core, is used by Quad Electroacoustics Ltd for a professional amplifier input stage.

Reducing the shunt inductance to a fairly low value has no adverse effect on the audio frequency response, though it gives an increase in low-frequency noise, which nevertheless still remains inaudible.

A different embodiment of the same broad principle, evolved quite independently, is described in Reference 34, and employs only a single op. amp. The secondary winding feeds the inverting input of this, but the feedback resistor, instead of being taken to this input, is taken to a tertiary winding. Assuming infinite op. amp. gain, this scheme ideally reduces the flux density right down to zero, which allows an extremely small transformer to give a very good performance.

An alternative way to achieve the above result, which I tried at the time the Figure 8.36 circuit was evolved, but found it unnecessary to adopt, is to apply a little positive feedback to one of the op. amps., making it into an open-circuit-stable negative resistance (see below) of magnitude equal to the secondary copper resistance of the transformer.

Transformers are often used to provide a high-level floating line output for feeding to other equipment. Because of the use of negative feedback, the output impedance is usually very low, but in self-contained studio installations the loading imposed on such an output is normally quite light. Line output voltage levels are usually expressed in dBu, i.e. decibels relative to 0.775 V r.m.s. – see footnote near the end of Section 8.3.11. Sometimes, however, such as in outside broadcast equipment, which has to feed into a long telephone line, the requirement is for a purely resistive output impedance of specific value, such as 75 Ω.

With early transistor equipment,[17,18] push–pull PNP output stages, with a centre-tapped transformer primary, were normally used, the negative feedback being preferably derived from a tertiary winding.[35,36]

The later availability of complementary transistors, and more particularly of suitable op. amps., has enabled such transformers to be driven in a 'single-ended' manner, the simplest such arrangement being that of Figure 8.37.

The distortion in the voltage across the primary is made very low by the negative feedback, but the secondary voltage will contain a small amount of distortion owing to the flow of the highly distorted magnetizing current in the

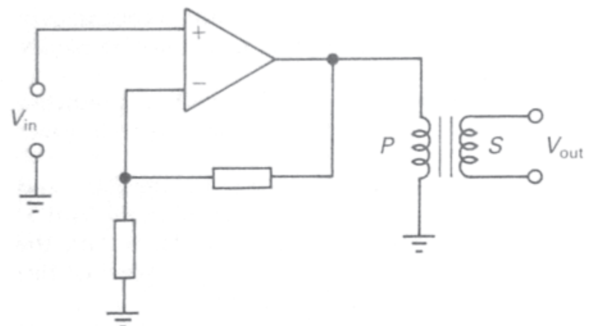


Figure 8.37 *Simplest op. amp. circuit for low-impedance driving of output line transformer. To avoid significant performance degradation caused by d.c. core polarization, the gain should preferably be limited to about ×2 and no appreciable d.c. voltage should be present at the input terminal.*

primary copper resistance (see Section 8.5.10). The distortion due to this cause is greater with small transformers than it would be with large ones, because of item (d) in Section 8.5.12, but is nevertheless likely to be much less than 1% for full output level at 30 Hz, given a reasonably good transformer.

There are two things that can be done to give a substantial reduction in the above distortion:

(a) Take the negative feedback, or some of it, from a tertiary winding. This will reduce the distortion of the flux waveform and therefore also of the output voltage.
(b) Insert negative resistance in series with the primary to annul the effect of the copper resistance.

Good results can be obtained in either of these ways, but (b) is especially attractive because it simplifies the transformer and reduces its cost, while requiring only trivial elaboration of the op. amp. drive circuit.

Negative resistance circuits are of two broad types – open-circuit stable and short-circuit stable; simple examples of these are shown in Figure 8.38. In Figure 8.38(a), with the input open-circuited, negative feedback is greater than positive, giving stability, but if the input is short-circuited, only positive feedback remains and the circuit runs into a saturated state. In circuit (b), a short-circuit on the input makes the negative feedback dominant and gives stability.

The (a) circuit, with values chosen to give a negative resistance of magnitude just slightly less than the primary copper resistance, can be inserted directly in series with the primary earth-return lead in Figure 8.37, and will give a good reduction in low-frequency distortion. However, this scheme as it stands is rather undesirable, since it is on the verge of instability, giving a large magnification of the op. amp. offset voltage, and being liable to run away with itself if resistance values change slightly.

The problem may be solved by inserting an electrolytic capacitor in series with the primary winding, the negative resistance then being set to be nominally equal to the copper resistance.

The circuit is now fully stable, but exhibits a sub-audio-frequency response peak caused by series resonance between the electrolytic capacitor and the primary shunt inductance. This can be removed by connecting resistors, ideally of value $\sqrt{(L/C)}$, across both the capacitor and the primary winding. The negative-resistance circuit then sees a constant-resistance network having a resistance

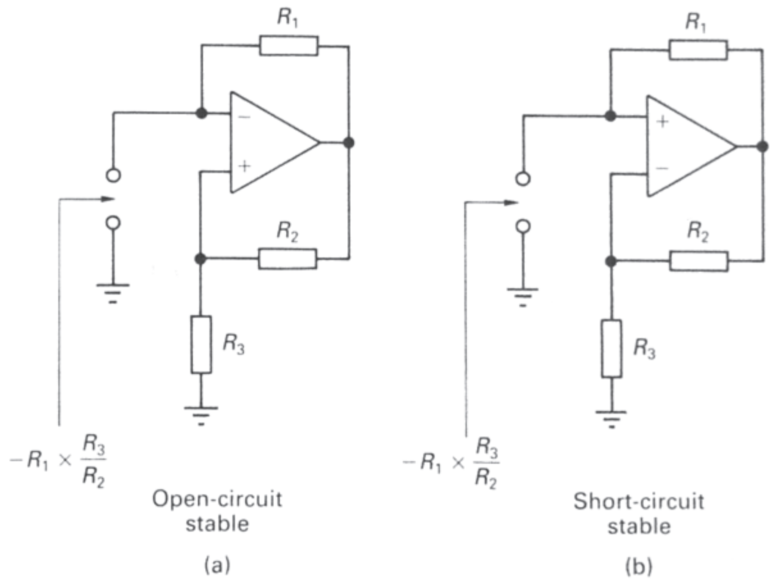


Figure 8.38 *Negative-resistance circuits.*

magnitude far higher than the magnitude of its own negative resistance, and the set-up is therefore thoroughly tame and non-resonant.

I have used the above scheme very effectively in a precision measurement context, with the electrolytic capacitor in series with the top end of the transformer primary. This makes available between the top end and earth, a voltage which is the true induced voltage, unadulterated by the voltage drop in the copper resistance. The ratio of this voltage to the secondary voltage is then determined with high precision purely by the turns ratio, if the windings are appropriately positioned.

Though the use of a separate op. amp. to produce the negative resistance in the above manner is a perfectly satisfactory practical scheme, an even more economical arrangement is shown in Figure 8.39. In this, stability is assured even if R_1 is increased well beyond the value that gives perfect annulling of the copper resistance, provided C_1 is not made excessively large. The effect of a very large C_1 value is to make the circuit liable to maintain a very-low-frequency relaxation oscillation if R_1 is increased a little beyond the critical value. A value of C_1 giving a response −3 dB at 5 Hz and −1 dB at 10 Hz is usually a sensible choice; then an increase in R_1 of even 20% above the correct value will not normally allow oscillation to be sustained. (Since the copper resistance of the winding has a positive temperature coefficient of 0.4%/°C, R_1 should ideally also have this coefficient.)

In the absence of C_2, and with C_1 adequately small, increasing R_1 by a large amount, say 100%, may first produce oscillation at a frequency of the order of 1 MHz, owing to series resonance in the transformer. Though not likely to be troublesome, the possibility of such oscillation is totally eliminated when C_2 is present, and the right value of C_2 is also effective in giving the circuit an excellent square-wave response.

A frequency-response flat to within ±0.1 dB from 20 Hz to 20 kHz is readily obtainable on no load.

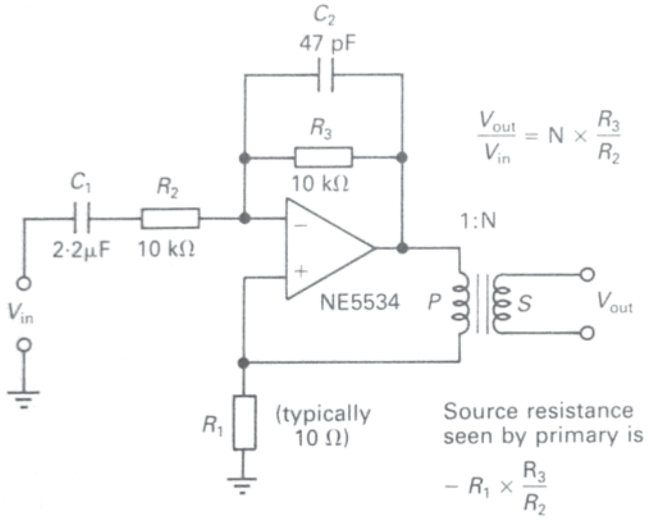


Figure 8.39 *Output circuit incorporating negative resistance to reduce distortion due to copper resistance of primary winding of line transformer.*

When a resistive load is applied, however, the presence of appreciable leakage inductance will cause the output to fall off at high frequencies, and this, at first sight, would seem to be an inherent weakness of the Figure 8.39 scheme.

The alternative tertiary-winding solution mentioned earlier can overcome this trouble, since, if sufficient care is taken, the tertiary winding can be so positioned on the bobbin that the leakage flux linking with it is equal to that linking with the secondary winding[35]. The negative feedback then effectively monitors the secondary voltage and thus ensures a flat frequency response at the output terminals. Even then, however, the primary-to-secondary leakage inductance causes the maximum output voltage obtainable on load, before voltage clipping occurs, to be less at high frequencies than at lower frequencies.

The most elegant solution is to reduce the total leakage inductance to an extremely low value by winding the primary and secondary together as a bifilar pair. This gives far lower leakage inductance than can be achieved in other ways – a ratio of main inductance to leakage inductance of well over 100 000:1 is readily obtained.

A simple bifilar winding, of course, restricts the transformer ratio to 1:1, but this limitation can be overcome as described later.

An NE5534 op. amp. on ±15 V supplies can comfortably give an output voltage swing of ±12 V, or 8.5 V r.m.s., so that, with a 1:1 transformer, this is also the output voltage and corresponds to +20.8 dBu. With a 600 Ω load on the output, the maximum obtainable output will be slightly reduced because of the winding resistances, totalling perhaps 20 Ω, but a level of +20 dBu should still be achieved. This requires a peak current of 18.3 mA, which the op. amp. is capable of supplying.

If the output resistance of the circuit is made up to 75 Ω by adding series resistors, and is fed to a 75 Ω load, the achievable output level is limited not by voltage clipping, but by the maximum output current that the op. amp. can turn on, which may be taken as 25 mA. This gives 23 mW mean to 75 Ω or approximately +14 dBm. The peak voltage at the primary is only about 4 V.

A higher output voltage, on no load or a light load, may be obtained by adopting a trifilar winding, with two sections in series.

With this 1:2 ratio, a no-load voltage level of over +26 dBu becomes available, but when a 600 Ω load is applied, the maximum output obtainable is limited to a peak voltage of 7.5 V by the maximum op. amp. current of 25 mA. This corresponds to a voltage level of +16.7 dBu, and the power to the 600 Ω load is, of course, +16.7 dBm.

On a 75 Ω load with 1:2 transformer, the maximum level obtainable is 6 dB less than with the 1:1 transformer, and is therefore only +8 dBm.

To meet all requirements when a 1:2 ratio is used, it may be necessary to increase the current capability of the drive circuit. This could be done by including complementary emitter-followers within the op. amp. feedback loop, but an alternative solution, which has appeared in Lundahl literature, is shown in Figure 8.40, and gives a doubling of the current capability. Moreover, the adoption of antiphase driving of the two primary sections has the great virtue that capacitive coupling from primary to secondary is ideally of a fully balanced nature, thus avoiding any capacitive common-mode output.

The reasons why 'multifilar' windings have come so much into their own in this application are:

(a) Simple ratios of 1:1, 1:2 or 1:0.5 are appropriate.
(b) The inherently much higher capacitances of such windings are of little consequence because of the low-impedance nature of the circuits, and because the negative feedback is taken directly from the primary rather than via transformer action from another winding. When tertiary-winding feedback is used, the behaviour of the transformer at frequencies far above the top of the audio band needs to be carefully taken into account.[35] An interesting combination of techniques, however, is to use tertiary-winding feedback, but with the normal separate tertiary winding replaced by one section of a multifilar winding.

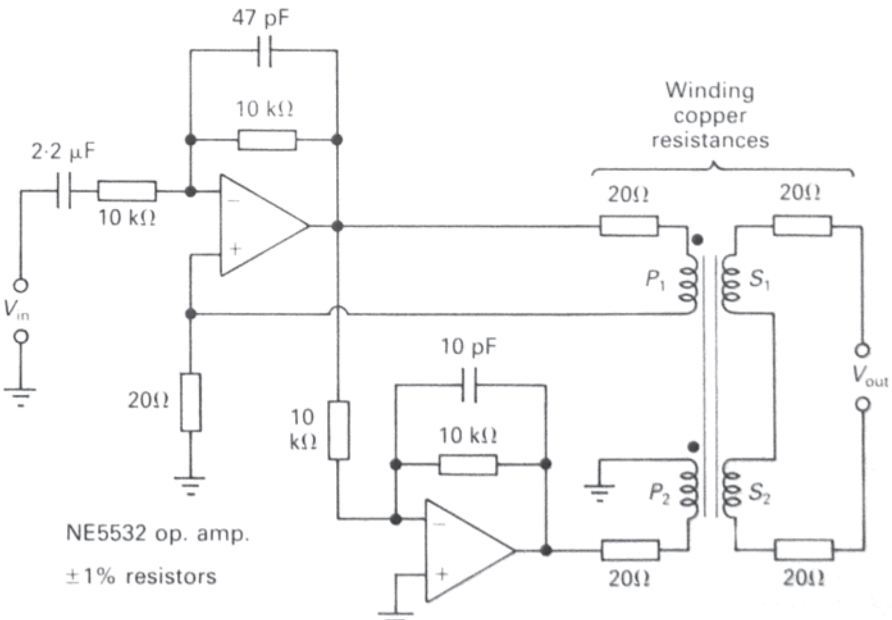


Figure 8.40 *Balanced op. amp. line output circuit employing 'multifilar' transformer, with negative-resistance compensation of primary winding resistance. The capacitively-coupled common-mode output is very low. Winding resistances are shown explicitly to aid understanding.*

(c) With multifilar windings, voltages of the same order as the input voltage exist between directly adjacent turns, but the low voltages involved eliminate any insulation problems.

Mumetal is not a satisfactory core material to use for these output transformers, since its relatively low saturation flux density would necessitate putting on so many turns that the copper resistances would be intolerably high, unless an uneconomically large size of core was used. The very high permeability of Mumetal, which is its special virtue, is not needed for hard-driven output transformers, and Radiometal, which is a nickel–iron alloy with about 50% nickel, is sometimes used. It has a saturation flux density of about 1.6 T.

A good alternative, having about the same saturation flux density, is grain orientated silicon steel (GOSS metal, invented in the mid 1930s, with extraordinary appropriateness, by Dr N. P. Goss!). This is marketed as Hipersil etc., and is used in the form of spiral-wound toroids, or as C-cores. The joining surfaces in the latter are ground flat to give intimate magnetic contact, and are held firmly together by a steel tension band. The double-saturation effect described in Section 8.5.11, and shown in Figure 8.32, is absent.

Further progress has been made, and continues to be made, in the evolution of improved transformer-core materials.

To conclude this Section, Figure 8.41 illustrates a novel technique for feeding audio signals at very low distortion through balanced lines of moderate length.

The receiving end, at the right, is basically that of Figure 8.36 but without the input resistors; it therefore presents a low impedance to the line. This low impedance appears also across the secondary of T_1, augmented somewhat by the line resistance and, at high frequencies, inductance. Thus the flux densities in both transformers are held very low, so that small Mumetal or ferrite cores may be used, which is neat and economical. A response flat to within ±0.1 dB from 20 Hz to 20 kHz is obtainable, with distortion of 0.01% or lower, and excellent signal-to-noise ratio.

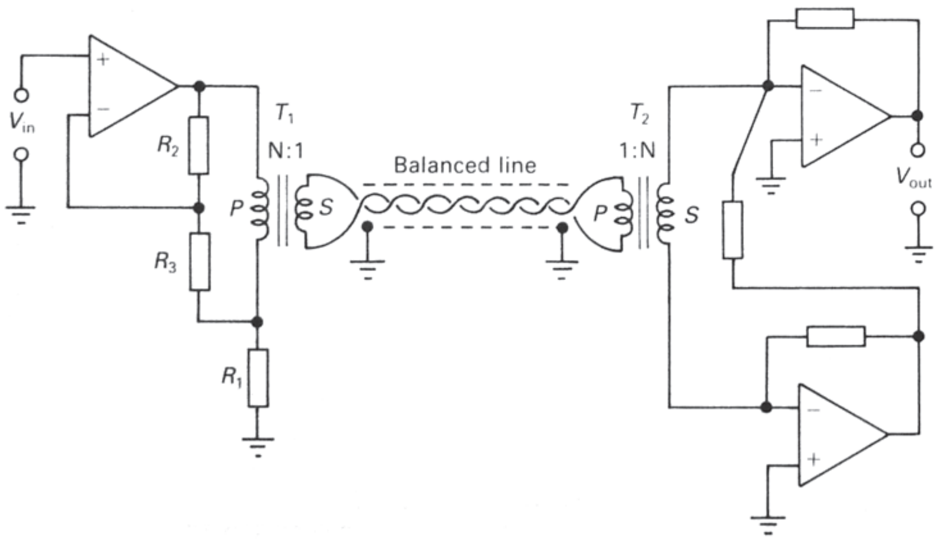


Figure 8.41 *Line operation on virtual-earth basis.*

The resistor network associated with the primary of T_1 applies a combination of current and voltage negative feedback such as to make the effective source resistance R_G feeding the primary have the value given by equation (8.38):

$$R_G = R_1 \times \frac{R_2 + R_3}{R_3} \tag{8.38}$$

R_2 and R_3 have relatively high values, giving negligible passive loading effect. This arrangement gives good stability, whereas employing only current feedback would be liable to lead to instability, and would result in the occurrence of gross overloading if the line was disconnected. A small capacitor should be shunted across R_2.

A ratio N of about 5:1 is generally appropriate, with R_G about 10 kΩ. There is no need for the transformers to have very low leakage inductance. When designing such a system it is helpful to bear in mind the typical microphone-line values $R = 0.07\,\Omega/\text{m}$ (go-and-return), $L = 0.7\,\mu\text{H/m}$ and $C = 150\,\text{pF/m}$. These give $Z_0 \approx 70\,\Omega$ and a velocity of propagation of about $10^8\,\text{m/s}$, i.e. about one third of the free-space velocity; wavelengths in the cable are therefore about three times shorter, for a given frequency, than they would be in free space.

It may be noticed that the series electrolytic capacitor of Figure 8.36 has been omitted in Figure 8.41. This is an economy that may be made, though it is liable to result in an offset at the amplifier output of a substantial fraction of a volt. However, if positive feedback is used to annul the secondary resistance of T_2, as mentioned in connection with Figure 8.36, then the capacitor must be included.

8.5.17 Higher-impedance microphone transformers

As mentioned in Section 8.5.2, it was desirable in the valve era to step the microphone impedance up to quite a high value to achieve a good noise performance and to subdue valve microphony. This, however, tends to make the frequency response much more dependent on variations in source impedance than with modern lower-impedance designs, since the resonance frequency no longer occurs several octaves above the audio band.

A very interesting example is the input transformer used in the BBC OBA/8 amplifier of Figure 8.5, in which the 300 Ω source resistance is stepped up to 300 kΩ at the valve grid. The transformer is quite large, using size 101A Mumetal laminations, which are 7.6 cm (3 in) tall. The secondary is wound in one narrow section in the middle of the bobbin, resulting in a total shunt capacitance of only about 50 pF, including the input capacitance of the top-grid valve.

The leakage inductance is intentionally made large, by having the low-profile primary well spaced away to the side of the tall secondary. A capacitor with a value of about 1.5 nF is shunted across this primary, so that the equivalent circuit, at medium and high frequencies, referred to the secondary, becomes approximately as in Figure 8.42(a), and functions as a π-section low-pass filter. The leakage inductance is such that the cut-off frequency is a little above 10 kHz, and the characteristic impedance is somewhat less than 300 kΩ; this requires a leakage inductance in the region of 7 H.

The effect of varying the source resistance over a wide range, as when series faders are used (see Section 8.2), is considerably less with such a π-filter design than when the input capacitor is omitted, and a response remaining within ±1 dB limits up to 10 kHz can be obtained.

It may be noticed, in Figure 8.42(a), that the symbols L and C have been used to denote ½-section values. This unconventional usage, much to be recommended, is due to my colleague E. F. Good, and results in the cut-off frequency for low-

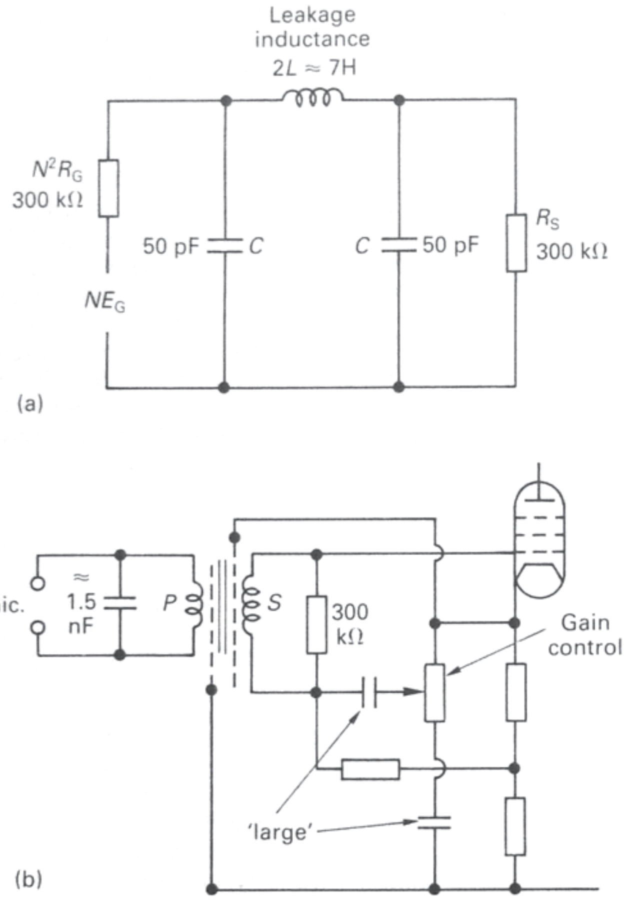


Figure 8.42 *Parameter values and input circuit relating to BBC OBA/8 amplifier illustrated in Figure 8.5.*

pass and high-pass filters, and the resonance frequency of tuned circuits, all being given by $f_0 = 1/2\pi\sqrt{(LC)}$ instead of by three different formulae. The characteristic impedance, however, is still given by $R_0 = \sqrt{(L/C)}$.

The manner in which the above transformer is incorporated in the OBA/8 circuit is shown in essence in Figure 8.42(b). The subtle use of two electrostatic screens prevents alteration in the setting of the negative-feedback gain control (which is ganged to a passive inter-stage control) from significantly affecting the value of the right-hand filter capacitance, which would cause the frequency response to vary with the gain setting.

A basic shortcoming of the OBA/8 design is that the presence of the passive 300 kΩ resistor across the transformer secondary causes a loss in signal-to-noise ratio, with a 300 Ω source, of ideally 3 dB. The possibility of virtually eliminating this loss by using negative feedback via a resistor of much higher value to terminate the filter had not been thought of in those early days.

There are many other applications in which the parameters of a transformer may be effectively exploited to obtain a wanted filter characteristic.

8.5.18 A 1000:1 test transformer

Figure 8.43 gives details of a simple but very useful test unit, which may be inserted in series with a balanced microphone line, with the microphone present, to check the frequency response and gain under normal working conditions.

It is intended for feeding from a 600 Ω oscillator. The internal 68 Ω resistor attenuates the oscillator e.m.f. by a factor of 10 (within 2%) and provides a source resistance close to 60 Ω for energizing the transformer primary. Thus an oscillator e.m.f. of, say, 1 V r.m.s. injects 0.1 mV r.m.s. (–78 dBu) in series with the microphone.

The frequency response is flat to well within ±0.1 dB from 20 Hz to 50 kHz, and the total harmonic distortion for a 20 Hz output level of 1 mV r.m.s. (–58 dBu) is less than 0.1%, being much less than this for higher frequencies and/or lower levels.

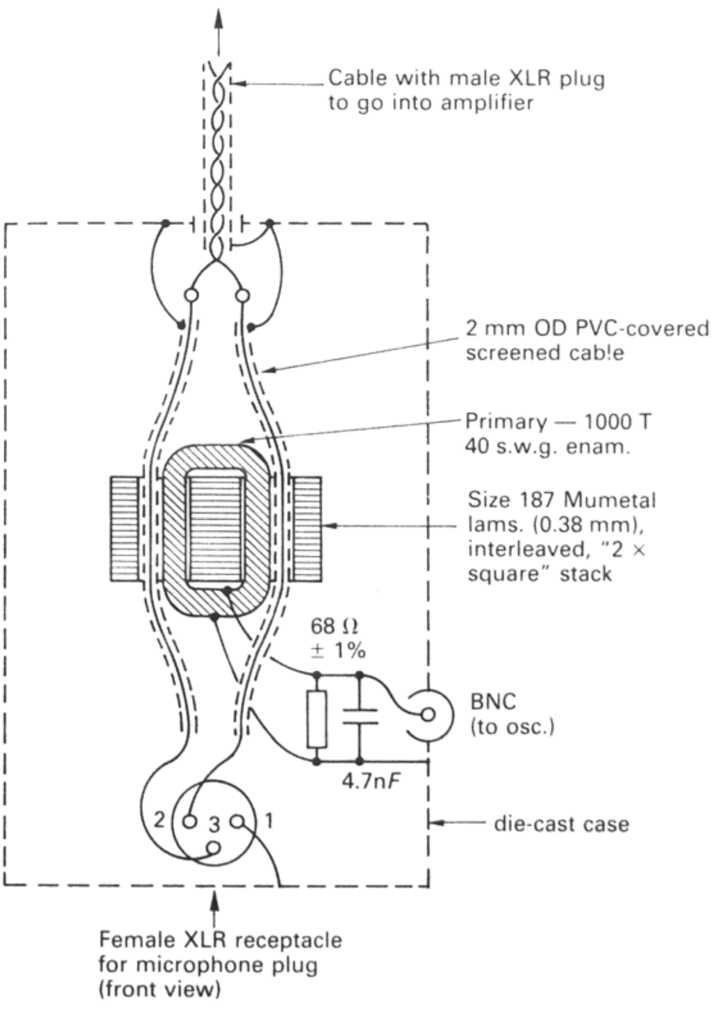


Figure 8.43 *Unit for testing microphone circuits. For clarity, the bobbin, and the top part of the core, have been omitted.*

The 4.7 nF capacitor across the input subdues the excitation of internal ringing of the 1000-turn winding, at frequencies in the region of 1 MHz, if a square-wave input is used, and an excellent non-overshooting 20 kHz square-wave output is then obtainable, with a rise-time of about 0.5 μs.

Owing to the rather curious configuration of the leakage flux field at such high frequencies, it is found that the output ringing is minimized when the effective 1-turn secondary formed by the screened leads is located midway between the two bobbin cheeks, and it is worth arranging for this to be the case.

At 20 Hz a virtually perfect square wave is obtained, the droop being only just discernible on careful inspection.

Magnetic screening of the transformer is quite unnecessary, but the effective loop area of the 1-turn circuit should be kept as small as possible, preferably by twisting the two pieces of cable together. Care should be taken, of course, to avoid allowing the cable screening to form a shorted turn – the screening of the two pieces must be earthed at one end only, as shown.

The XLR connectors and their contact numbering are arranged in accordance with the widely adopted IEC268/BS5428 Standard.[37]

8.5.19 A transformer for loudspeaker talk-back

A good loudspeaker may be operated in reverse as a very high-quality microphone.

It is shown in Reference 38 that if the loudspeaker output current is fed to an amplifier having an extremely low input impedance, and a frequency-response rising with frequency at 20 dB/decade throughout the audio spectrum, then the axial frequency response as a microphone will be identical to that obtained in normal loudspeaker use.

The Reference gives details of more elaborate circuit arrangements than that presented here, which provide lower distortion, better signal-to-noise ratio, and more accurate equalization, but the simple ferrite-cored transformer shown in Figure 8.44 constitutes a highly satisfactory practical compromise, very easy to implement.

The primary inductance is about 32 μH, which has a reactance of 2 Ω at 10 kHz, and because a ferrite core is used, the reactance is very linearly proportional to frequency. Thus, with an 8 Ω loudspeaker, a reasonable approximation to the ideal zero input impedance is provided.

Sound pressure levels up to at least 100 dB can be handled with adequately low distortion.

The 680 pF capacitor across the secondary, augmented by up to about 100 pF of cable and/or amplifier input capacitance, gives resonance at approximately 20 kHz, the resonance being well damped by the 22 kΩ resistor. The capacitor serves to suppress RF interference, and the damping, in addition to giving more closely the required frequency response, is also beneficial in reducing the risk of oscillation when this inductive source is connected to a microphone amplifier having internal negative feedback. A balanced output may be taken if preferred.

The use of a centre-tap on the bifilar primary is the best choice for this application, since unscreened loudspeaker cable is often used. It is found that this usually gives no interference problems. But it is wise to keep the loudspeaker away from television sets, whose line time-base magnetic field can cause trouble.

Provided the transformer is not much less than 1 m from mains transformers, no audible hum will be experienced even if no screening can is employed. The use of a gapless transformer core is preferable in this respect to having a gapped core with more turns, though the latter would give reduced distortion.

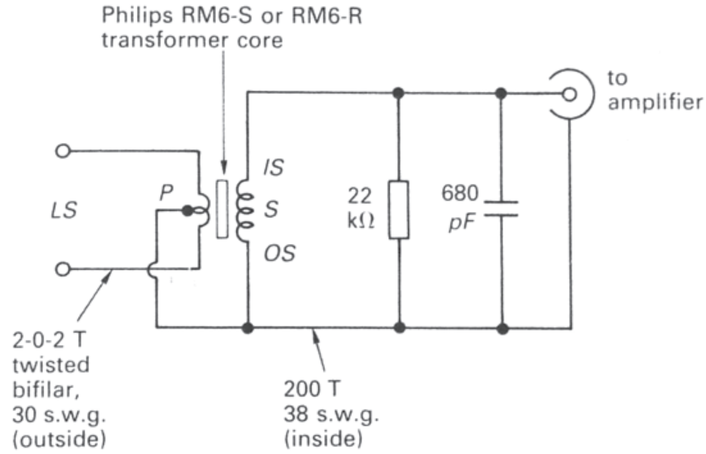


Figure 8.44 *Loudspeaker-talk-back transformer unit. An additional 1-0-1 T primary may be provided if desired, to give a more accurate frequency-response at high frequencies for measurement purposes, but at the sacrifice of 6 dB of signal-to-noise ratio.*

The transformer core employed has a minimum A_L value of 1900 nH (type RM6-S) or 2000 nH (type RM6-R). If only a gapped RM6 core is to hand – e.g. as obtainable from *R–S* Components – it may be very easily converted to the required gapless core by placing a piece of fairly fine wet-or-dry abrasive paper, used dry, on a good flat surface, and rubbing the appropriate half of the core on this for a minute or two.

Applications of the above principle to the subjective assessment of loudspeakers, to frequency response measurement, and to absolute sensitivity determination, are discussed in Reference 38.

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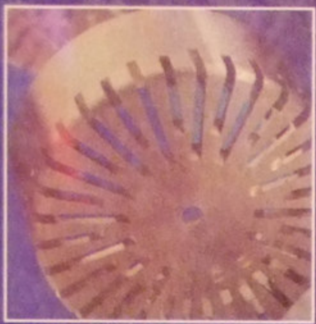

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