The Journal of

THE BRITISH INSTITUTION OF RADIO ENGINEERS

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"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."

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INTERPRETATION OF RESPONSIBILITY*

ALL professional bodies employ the term "responsibility" in listing the qualifications required for membership. It is a term also used, albeit loosely, in advertisements for a very wide range of appointments. In fact, of course, every employed person has some degree of responsibility. The term has, however, a more precise definition in regard to the qualities required of an applicant for corporate membership of a professional body. In the case of engineering Institutions, the emphasis is on *technical* responsibility and the interpretation is of particular importance when considering proposals for membership.

The most obvious examples are:—

(i) technical responsibility for the work of other personnel;

(ii) technical responsibility for continuous development of research projects;

(iii) technical responsibility for production.

Where an applicant's work involves the supervision of a number of assistants, his responsibility can usually be measured by the number and qualifications of his staff.

Where an engineer is not responsible for other technical personnel, but has considerable responsibility for technical work, as in (ii), the problem becomes more difficult. In such a case satisfactory supporting evidence would be by the publication of a suitable original paper or by the candidate holding patents.

In most of these cases evidence is available in the form of publication work done which is the legitimate ambition of every engineer and research worker. Experience has shown that such ambitions are not inconsistent with the interests of firms engaged in competitive industry; indeed such publications are of great commercial value, in addition to being extremely valuable in assessing the abilities of engineers.

There remains, therefore, the problem of the engineer who, because of the structure of the organization for which he works, finds it difficult to obtain permanent establishment in the particular grade which would ensure his eligibility for corporate membership of his professional body. On the other hand, the Institution is judged by the standing of its members and therefore by the interpretation which it places on the term "responsibility." In those circumstances, the discretionary powers of the Membership Committee are restricted.

It needs to be emphasized, therefore, that election to corporate membership recognizes that the candidate has achieved a satisfactory balance of theoretical knowledge, technical experience *and* technical responsibility.

The first requirement is, therefore, evidence of satisfactory training and proof of knowledge —normally recognized by election to Graduateship. The value of such training lies not so much in the actual facts learned, but in the training of the mind to grapple with technical problems. Thus equipped, the Graduate is able to meet theoretical and practical problems which occur in actual industrial practice and, by such experience, is able to go on and acquire responsibility.

Every corporate member also incurs responsibility when proposing a candidate for election. In this sense, the proposer is entrusted to uphold the quality of membership of his professional body.

* A similar editorial was published in the December 1953 *Journal* but [illegible] revised version is published for the guidance of members who sometimes experience difficulty in deciding whether they can support a proposal for membership.

INSTITUTION NOTICES

OBITUARY

The Council has learned with regret of the deaths of the following members and has expressed sympathy with their relatives:—

Captain Andrew Macfadyen Houston Fergus (Member) died on the 24th December at the age of 63. Captain Fergus served in the Royal Artillery during the 1914-18 War as a Signals Officer and subsequently practised as a consultant engineer in Scotland and in Jersey. He returned to England in 1937 and settled in Surrey. During the last war he carried out special radio monitoring work for the War Office. First elected an Associate Member of the Institution in 1927, Captain Fergus was transferred to full Membership in 1929; he served as a member of Council from 1939–41.

* * *

Edward Hedgecock (Associate Member) died recently in New Westminster, British Columbia, at the age of 56. Mr. Hedgecock served during the 1939-45 War as a Warrant Officer in the Signals Branch of the Royal Navy and in 1949 went to Canada. He was elected an Associate Member of the Institution in 1942.

* * *

Squadron Leader William Alfred Spacey (Associate) died suddenly on 7th October, aged 45. At the time of his death he was serving as a Squadron Commander at No. 3 Radio School, R.A.F. Squadron Leader Spacey was elected an Associate in March 1956.

Radio and Electronic Component Show

This year's Radio and Electronic Component Show will be held at both Grosvenor House and Park Lane House, Park Lane, London, W.1, from April 8th to 11th inclusive.

The R.E.C.M.F. has discontinued the lapel badge system of admission for visitors: this year double tear-off tickets will be issued which will admit to both sections of the Show. These tickets will be issued from the R.E.C.M.F. offices and all Brit.I.R.E. members wishing to attend should apply in writing direct to The Secretary, Radio and Electronic Component Manufacturers' Federation, 21 Tothill Street, Westminster, S.W.1, and *not* to the Institution.

New Year Honours

The Council of the Institution has congratulated Flight Lieutenant Francis William Benson (Associate Member) on his appointment as a Member of the Military Division of the Most Excellent Order of the British Empire.

Flight Lieutenant Benson is with No. 90 Group R.A.F. and is concerned with ground radar equipment. He was transferred to Associate Membership in 1955.

1957 Physical Society Exhibition

The Physical Society Exhibition of Scientific Instruments and Apparatus takes place from Monday 25th to Thursday 28th March, inclusive, at the Halls of the Royal Horticultural Society in Westminster, London.

The opening ceremony will be performed by Professor P. M. S. Blackett, F.R.S., at 11 a.m. on Monday, 25th March, and the Exhibition will be open as follows:

Monday: 10.30 a.m.–7 p.m.
(members only 10.30–2 p.m.)

Tuesday: 10 a.m.–9 p.m.

Wednesday: 10 a.m.–7 p.m.

Thursday: 10 a.m.–4.30 p.m.

The Physical Society has provided special admission tickets to enable Institution members to visit the Exhibition on the Physical Society's Members' morning, Monday, 25th March, from 10.30 a.m.–2 p.m., when the Exhibition is not so crowded. These special tickets may be obtained on application to the Librarian of the Institution, 9 Bedford Square, London, W.C.1. Members are asked to note that these are the *only* tickets available from the Institution this year.

Entrance to the Exhibition at all other times is by ticket only and these may be obtained, free of charge, from the offices of the Physical Society, 1 Lowther Gardens, Prince Consort Road, S.W.7.

In connection with the Exhibition a handbook of Scientific Instruments and Apparatus of some 300 pages will be published in which are included detailed descriptions of the various items to be exhibited. It is available at 6s. 0d., 7s. 6d. post free, from the Society.

1957 CONVENTION—"ELECTRONICS IN AUTOMATION"

June 26th to July 1st—Cambridge

GENERAL ARRANGEMENTS

Lectures

All lectures will be held at the Cavendish Laboratory, Free School Lane, in the Clerk Maxwell and/or the Green Lecture Theatres.

As preprints of all papers will be available before the Convention, authors will present their contributions briefly during the first half of each session; after a short break for refreshments the second half of the session will be devoted to a discussion of the papers.

Preprints

Registration for the Convention will entitle delegates to receive in advance preprints of all papers being presented. These will *not* be available to persons not attending the Convention until *after* it has been held. A limited number of complete sets of preprints will then be available at a charge to be announced later. All papers presented at the Convention will subsequently be published in the Institution's *Journal*, with a record of the discussions.

Accommodation

Limited accommodation is available in King's College for delegates staying for the *whole* period of the Convention, i.e. from Wednesday evening, June 26th, to Sunday afternoon, June 30th. Arrangements can also be made for those wishing to stay in College until Monday morning, July 1st, when the Convention officially disperses and when a number of special visits will be arranged.

Delegates intending to be present for only part of the Convention should make their own arrangements regarding accommodation.

Convention Banquet

Following the pattern of previous Institution Conventions, an official Banquet has been arranged for Friday, 28th June, at 7.30 p.m. in King's College Hall. All persons staying for the whole period of the Convention and resident in King's College will be invited to attend. Members not resident in King's College should make early application for tickets to attend the Banquet (which will cost £2 2s. each).

Group Discussions and Films

The Chairman of each session will make arrangements for group discussions to be held in King's College subsequent to the presentation of papers and any demonstrations.

Programmes of technical films on relevant aspects of automation will be shown in King's College on the evening of Wednesday June 26th and at other times to be announced later.

Visits

Although the programme of the Convention is very full, it is felt that there will be many members who will wish to take the opportunity to visit firms and other establishments in the Cambridge area. Some visits will therefore be arranged to take place on the morning of Monday, July 1st, following the official conclusion of the Convention.

Clerk Maxwell Memorial Lecture

The Clerk Maxwell Memorial Lecture was founded by the Institution in 1951. This year's lecture will be given by Professor Sir Lawrence Bragg, F.R.S., Director of the Davy Faraday Laboratory of the Royal Institution, on the evening of Thursday, June 27th.

Registration

The charges for registration are as follows:—

For the whole period of the Convention with accommodation in King's College (June 26th-June 30th inclusive), meals, Convention Banquet and preprints:

Members	...	£12 10s. 0d.
Non-members	...	£15 10s. 0d.

Registration covering attendance at all six main sessions and the summing-up session, including preprints but exclusive of accommodation and catering:

Members	...	£4 0s. 0d.
Non-members	...	£7 0s. 0d.

Registration forms for members are inserted in this *Journal*.

Forms for non-members may be obtained on application to the Institution.

"ELECTRONICS IN AUTOMATION"

Programme of the 1957 Convention

Wednesday, June 26th (Evening)

Registration of members staying in King's College.

7—7.30 p.m. Dinner for residents only.

During the evening a programme of films on electronics and automation will be shown in the *Reading Room.*

Thursday, June 27th

9.00 a.m. Opening address by the President of the Institution, Mr. G. A. Marriott, B.A.

Session 1

Office Machinery and Information Processing

Chairman: A. D. Booth,
D.Sc., Ph.D., M.Brit.I.R.E.

Scope of Session:

Computing elements, logical design and input/output devices in the field of information processing and automatic control.

9.15 a.m. Presentation of papers.
10.45 a.m. Morning coffee.
11.00 a.m. Discussion of papers.
1.00 p.m. Lunch.

Session 2

Machine Tool Control

Scope of Session:

The use of electronics in the control of speed or position of machine tools; systems using both analogue and digital computation will be also considered, together with ancillary equipment.

2.30 p.m. Presentation of papers.
4.00 p.m. Afternoon tea.
4.15 p.m. Discussion of papers.
6.30 p.m. Dinner.

8.00 p.m. **The Third Clerk Maxwell Memorial Lecture** to be delivered by Professor Sir Lawrence Bragg, F.R.S.

Friday, June 28th

Session 3

Chemical and other Processes

Chairman: Denis Taylor,
Ph.D., M.Sc., M.Brit.I.R.E.

Scope of Session:

The use of electronic equipment in the control of manufacturing processes from raw material to finished product. Determination of system transfer functions, and the use of continuous recording and sample testing in, for example, the chemical industry, will be considered.

9.15 a.m. Presentation of papers.
10.45 a.m. Morning coffee.
11.00 a.m. Discussion of papers.
1.00 p.m. Lunch.

Session 4*

Automatic Measurement and Inspection

Chairman: J. E. Rhys-Jones,
M.B.E., M.Brit.I.R.E.

Scope of Session:

Presenting some examples of the introduction of automation into the field of measurement and inspection. This includes examples of sensing or measuring together with means of handling the data so obtained.

2.30 p.m. Presentation of papers.
4.00 p.m. Afternoon tea.
4.15 p.m. Discussion of papers.

7.30 p.m. Reception by the President and Mrs. G. A. Marriott.

8.00 p.m. **Convention Banquet** in King's College Hall.

In addition to delegates staying in King's College a limited number of guests, including ladies, may be invited. Delegates requiring additional tickets should make early application.

**Note.* The attention of members is drawn to the rearrangement of Sessions 4, 5 and 6.

Saturday, June 29th

Session 5*

Simulators

Chairman: Professor D. G. Tucker, D.Sc., Ph.D., M.Brit.I.R.E.

Scope of Session:

Analogue computers and their uses, e.g.

(1) in determining system behaviour by simulating known characteristics of the operational system,

(2) in deducing system characteristics from data recorded on a system in normal operation,

(3) in solving complex systems of equations arising from problems of aircraft structures, field distribution, linear programming, etc.

9.15 a.m. Presentation of papers.
10.45 a.m. Morning coffee.
11.00 a.m. Discussion of papers.
1.00 p.m. Lunch.

Session 6*

Automation in the Electronics Industry

Chairman: L. H. Bedford, C.B.E., M.A., B.Sc., M.Brit.I.R.E.

Scope of Session:

The title, almost an inversion of the main theme, includes:—

(1) Automation of assembly processes based principally on the printed circuit.

(2) Automation of production of passive components, e.g. resistors and capacitors (other than printed elements).

(3) Automation of production of active circuit elements, viz. thermionic valves and semi-conductor devices.

2.30 p.m. Presentation of papers.
4.00 p.m. Afternoon tea.
4.15 p.m. Discussion of papers.
6.30 p.m. Dinner.

Sunday, June 30th

Morning Service in King's College Chapel.

1.00 p.m. Lunch.
3.00 p.m. Summing-up Session.

A general discussion on the lessons of the Convention. To be opened by the Chairmen of all the Sessions.

Monday, July 1st

Arrangements are being made for parties to visit industrial organizations where electronic equipment is either manufactured, or being used in automation.

Other visits may be arranged to organizations manufacturing instruments, electronic equipment and motor cars, and to research establishments. Small groups will also be invited to see the work in progress in the Mathematical Laboratory of the University of Cambridge.

Full details of these visits will be circulated to members who have registered for the Convention and requests to participate in such visits must be made before the commencement of the Convention. In some cases there may be a small charge to cover transport.

Demonstration of Equipment

Provision is being made for authors of papers to demonstrate equipment in an adjoining room in the Cavendish Laboratory. Demonstrations will be given before and after each session, and an exhibition of equipment will be open to delegates during the period of the Convention.

All equipment will be relevant to the papers presented. In cases where the exhibition of a large installation is impracticable, models, photographs or similar portable demonstration material may be displayed.

Hotel Reservations

Members who have not reserved accommodation in King's College are advised to secure accommodation elsewhere in Cambridge as early as possible. A list of Hotels in the district may be obtained from the Institution, but members must make their own subsequent arrangements.

Garage Accommodation

Garage accommodation in Cambridge is very limited and parking is *not* permitted within the precinct of King's College. A list of addresses providing garage facilities may also be obtained from the Institution and members are advised to make arrangements with a garage as early as possible.

NINTH EXTRAORDINARY GENERAL MEETING

In accordance with the notice circulated to all corporate members and also published in the December 1956 issue of the *Journal*, an Extraordinary General Meeting of the Institution was held at the London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1, on Wednesday, 30th January 1957, commencing at 6 p.m.

In the absence of the President, who was indisposed, the Chair was taken by Mr. William E. Miller, M.A. (Member), a Past President of the Institution, who was supported by other Officers and members of Council. Twenty other corporate members were present at the commencement of the meeting and the Secretary reported that a number of proxies had been received in favour of the motion.

Mr. Miller explained to the meeting that the first main change to be made was the introduction into the Articles of the terms *electronics* and *electronic engineering* to implement the alterations which had previously been made in the Memorandum of Association; the proposed changes to the Articles gave more definite expression to the qualifications required of members in radio and electronic engineering.

Adoption of the new Articles would also give effect to the increase in subscriptions recommended by the Council and referred to in the Annual Report of the Institution which was approved at the Annual General Meeting held on 30th October, 1956.

For these reasons, legal advice and help had been secured in completely redrafting the Institution's Articles of Association and before moving adoption of the Special Resolution Mr. Miller invited comment.

Mr. D. C. H. Mellon (Member) stated that although it was not his intention to vote against the motion, he wished to express an opinion on the steps being taken to enable the Institution's Articles to embrace electronics and electronics engineers. Originally the Institution's membership was concerned with electronic devices capable of radiating electromagnetic waves, and he felt that the Institution should not be concerned with matters outside that field.

In reply, the Chairman stated that the Council had given very careful consideration to such points as had been raised by Mr. Mellon. During the last few years the Professional Purposes Committee had been charged with the responsibility of considering such changes as might clearly indicate the Institution's work beyond the usually accepted field of radio communication, which included television and radar. Recommendations that the Institution might embrace in its title the name "Electronics" had also been considered but it had been generally agreed that there was no need for the Institution to depart from its main title, having regard also to the general acceptance of the definition of electronics.*

Mr. Miller also drew attention to the fact that the Institution's Graduateship Examination had recently been amended to include electronics as a subject, and to make provision for engineers who were not concerned only with communication techniques.

On behalf of the Council of the Institution, the Chairman then moved from the Chair that the corporate members of the Institution be asked to approve the Special Resolution:

> "That the Articles of Association contained in the printed document laid before this Meeting and subscribed for identification by the President be and the same are hereby adopted as the Articles of Association of the Institution in substitution for and to the exclusion of its existing Articles."

The motion was seconded by Mr. G. B. Ringham (Member) and, with the one abstention, was carried unanimously.

The meeting concluded at 6.30 p.m.

The revisions to the Articles of Association have been approved by the Board of Trade. Corporate members have already received a copy of the revised Articles and non-corporate members will be able to peruse them in detail in the next issue of the Institution's List of Members which will be circulated by the end of April 1957.

* In a report published by the Institution in 1944, electronics is defined as describing "the wider uses of the radio valve and kindred devices: it is radio technique at work in new ways and in widely diverse fields."

THE APPLICATION OF TRANSISTORS TO A.M. BROADCAST RECEIVERS *

by

B. F. C. Cooper, B.Sc., B.Eng. †

SUMMARY

The performance specification of a transistor receiver, which is to be competitive with existing four-valve portable receivers, is first laid down. It is shown that with a single-ended class A output stage a minimum of six transistors is needed to meet this specification, or seven transistors when using a class B output stage. The latter design is to be preferred for reasons of running economy. Problems encountered in the design of the component circuits of the receiver are considered in some detail, and the design of an experimental receiver is described. Running costs of transistor and valve receivers are compared.

1. Introduction

With the advent of transistors capable of performing efficiently at broadcast frequencies the transistorized broadcast receiver has become a commercial proposition. A number of papers has been written on this subject[1-6] and production models have already appeared on the American market. Their advent on the Australian market should therefore be not very far distant. The emphasis at the present time is naturally on portable radios where the greatest benefit is reaped from the inherent running economy of transistors. The low running costs of a transistor receiver will compensate for the higher first cost which may be expected in view of the fact that more transistors, resistors, capacitors, etc., are required to make a receiver with performance comparable to that of existing valve receivers.

This paper discusses the circuit problems involved in the design of transistor receivers. A particular experimental model is described and the running cost is compared with that of a typical valve-equipped portable radio.

2. General Considerations

Although adequate performance can be achieved in metropolitan areas with quite simple transistor receivers this paper will be concerned only with superheterodyne receivers which are competitive with valve receivers of the same type.

Initially the performance of a 4-valve portable radio might well be taken as a target. Such a receiver having a loop aerial followed by a mixer, an i.f. amplifier, a detector/audio frequency amplifier, and an output stage, has typical performance figures as follows:

Maximum audio output	250 mW
Overall sensitivity for 50 mW output and 30% modulated signal	200 microvolts/metre‡
Equivalent noise side-band input (e.n.s.i.)	10 microvolts/metre

Since transistors amplify signal power rather than signal voltage it is necessary to compute first the signal power available from the loop aerial in order to determine the overall gain required in the transistor receiver.

Now the voltage applied to the grid of the converter valve in the receiver considered above will be approximately 100 microvolts for a field strength of 200 microvolts per metre. In this case the loop is virtually unloaded but since it has a resonant impedance of $Q\omega L$, where L is loop inductance, it is capable of delivering a power $E^2/4Q\omega L$ to a matched load, where E is the open-circuit voltage across the loop. We have typically $L \cong 200 \mu H$, $Q=100$, and for $\omega = 2\pi \times 10^6$, and $E=10^{-4}$ volt the available signal power is 2×10^{-14} watts. This

* Reprinted from *The Proceedings of the Institution of Radio Engineers, Australia,* Volume 17, October 1956. (Paper No. 383.)

† Division of Radiophysics, C.S.I.R.O., Sydney, N.S.W.

U.D.C. No. 621.396.62:621.315.59.

‡ This is a fairly representative sensitivity although some radios of this type may reach 100 microvolts/metre.

figure is the available carrier power, and the available sideband power is smaller by a factor $m^2/2$ where m is the modulation factor. Here $m=0{\cdot}3$ so that the available sideband power is 9×10^{-16} watts. To achieve an output power of 50 mW an overall power gain of $4{\cdot}4\times10^{13}$ or approximately 136 db is therefore required.

A five-valve portable receiver with r.f. stage will have a gain some 20 to 26 db higher than a four-valve receiver, and the equivalent transistor receiver will therefore need that much extra gain. Considering, however, the receiver of lower sensitivity a few preliminary remarks can be made about the number of transistors required to achieve the necessary power gain. The transistor receiver will contain the same component circuits as its valve counterpart, namely oscillator-mixer stage, i.f. amplifier, second detector and audio amplifier. Since present transistors give higher gain at audio than at radio frequencies it is desirable to use as much audio gain as possible in order to reduce the amount of radio frequency gain required. However this process cannot be carried too far since the detector will be required to operate at a low power level where it is difficult to maintain linearity and achieve a satisfactory a.g.c. characteristic. Tests show that this point is reached with transistor detectors at a detected power level of about 1 microwatt. Hence there is nothing to be gained by providing more than about 55 db of audio gain and this can be achieved with a driving stage followed by an output stage.

This leaves about 80 db to be provided in the convertor, i.f., and detector stages. Up-to-date transistors can provide 30 db or more gain in i.f. stages and 20 db in convertor stages when working into matched loads. In practice 2 db loss per stage should be allowed for coil losses and mismatching. A gain of 20 db relative to sideband power can be obtained in a transistor detector with optimum matching to the following audio stage. However as described later a simple resistance-capacitance inter-stage coupling will usually be preferred after the detector, in which case a detector gain of approximately 10 db can be realized. Allowing 30 db combined gain in the convertor and detector stages the i.f. stages must provide about 50 db of gain and this can be done comfortably with two high-gain transistors. Transistors available for the experimental receiver described in this paper were not of advanced design and a third i.f. stage was found necessary.

It is seen from the foregoing that the minimum number of transistors which may be used corresponds to an oscillator-mixer transistor, two i.f. transistors, a detector, an audio driving transistor, and a class A output transistor, i.e. a total of six transistors. Alternatively where a class B push-pull output stage is used there will be a total of seven transistors.

3. R.F. and I.F. Amplification

Problems arising in r.f. and i.f. amplification can best be studied with the aid of a small-signal equivalent circuit. Equivalent circuits which are reasonably accurate at high frequencies are given in Fig. 1 for the grounded-base connection and in Fig. 2 for the grounded-emitter connection.

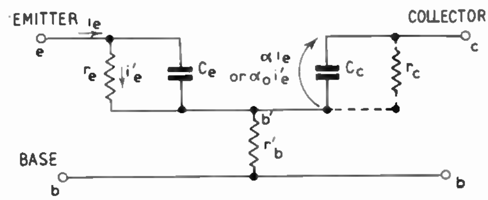

Fig. 1. High-frequency grounded-base equivalent circuit.

In Fig. 1, r_e is the dynamic resistance of the emitter junction (approximately $26/I_e$ ohms, where $I_e=$ d.c. emitter current in mA) and C_e is an equivalent capacitance which results mainly from the storage of minority carriers in the base of the transistor. α is the current transfer factor of the transistor and C_c and r_c are the equivalent capacitance and resistance of the collector junction. r'_b is the bulk semiconductor resistance which lies between the base contact and the active part of the transistor. Usually $r_c \gg 1/\omega C_c$ at high frequencies so that r_c can be ignored. The portion of the high-frequency emitter current which flows in C_e is a minority-carrier charging current which is not available for augmenting the collector current, while the remainder i'_e which flows in r_e can be regarded as the useful component of the emitter current. Neglecting transit time dispersion, a constant fraction of α_0 of this current reaches the collector junction independent of frequency, so that the active current generator has a strength $\alpha_0 i'_e$. Alternatively the current generator can be referred to the total emitter current and written as αi_e. It may then be seen that α and α_0 are related by the expression

$$\alpha = \frac{\alpha_0}{1+j\omega C_e r_e}$$

$$= \frac{\alpha_0}{1+j\omega/\omega_a} \text{ where } \omega_\alpha = 1/C_e r_e.$$

Here ω_α is the α cut-off angular frequency as conventionally defined.

Since $i'_e = v_{eb}'/r_e$ the current generator may alternatively be labelled $g_m v_{eb}'$ where $g_m = \alpha_0/r_e$. g_m is also frequency-independent.

The parameter g_m is used popularly in the grounded-emitter equivalent circuit of Fig. 2 which is derived from Fig. 1 by a straightforward transformation. Here the active

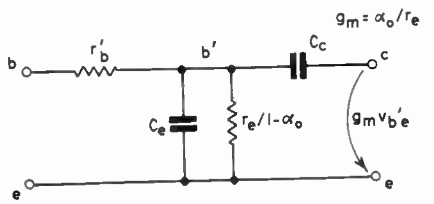

Fig. 2. High-frequency grounded-emitter equivalent circuit.

current generator has been placed across the output terminals rather than across C_e. These equivalent circuits show that r'_b is a detrimental factor which reduces the voltage applied to the active portion of the transistor and also gives rise to feedback in the grounded-base connection The collector capacitance C_c also contributes to feedback in both connections. Consequently low values of C_c and r'_b are desirable in a high-frequency amplifying transistor. The interelectrode feedback may cause instability in a tuned transistor amplifier, owing to the input resistance going negative at a frequency just below the resonant frequency of the collector tuned circuit. With most transistors, parameter values are such that the feedback is more objectionable in the grounded-base connection than in the grounded-emitter connection where the resistance r'_b has a stabilizing influence on the input resistance.

In general the power gain of a narrow-band tuned transistor amplifier commences to fall away at frequencies well below the α cut-off frequency. A typical characteristic for the grounded-emitter connection is shown in Fig. 3.

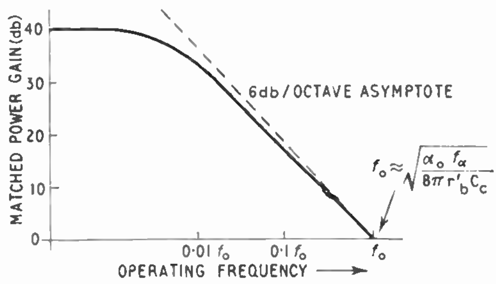

Fig. 3. Typical gain-frequency characteristic for tuned grounded-emitter amplifier.

It is seen that the power gain falls off gradually in the middle frequency range and then more steeply as the high-frequency amplifying limit is approached. As shown in Appendix 1 this limiting frequency, i.e. the frequency at which the available power gain is unity, is given to a good approximation by the expression

$$f_0 = \sqrt{\left(\frac{\alpha_0 f_\alpha}{8\pi r'_b C_c}\right)}$$

This expression shows that, from the power gain view-point, low values of r'_b and C_c are equally as important as a high α cut-off frequency in a radio-frequency transistor. Transistors designed for operation in the broadcast band are currently being made with α cut-off frequencies of 5 Mc/s or more and values of r'_b and C_c of the order of 100 ohms and 15 pF respectively.

The basic circuit of a grounded-emitter i.f. amplifier is shown in Fig. 4 (*a*). Single-tuned interstage transformers are shown here but double-tuned transformers, with the secondary suitably tapped to match the transistor input impedance, may be used to obtain greater selectivity. Single-tuned transformers can, however, be made much more compact than double-tuned ones, which is an advantage in transistor receivers. It is worthy of note, also, that two single-tuned stages, with their resonances staggered by amounts $\pm f_0/2Q$ from the central frequency f_0, can provide exactly the same selectivity characteristic as a double-tuned stage with the same values of L, C and Q. However stagger tuning may not be generally acceptable because of alignment problems.

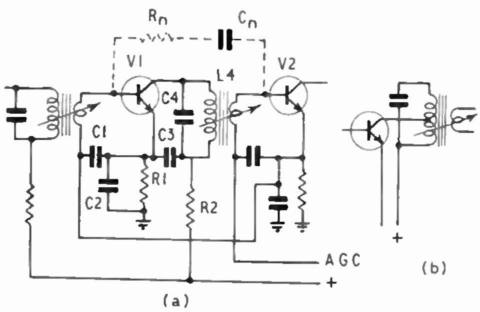

Fig. 4. Grounded emitter i.f. amplifier circuit.

The value of tuning capacitance required with the arrangement of Fig. 4 (*a*) may be inconveniently large, in which case it may be preferable to add turns to the collector winding as shown in Fig. 4 (*b*) so that a smaller tuning capacitor may be used. In order to minimize coupling losses in the interstage transformers the unloaded Q of the tuned circuits (Q_0) should be much larger than loaded Q (Q_L)*. The coupling loss for various values of Q_0/Q_L is shown in Table 1.

Table 1

Q_0/Q_L	Coupling Loss in db $= 10 \log_{10} (1 - Q_L/Q_0)$
5	1
2·7	2
2	3

Since the value of Q_0 will often be limited by the capabilities of miniature coils, the design of the coupling transformers may involve a compromise between selectivity and coupling loss. Consider, for example, the problem of providing selectivity in a transistor receiver comparable to that of a typical valve receiver which has 40 db attenuation at 10 kc/s off resonance. The transistor receiver may be expected to have three i.f. transformers, and calculations show that three single-tuned transformers having $Q_L = 100$ or three double-tuned transformers having $Q_L = 68$ would provide the necessary 40 db attenuation at 10 kc/s off resonance. Since values of Q_0 greater than 180 are difficult to achieve with miniature transformers the adoption of single-tuned transformers with $Q_L = 100$ would involve prohibitively large coupling losses and the use of double tuned transformers would appear to be advisable. Alternatively if the intermediate frequency were lowered to say 250 kc/s the required selectivity could be obtained with much lower Q values. While a lower-than-standard intermediate frequency would also give slightly higher power gain with present transistors its advantages appear to be offset by other troubles such as image-frequency interference.

Although the selectivity of present-day valve receivers guarantees a high degree of adjacent-channel rejection this is achieved at the expense of rather severe sideband cutting which results in poor audio fidelity. It is possible that a more reasonable compromise between selectivity and audio fidelity can be adopted in transistor receivers which will ease the requirements placed on the Q of the coupling transformers.

Other points to be noted in Fig. 4 are that a.g.c. is obtained by varying the d.c. base bias voltage which in turn varies the emitter current from cut-off up to a chosen maximum value. Where more than one i.f. stage is controlled, as in Fig. 4 (*a*), it is preferable to supply the base voltage of V1 from the emitter of V2 in order to minimize the d.c. loading on the detector. Bypass capacitors for the collector and base supplies should be returned to the emitter rather than to ground in order to provide the most direct return path for the signal currents circulating within each stage. The input impedance R_i of i.f. transistors at $I_e = 1$ mA varies considerably at present and may be anywhere in the range 200 to 500 ohms while the output impedance R_o may vary between 10,000 and 20,000 ohms. The interstage transformers should therefore be designed to suit the specific transistors used. Both R_i and R_o are accompanied by appreciable shunt capaci-

* This problem is not ordinarily met in valve i.f. amplifiers where the coupling transformers are permitted to absorb virtually all of the signal power.

tances which become incorporated into the tuning capacitances. It should also be noted that with correct matching, the shunt resistance across the collector-tuned circuit is $R_o/2$ and the required value of tuning reactance referred to the collector will be $R_o/2Q_L$. In gain-controlled stages the input and output impedances tend to rise considerably at low values of emitter current. This will cause pronounced variations in Q_L unless the initial ratio Q_0/Q_L is low. The latter point may be taken as an argument in favour of inefficient tuned circuits in gain-controlled stages.

Neutralization may also be necessary in high gain i.f. stages. This may be accomplished by connecting a resistance R_n and capacitance C_n in series between the bases of successive transistors, provided the interstage transformers are connected to obtain a phase reversal between primary and secondary voltages. As shown in Appendix 2 and verified by actual measurement, the value of C_n is nearly independent of emitter current but the required value of R_n decreases with emitter current. In practice satisfactory neutralization should be obtained if R_n is omitted.

4. Frequency Changing

Frequency changing may be carried out with separate local oscillator and mixer transistors, or the two functions may be combined in one transistor. An alternative possibility is to use a crystal diode as mixer, but owing to the conversion loss in the diode, an additional stage of i.f. amplification will usually be required.

Circuits which are representative of two-transistor and single-transistor mixers are shown in Fig. 5 (*a*) and (*b*). The single transistor circuit gives virtually the same conversion gain as the two-transistor circuit and should prove more popular on economy grounds. It does, however, exhibit a certain amount of reaction between the aerial and local oscillator tuning.

An extensive analysis of transistor mixers has been carried out by Zawels[7] who finds that optimum conversion gain is achieved by operating the mixer transistor with an injected local oscillator signal of 0·2–0·3 volts and a d.c. emitter current of 0·2 to 0·3 mA. In Fig. 5 (*a*), a voltage of this order of magnitude is generated at the base of the oscillator transistor and is injected into the emitter of the mixer transistor. The resistor R1 is here adjusted so that the rectified emitter current has the required value.

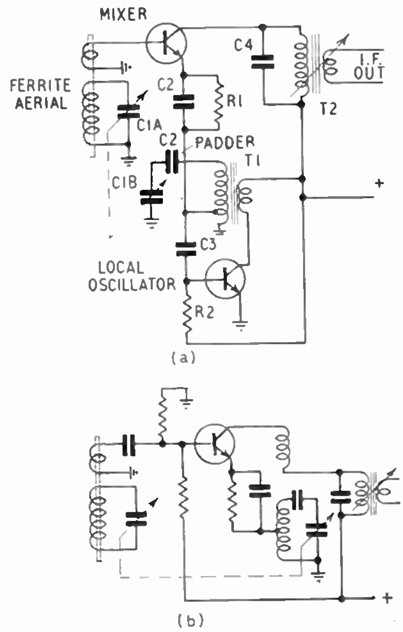

Fig. 5. Local oscillator-mixer circuits.

The transformer T1 is designed to provide an approximate impedance match between base and collector of the oscillator transistor and at the same time the transistor is tapped well down on the tuned circuit so that transistor capacitances have a minor effect on the oscillation frequency. The base tap should not be too low down on the tuned circuit, however, since the circuit losses may load the transistor so heavily that oscillations may not be maintained over the whole tuning range. These remarks apply particularly where the oscillator transistor has an inherent oscillation limit not far above the broadcast band. A practical design is discussed in a later section.

In the circuit of Fig. 5 (*b*) oscillations occur by virtue of the feedback from collector to emitter of the mixer transistor. The signal is injected into the base and the resultant i.f. component of the collector current generates an i.f. voltage across the tuned i.f. transformer.

In both Fig. 5 (*a*) and (*b*) optimum transfer of power from the ferrite rod aerial takes place when the impedance looking back into the coupling winding is equal to the mixer input impedance. This means, in effect, that the aerial is loaded down to a *Q* of half its unloaded value. Conditions for optimum signal-to-noise ratio have not been investigated yet but probably require that the loading on the aerial should be somewhat lighter than for optimum power transfer.

5. The Second Detector

Although a diode can be used as second detector it is preferable to use a transistor since automatic gain control of the i.f. stages requires an appreciable amount of d.c. power and this can be obtained from a transistor detector with a much lower i.f. power input than a diode requires. For reasonably distortion-free detection a signal input of 0·2 volts or more is required. Below this level the detector rapidly assumes a square law characteristic. A basic grounded-emitter detector circuit is shown in Fig. 6 (*a*). For best low-level action the emitter diode is forward-biased by a potential in the range 0·1–0·15 volt obtained from the divider R1 and R2. The rectified a.c. and d.c. components of the emitter current are amplified in the collector circuit and the i.f. fluctuations are filtered out by means of C2 which at the same time is made small enough to minimize the loss of high audio-frequency modulation components. An a.g.c. voltage is obtained from the collector after filtering out audio frequency voltages by means of R4 and C3. Transformer coupling could be used to improve the match between the detector and the following audio stage but resistance-capacitance coupling is simpler and cheaper although sacrificing some power gain.

Owing to the use of constant voltage bias in this circuit the no-signal emitter current is very temperature-sensitive, and as the collector current varies in close step with the emitter current the no-signal a.g.c. voltage may fluctuate to an undesirable extent. Furthermore with constant voltage bias the no-signal current is liable to vary considerably with changes of transistor. By substituting a junction diode or negative-temperature-coefficient resistor for R2 the bias can be compensated for temperature variations, but this method does not affect the variations between transistors.

An improved circuit which gives an approximation to constant current bias is shown in Fig. 6 (*b*). Here the voltage drop across R2 is made several times greater than emitter-to-base voltage. The resistance R5 then defines the emitter current fairly closely since the voltage drop across R5 is approximately equal to the drop across R2. If in this circuit C4 is made only large enough for i.f. by-passing, the linearity of detection of strong signals is greatly improved because the presence of R5 makes the emitter-base circuit act like a conventional linear diode detector. The overall gain is, however, very much reduced because of the degenerative action of R5. If, on the other hand, C4 is made large enough to bypass audio frequencies the full sensitivity relative to modulation power is restored, although the linearity becomes poor once more.

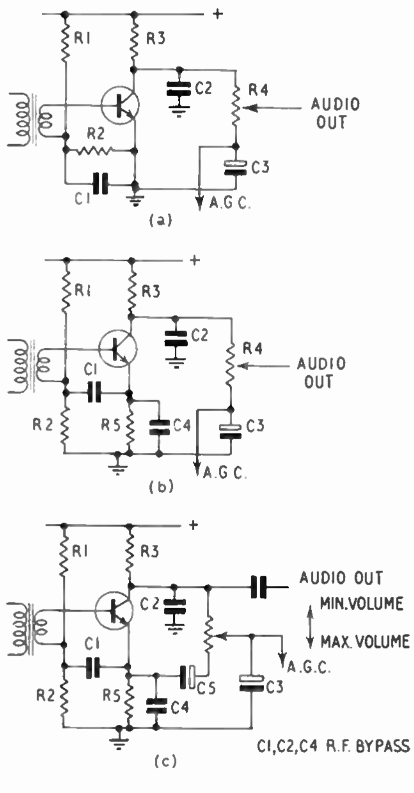

Fig. 6. Second detector circuits.

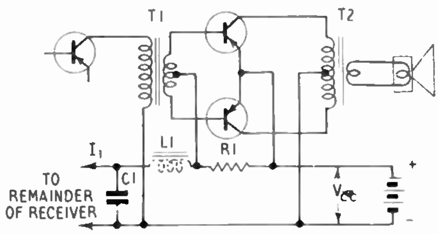

Fig. 7. Audio output circuit.

Normally good quality is not expected in the reception of weak signals and sensitivity is paramount, but with strong signals quality is paramount and some loss of sensitivity can obviously be tolerated. The circuit of Fig. 6 (*c*) has been devised to combine both modes of detection. Here the volume control R4 places a large bypass C5 across C4 when at its maximum setting. Towards the minimum setting of R4 only the i.f. bypass C4 is effective across R5 and the audio output is progressively reduced by the shunting effect of the top portion of R4. This circuit shows a small amount of "break through" at minimum volume, owing to the finite impedance of C3. C3 should therefore be a large capacitance to minimize this effect as well as to prevent variations in the low frequency response at low volume levels. More elaborate volume control circuits, along the lines suggested by Shea[8], can be used to circumvent these effects.

6. Audio Amplification

The most important question to be decided about the audio end of the receiver is whether to use a class A or a class B output stage. A class A output stage undoubtedly poses fewer design problems than a class B stage and is cheaper in first cost since it can be single-ended, whereas a class B stage must, of course, be push-pull. While a class A stage is capable of an efficiency approaching 50 per cent. at full output it draws a steady power at all volume levels of not less than twice the desired maximum output power. By contrast, the class B stage draws very little idling power and gives an efficiency approaching the theoretical maximum value of 78 per cent. at full output.

The design of transistor audio amplifiers has been extensively treated in the literature (see, for example, Shea[8]), and it will suffice here to treat briefly some of the more important aspects of the design of the output stage. Fig. 7 shows a typical push-pull output circuit. Here the output transformer is designed to present a load of $2(V_{cc}-V_M)^2/P_o$ ohms between collectors, where V_{cc} is the collector supply voltage, V_M is the minimum instantaneous collector voltage and P_o is the desired output power. In practice, V_M is usually only a fraction of volt so that if V_{cc} is 6 volts or more little error is introduced by neglecting V_M.

For a given value of load resistance the transfer and the input characteristic of each transistor will be of the general form shown in Fig. 8 (*a*) and (*b*). These curves commence with a region where the collector and base currents are building up exponentially from a very small initial value. Next comes a more or less linear region, after which the collector

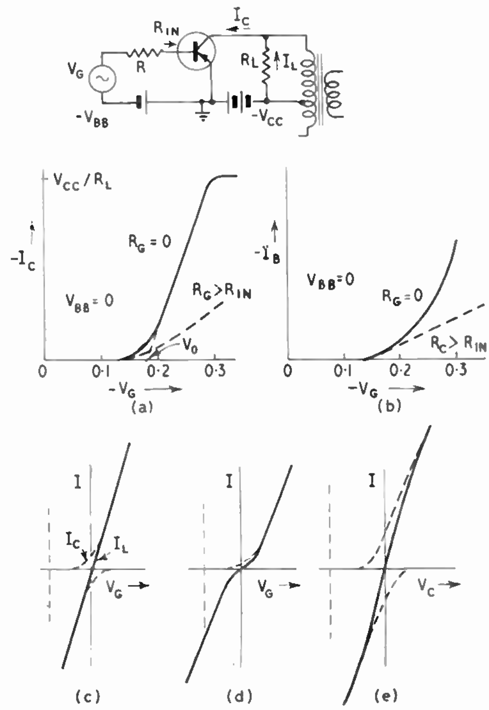

Fig. 8. Audio output circuit characteristics.

(*a*) Transfer characteristic for one transistor.

(*b*) Input characteristic for one transistor.

(*c*) Push-pull transfer characteristic with the correct value of V_{BB}.

(*d*) Same as for (*c*) but with V_{BB} too small.

(*e*) Same as for (*c*) but with V_{BB} too large.

current saturates, owing to "bottoming" of the collector.

Push-pull transfer characteristics may be constructed by methods very similar to those used in analysing valve amplifiers and are shown in Fig. 8 (*c*), (*d*) and (*e*) for three values of the base bias voltage V_{BB}. For values of V_{BB} less than the "extrapolated cut-off" bias shown as V_0 in Fig. 8 (*a*) the push-pull transfer characteristic shows a discontinuity at the centre which gives rise to a step in the output waveform. Optimum linearity results when $V_{BB}=V_0$. Values of $V_{BB}>V_0$ may give rise to a certain amount of distortion at moderate power levels, owing to an inflection in the transfer characteristic, and at the same time the idling current will be unnecessarily high.

The linearity of the individual transfer characteristics at low values of I_c can generally be improved by making the driver impedance relatively high as shown by the broken line in Fig. 8 (*a*). The idling current for the extrapolated cut-off bias is then considerably reduced. It should be noted that to maintain a high driving impedance at low frequencies the primary inductance of the driving transformer must be greater than required for maintaining frequency response alone. In practice there is little to be gained by adopting a driving impedance greater than two or three times the averaged transistor input impedance.

The optimum bias may lie anywhere in the range 0·1 to 0·2 volts at room temperature and has an appreciable negative temperature coefficient. Temperature compensation should be provided where the ambient is likely to vary more than $\pm 10°$ C.

An economical way of providing the required bias using a "back-bias" resistor (R1) is shown in Fig. 7. This method is satisfactory so long as the current drawn by the remainder of the receiver (I_1) is reasonably constant. Alternatively a separate voltage divider can be connected across the battery where the extra drain is acceptable. Temperature compensation can be provided by using a suitable negative-temperature-coefficient (NTC) bias resistor or by the use of a junction diode connected so that the current passes in the low resistance direction.

Care must be taken in applying the back-bias method if the peak base current of the output transistors becomes comparable to I_1. The nett current through R1 will then fall to a low value and if R1 is a junction diode it will tend to "cut-off." Severe distortion will then result. However if R1 is a linear resistance and is decoupled from the reminder of the receiver by the inductance L1 the bias will be correct at the cross-over instant since the instantaneous base current is then zero. The resistance R1 is then, in effect, added to the input impedance of the output transistors. However R1 will typically be about 40 ohms for $I_1=4$ mA, and since R_{in} is typically 100 to 200 ohms the increase in input impedance is not serious. Where the base current of the output transistors at full output is small compared to I_1, L1 can be omitted and R1 can be bypassed.

The design of the output and driving transformers follows principles well-established in class B valve amplifiers. However the loading on the driving transformer is much more constant in a transistor amplifier than in a valve amplifier. For best audio quality the leakage reactance between the two half secondaries of the driving transformer and between the two half primaries of the output transformer should be kept to a minimum, otherwise objectionable cross-over transients may occur. This problem is easily solved by the use of bifilar windings. In small portable sets, however, acceptable quality can be obtained with plain transformer windings.

Where the load impedance is low enough to permit direct operation into a loudspeaker voice coil and matched pairs of *pnp* and *npn* transistors are available the output transformer can be eliminated by the use of a complementary symmetry circuit.[9] Alternatively a centre-tapped voice coil would be very convenient for use with *pnp* transistors.

7. Experimental Receiver

An experimental receiver was designed along the line indicated in the preceding sections and its circuit is shown in Fig. 9. The set uses five selected RD2517A *npn* transistors (Germanium Products Corp. U.S.A.) in the mixer-oscillator, i.f., and detector stages. Although not specifically designed for such application these transistors happened to be the best on hand at the time of building the set. Several other American transistor types giving excellent performance in r.f. stages are now available and there are also similar types available from European sources.

The aerial rod is of 9 mm diameter grade 4B ferroxcube cut to a length of 7½ inches. It has a tuned winding consisting of 115 turns of 7/41 litz wire space-wound on a thin sleeve 3½ inches long. The sleeve is movable for fine adjustment of its inductance. A coupling winding of 5 turns, also on a movable sleeve, is wound at one end of the primary. The unloaded Q of the aerial rod when mounted adjacent to the chassis is approximately 160 and this value is further reduced to approximately 80 when the coupling winding is adjusted for optimum transfer of power to the mixer. Experimentally it was found that matching was complicated by a large variation of the input impedance of the mixer transistor over the broadcast band. In a typical case the input impedance was found to vary from 5000 ohms at 550 kc/s to 600 ohms at 1500 kc/s. As no theoretical or practical reason for this behaviour could be discovered in a reasonable time the matching was made correct at 1 Mc/s and the mismatch at other frequencies was accepted.

It was found that the broadcast band could be covered with a tuning capacitance of 5 to 110 pF—a figure which is made possible by the low value of effective capacitance coupled into the tuned circuits by the mixer transistor. However it is now felt that a more practical value of the tuning capacitance would be in the range 150 to 200 pF max. This would still be quite small physically and would provide more tolerance to accidental stray capacitances. The two-gang tuning capacitor was made from a standard midget single-gang capacitor with some of its rotor and stator plates omitted. The stator was then separated into two insulated sections.

The oscillator and i.f. transformers were wound on ferroxcube pot cores type D14/8 grade 4B, having an adjustable tuning slug. An air gap of 0·015 inch was ground in the centre post of these cores. This was found to give approximately the optimum Q at 455 kc/s. The oscillator transformer was wound with a tuned winding of 78 turns of No. 38 B & S enamel and single silk-covered wire tapped at four turns from the bottom for the emitter connection. The collector winding consisted of 18 turns of No. 38 wire. It was found advisable to use layer insulation of thin polythene tape in order to minimize the self-capacitance of the coil and improve its Q at 2 Mc/s. It is probable that a pie-wound, self-supporting coil would be preferable to the present coil which is wound on the bobbin provided with the ferroxcube core.

The i.f. transformers T2, T3 and T4 were designed for an impedance ratio of 10,000 ohms to 300 ohms. The primaries have 45 turns of 9/41 litz wire and the secondaries 8 turns. The unloaded Q is 180. Owing to a design oversight the adopted tuning capacitance of 1000 pF leads to an undesirably low value of

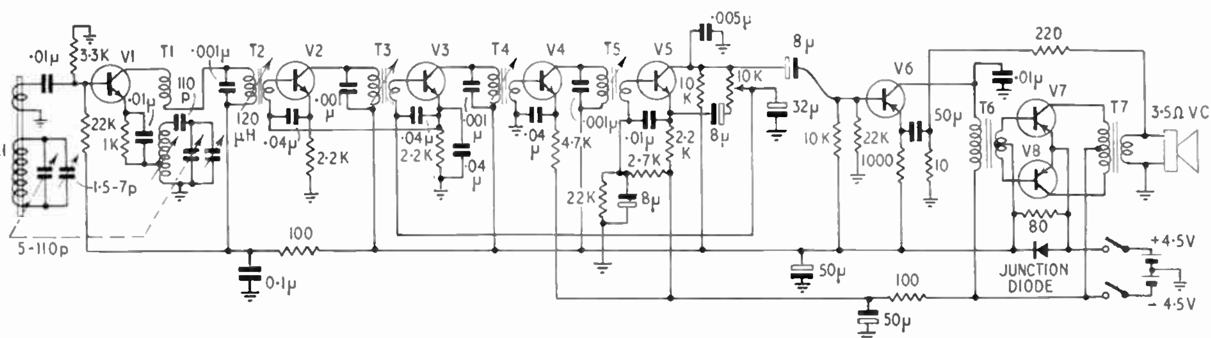

Fig. 9. Experimental receiver circuit.

V1–V5 are Germanium Products Corp., U.S.A., transistors type RD2517A (selected) or RD2521A. V6 is OC71. V7 and V8 are experimental transistors (alternatively OC72 or 3X/300N.)

the loaded Q (about 15). This should preferably be increased to 40–50 by an increase of the C/L ratio. Owing to the higher input impedance of the detector (about 600 ohms) a 12-turn secondary was used on the transformer T5. Neutralization of the i.f. transistors was found to be unnecessary owing to the relatively low stage gain.

The detector was designed along the lines described earlier. It operates with a standing emitter and collector current of 0·2 mA. As the midpoint of the 9 volt supply battery is earthed the collector of V5 stands at approximately 2·5 volts above earth. This establishes a current of approximately 1 mA in the i.f. transistors V2 and V3. When a strong signal is tuned in the collector voltage falls until it reaches a value of a few tenths of a volt above earth at which point V2 and V3 are nearly cut off. The third i.f. transistor V4 is operated with a constant emitter current of 1 mA defined by the 4,700 ohm series emitter resistance. Adequate stabilization and linearization of the detector is achieved through the use of a 2,200 ohm series emitter resistance. Higher values than this were found to lead to overloading of the third i.f. stage on strong signals. The measured percentage of second harmonic distortion in the detector output for a 30 per cent. modulated i.f. signal was 7 per cent. for weak signals, decreasing as the signal strength increased to an amount indistinguishable from that already present in the modulation envelope of the signal generator (about 1·5 per cent.).

In the audio output stage the opportunity was taken to test some experimental transistors which, although operating well below their maximum power rating, had the advantage of a relatively low input resistance. Comparable results would be obtained from the type OC72 or 3X/300–302 N. The driving transistor is an OC71 operating in a stabilized bias circuit at an emitter current of 1·5 mA. A turns ratio of 5 to 1 overall was adopted for the driving transformer while the output transformer was designed to present a load of 400 ohms from collector to collector.

In the absence of a suitable NTC resistor a junction diode was used to obtain temperature compensation bias for the output transistors. Some difficulty was experienced with reduction of current in the diode at high output levels and improved performance was obtained at the expense of some loss of temperature compensation by shunting the diode with an 80 ohm resistor. About 8 db of negative feedback was applied over the audio stages by means of a resistive divider connected between the voice coil and the emitter of V6.

Table 2

Comparative Running Costs of Battery-Operated Valve and Transistor Receivers

	Typical 4-valve portable	Transistor Receiver Class B output	Transistor Receiver Class A output
Power output (mW)	250	250	250
"A" battery consumption (mW)	375 (1·5 V at 250 mA)	—	—
"B" battery consumption (mW)	1215 (90 V at 13·5 mA)	90 (9 V at 10 mA)	540 (9 V at 60 mA)
"A" battery cost (shillings)	21/-	—	—
"B" battery cost (shillings)	18/-	6/9 (6 size-"D" cells at 1/1½d.)	approx. 21/-
Battery life at 2 hr/day (hr)	250	500	250
Battery cost pence per hour	1·9	0·16	1

When tuned to a strong signal the total battery drain is 6 mA at zero volume. At full volume the measured audio power delivered to a dummy load is 300 milliwatts with 10 per cent. 3rd harmonic distortion for a drain of 60 mA. When playing speech or music at a moderate volume level the average drain is approximately 10 mA.

The sensitivity for 50 milliwatts output is 300 microvolts per meter at 1400 kc/s falling to 600 microvolts per metre at 550 kc/s. This falling off is apparently due to the aerial mis-matching at the lower frequency end of the band. Equivalent noise sideband input is 22 microvolts per metre. However no attempt has been made to secure the optimum noise performance. Owing to the low operating Q of the i.f. transformer mentioned earlier the adjacent channel attenuation is only 10 db. The response of the audio section is 3 db down at 80 c/s and 4000 c/s. Under ambient temperatures varying between 0 and 50° C the zero-volume battery current varies from 5 mA to 7 mA and the no signal a.g.c. voltage varies from 1·5 volts to 2·5 volts. This performance is probably adequate for Australian conditions. The performance of the receiver towards the end of the useful battery life has not been measured.

The set as shown uses two flat 4½ volt cells which are estimated to have a life of 200 hours at an average drain of 10 mA for 2 hours per day. Better economy would be obtained from the use of six size-D torch cells which would have a life of 500 hours under the same conditions. In Table 2 a comparison is made between the running costs of a typical full-size 4-valve portable radio and a transistor radio. In the third column the running cost of a hypothetical transistor radio having a 250 mW class A output stage is shown to emphasize the advantages of class B operation. It may be remarked in passing that a running cost of 0·16 pence per hour is not much more than the running cost of a typical small mains-powered radio which at, say 50 watts consumption, costs approximately 0·12 pence per hour to run.

8. Acknowledgment

The author is indebted to Mr. C. D. Hogarth for constructing the experimental radio described herein and to Dr. L. W. Davies for providing experimental power transistors.

9. References

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10. Appendix 1: Approximate High Frequency Power Gain of a Grounded-emitter Tuned Amplifier

In using the equivalent circuit of Fig. 2 to analyse the operation of a tuned grounded-emitter amplifier in the neighbourhood of the α cut-off frequency, certain simplifying approximations are permissible. At the α cut-off frequency we have $1/\omega_a C_e = r_e \cong 25$ ohms.

Since r'_b is typically 100 to 200 ohms the input impedance is approximately resistive and equal to r'_b.

Furthermore the resistance $r_e/(1-\alpha_0)$ appearing across C_e is typically of the order of

500 ohms and can be ignored. Since $C_c \ll C_e$ the output impedance consists of a capacitance C_c shunted by an impedance resulting from the activation of the output current generator by feedback from the point c to the point b′. It may be seen that if unit voltage is applied to the output terminals the active current generator draws an in-phase current of approximate magnitude $g_m C_c/C_e$. Hence the equivalent output resistance appearing across C_c has a magnitude $C_e/g_m C_c$.

Under operating conditions the capacitance C_c is tuned out and for maximum power gain we have:

Generator resistance $r_g = r'_b$

Load resistance $r_l = C_e/g_m C_c = 1/\alpha_0 \omega_\alpha C_c$.

For an internal generator voltage v_g we obtain

$$v_{b'e} \cong v_g/2j\omega C_e r'_b$$

and $v_{out} = g_m v_{b'e} r_l/2$.

The power gain is given by

$$\left|\frac{v_{out}}{v_g}\right|^2 \times \frac{4r_g}{r_l}$$

This reduces after manipulation to

$$\frac{\alpha_0 \omega_a}{4\omega^2 C_c r'_b}$$

Alternatively we may write

$$\text{Power gain} = f_0^2/f^2$$

$$\text{where } f_0 = \sqrt{\left(\frac{\alpha_0 f_a}{8\pi r'_b C_c}\right)}$$

It should be emphasized that this analysis gives only an approximate indication of the performance of a transistor at the upper end of its frequency range.

11. Appendix 2: Neutralization of a Grounded-emitter Tuned-Amplifier

Referring to Fig. 4 (*a*) exact neutralization is obtained when the current fed back through the neutralizing network is equal and opposite to the current fed back through the transistor. Allowing for a turns ratio of n in the interstage transformer this is equivalent to saying that the neutralizing admittance must be equal to n times the short-circuit feedback admittance of the i.f. transistor.

Using the equivalent circuit of Fig. 2 it may be shown that the feedback admittance with the input short circuited is given by

$$y_{bc} = \frac{j\omega C_c}{[1+(1-\alpha_0)r'_b/r_e + j\omega C_e r'_b]}$$

This admittance corresponds to a resistance $r'_b C_e/C_c$ in series with a capacitance

$$C_c/[1+(1-\alpha_0)r'_b/r_e]$$

Hence for exact neutralization we have

$$R_n = r'_b C_e/nC_c$$

and $$C_n = nC_c/[1+(1-\alpha_0)r'_b/r_e].$$

As the term $(1-\alpha_0)r'_b/r_e$ is generally much smaller than unity the value of C_n is fairly close to nC_c. However as $C_e(=1\omega_a r_e)$ decreases with emitter current the required value of R_n also decreases with emitter current.

HEINRICH HERTZ

HEINRICH HERTZ, whose name is commemorated in the second Institution Premium, was born 100 years ago this month, on the 28th February 1857. Clerk Maxwell, whose name has been given to the senior award, is always regarded as the founder of modern electromagnetic theory and similarly Heinrich Hertz was the pioneer of electromagnetic radiation experiment. Indeed to a very large extent the general acceptance of Maxwell's theory was due to the work of Hertz.

Hertz was born in Hamburg and while still at school constructed optical and mechanical instruments as a hobby. On leaving school, he intended to become an engineer, but whilst studying became more and more interested in pure science. He eventually decided to take up an academic career and in 1878 went to work at Berlin in Helmholtz's laboratory. He soon became his assistant and carried out much experimental work in electricity. In 1883 he went to Kiel University where he had received an appointment as *privatdozent* ("private lecturership") and began a serious study of Maxwell's theory which had been published in book form some ten years earlier.

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His first publication was a justification of Maxwell's equations on theoretical grounds. Continental theory was based on the concept of direct action at a distance whereas Maxwell's theory involved the motion of a displacement current in a dielectric medium. Hence Hertz's paper attracted some interest and on his appointment in 1885 as Professor of Physics at Karlsruhe he began to seek direct experimental evidence for the Maxwellian theory.

In his early experiments he used the spark discharge of an induction coil through what we would now call a tuned circuit and detected the effect in a similar piece of wire bent into a rectangle with a short air-gap between the ends of the wire. Using this elementary transmitter and receiver, which operated on a wavelength of about 24 cm, he carried out interference experiments by transmitting waves along a wire and through the air simultaneously. He thus proved that the electromagnetic wave was propagated with a finite velocity and he subsequently showed that this was of the same order as the velocity of light. Other experiments demonstrated the similarity between electric waves and light waves since he was able to reflect them and also show refraction and polarization effects.

This particular line of work was extended some ten years later by Sir J. C. Bose, the Indian scientist, after whom another Institution Premium is named.

It is strange that during his next appointment, that of Professor of Physics at Bonn, where he went in 1889, Hertz transferred his interests completely to the fundamental principles of mechanics and did no further work on electromagnetic waves, on which his claim to fame rests. Indeed, high frequency oscillations are often called "hertzian-waves" and the unit of frequency is similarly known as the hertz.

Hertz died on 1st January, 1894, just before his 37th birthday.

APPLICANTS FOR ELECTION AND TRANSFER

As a result of its February meeting the Membership Committee recommended the following elections and transfers to the Council. This list also includes studentship registrations which were considered at the meeting held in January.

In accordance with a resolution of Council and in the absence of any objections, the election or transfer of the candidates to the class indicated will be confirmed fourteen days after the date of circulation of this list. Any objections or communications concerning these elections should be addressed to the General Secretary for submission to the Council.

Direct Election to Member
CLARK. John Eric. *Leigh-on-Sea.*

Transfer from Associate Member to Member
LAMPITT. Robert Alfred. *Wolverhampton.*
PILGRIM. Commander Kenneth Arthur William. R.N. *Gosport.*

Direct Election to Associate Member
HATCHER. Walter Douglas. B.Sc.(Eng.) *London. S.E.19.*
JONES. Flt. Lt. Frank. Dip.El.. R.A.F. *Wilmslow.*
NORMINGTON. Alfred Charles. B.Sc.(Eng.). *Manchester.*
PHILLIPS. Glyn Jones. B.Sc. *Cardiff.*
RUDKIN. Alfred Charles Frederick. *Walton-on-Thames.*

Transfer from Associate to Associate Member
BUDDEN. George Fothergill. *Altrincham.*

Transfer from Graduate to Associate Member
BRULEY. John. *Bushey.*
SITARAM. Racherla. M.Sc. *Chromepet. Madras.*
SMITH. Frederick George. *Cardiff.*

Direct Election to Associate
BRADFORD. Douglas Stephan. *Lark Hill.*
MAYHEAD. Lawrence Victor. *Kingston on Thames.*

Transfer from Student to Associate
ENDALL. John Richard. *Birmingham.*

Direct Election to Graduate
BELL. Ernest Clifford. B.Sc.(Eng.). *Bradford.*
BROWN. Andrew Duncan. *Orpington.*
COOPER. John Derek. *Sleaford.*
COWE. 2nd Lt. Dan Williamson. B.Sc.. R.E. *Irak.*
HANDS. Edward. B.Sc.(Eng.). *Birmingham.*
JACK. Capt. James McLay. B.Sc.. R.E.M.E. *B.A.O.R.*
MAHONEY. John. *Chelmsford.*
PARKS. John Ronald. *Isleworth.*
PINTO. Cyprian. *Kanpur.*
POEY. Patrick. *St. Margaret's Bay.*
POTTS. John Robert. *Rutherglen.*
RETTIE. Alister Brian. *Hatfield.*
ROSE. Milton. B.Sc. *London. N.W.9.*
SMITH. James G. *London. S.E.15.*
TUDOR. Ernest Rhodes. *Trumpington.*
WRAY. Robert William. B.Sc. *Sutton.*

Transfer from Associate to Graduate
BEVERIDGE. Lieut. Stuart. R.N. *Edinburgh.*
BOOTH. Charles Hector. *Greenford.*

Transfer from Student to Graduate
GILVARY. David Francis. *Coventry.*
HALLWORTH. Robert Philip. *London. E.6.*
RANGASWAMY. Setlur Venkataranga Iyengar. B.Sc. *Bangalore.*
SANDYS. Maurice Arthur. *London. W.14.*

STUDENTSHIP REGISTRATIONS

January 1957
AGRAWAL. Virendra Krishna. *Bombay.**
AHUJA. Arjan Dass. B.A. *New Delhi.*
CAMPBELL. Ian Malcolm. *Edinburgh.*
CHENNEOUR. Capt. Kenneth. R.E.M.E. *Arborfield.*
CHESTER. Michael William. *Beccles.*
DUGGAL. Didar Singh. B.A. *Agra.*
GEORGE. Prince Festus. *London. W.8.*
HAMILTON. Jonathan Olufunso. *London. W.2.*
HELPS. John David. *H.M. Forces.*
HOBBS. Thomas William. *H.M. Forces.*
IFIDON. Rowland Oke. *London. W.2.*
JASWANT SINGH. Ghoman S. *Gurdaspur.*
JONES. Eric. *Wrexham.*
KILGANNON. Peter. *Wyton. Hunts.*
McCALL. Robert Watt Sutton. *Ilford.*
NAWI. Khedaer A. *Tel-Aviv.*
NUNES. Edward Vincent. *Bombay.**
OM NARAYAN GARG. M.Sc. *Kanpur.*
PARAMALINGAM. Sivaguri. *Neerveley. Ceylon.*
PARKIN. Oswald Theodore. *Poole.*
POWER. Edward Plunkett. *Dublin.*
SAHATHEVAN. Sinniah. *Valvetty Valvettiturai. Ceylon.*
SALISBURY. Henry Nance. *Aberystwyth.*
SHAH. Rajendra Kantilal. B.Sc. *Bombay.*
SHARMA Kamal Kishore. *Bombay.*
SKINNER. John. *Bristol.*
SMITH. Harold Albert. *Salford.*
STEELE. Michael. *Hitchin.*

February 1957
BASSETT. Edward James. *Cheltenham.*
BASU. Amiyar Kumar. B.Sc. *Cuttack.*
CARREYETT. Trevor Walter. *Bristol.*
CHANDRASEKHARAN. Panthaloor. *Bangalore.*
COPPACK. Kenneth Norman. *Chester.*
EKAMBARAM. Conjeeveram. *Bombay.*
GELDARD. David Scott. *H.M.S. Keppel.*
GIFFORD. Plt. Off. Robert. R.A.F. *Leamington Spa.*
GRACE. Peter Brian. B.Sc. *Marlow.*
KADAM. Padmakar Gopal. *Dwarka.*
LAMA. Prem Kumar. *Agra.*
McKENZIE. John George. *London. W.12.*
MATTHEWS. Donald Charles Mackenzie. *Hemel Hempstead.*
NEWRICK. Roy William. *Beccles.*
PAPWORTH. George Geoffrey Guy. *London. W.2.*
REED. Christopher. *Wells.*
SMITH. Derek Travers. *H.M.S. Birmingham.*
TOWNLEY. Michael Edward. *St. Albans.*
WATERLANDER. Cornelis Johannes. *Ouyen. Victoria.*
WHITEHILL. William Kenneth. *Newport. Mon.*
ZAKI. Mohamad. B.E. *Newcastle-upon-Tyne.*

* Reinstatement.

AMPLITUDE MODULATION OF MICROWAVES BY TUNABLE TRANSMISSION WAVEGUIDE FILTERS

by

M. H. N. Potok, B.Sc., Ph.D. (Associate Member)†
and J. Barbour, A.R.T.C. (Hons.)‡

SUMMARY

Amplitude modulation of microwaves can be obtained by shifting the pass-band of a transmission filter by the modulating signals. A linear response and a bandwidth of over 4 kc/s at 9000 Mc/s has been obtained using simple components and circuits.

1. Introduction

Recent work on microwave generation has resulted in the development of a variety of oscillators ranging from the now more conventional klystron and magnetron to the latest backward travelling-wave tubes.[1]

All of these can be frequency modulated with comparative ease but, unlike the conventional valve oscillators, they cannot be amplitude modulated. This led to various attempts at amplitude modulation of the generated signal outside the generator.

The simplest form such a modulator can take is a rotating vane made of a lossy or reflecting material entering the waveguide through a central longitudinal slot. The shape of the envelope depends on the shape of the vane and this method could not be adapted easily to amplitude modulation by speech or music.

Instead of using a vane, a gas-discharge attenuator could be employed to vary the u.h.f. carrier attenuation in step with the modulating signal.[2] This appears to give satisfactory results up to a modulating frequency of 1 Mc/s.[3]

Ferrites also offer promising solutions of which at least two have been described. The more commonly known one utilizes the property of a ferrite to cause the rotation of the electromagnetic field associated with microwaves as they move past, the degree of rotation being determined by the magnetization of the ferrite.[4] If a non-lossy ferrite were placed in a circular guide terminated by a rectangular one designed to pass only the TE_{10} mode, the amplitude of the wave transmitted would depend on the degree of rotation of the axes of the electromagnetic field. Since the magnetization of the ferrite can be brought about by an audio-frequency signal, the output will be amplitude modulated, distortion being kept in check by suitable feedback arrangement.[5]

Another method is to use a lossy ferrite. It is found that the loss in some ferrites varies at microwave frequencies, depending on their magnetization. If such a ferrite were placed across an absorption cavity placed along the waveguide and tuned to the frequency of the transmitted wave, then changes in magnetization of the ferrite core brought about by an external coil, supplied by an audio-frequency signal, will produce changes in the absorption of power by the cavity, resulting in amplitude modulation of the output power.[6] Good results are claimed up to a modulating frequency of 1·3 Mc/s.

The absorption of power by the cavity could also be varied by changing its resonant frequency and this method has also been suggested for amplitude modulation of microwaves.[7]

The method described below, although thought out quite independently, seems to be a development of the last mentioned principle. Here the wave passes through a waveguide filter of the two-irises or two-posts type. Such a filter can be tuned within a few per cent. on either side of its resonant frequency by the

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insertion of a plunger at the centre of the filter[8] as in Fig. 1. This method, like the non-lossy ferrite but unlike the gas-discharge and absorption cavity methods, produces a reflected signal which is also amplitude modulated and can be made use of.

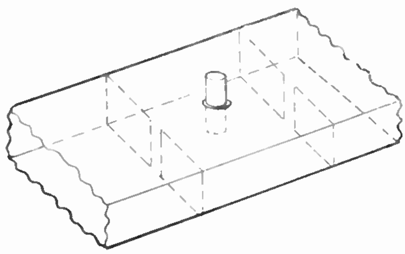


Fig. 1. Two-iris filter tuned by central post.

2. Characteristics of Tuned Filter Modulator

The method of modulation is best explained by reference to Fig. 2. The pass-band of the filter is shifted by varying the penetration of the plunger, actuated by audio-frequency signals via a speech-coil, etc. In its quiescent position, the filter is mistuned so that the carrier frequency corresponds to the steep portion of

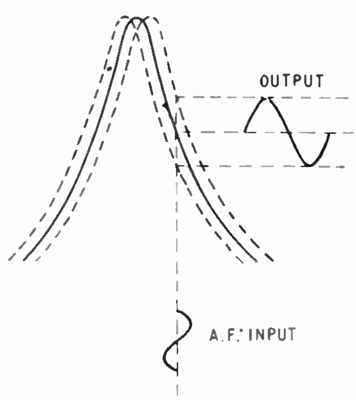


Fig. 2. Method of filter modulation.

the filter response characteristic, hence shift of the pass-band results in amplitude modulation of the output. It is clear that for linear modulation, the carrier frequency must lie on the linear portion of the filter response characteristics, hence if modulation index of the order of 80 per cent. is desired then almost half the available power will be reflected by the filter in its quiescent state. At power levels employed in microwave transmission this is not a serious loss. The reflected modulated wave can be isolated by a suitable directional coupler, and made use of, if required.

2.1. *Sensitivity and Efficiency*

The sensitivity of the device expressed as relative output variation for a unit movement of plunger within the linear region was found to increase—as expected—with the Q of the filter, being a function of the rate of fall of power output when filter is off tune. It also depends on the mean depth of penetration of the plunger since its reactance increases rapidly as its length inside the guide approaches a certain proportion of guide depth,[9] as seen in Fig. 3. Thus, for example, of two filters both

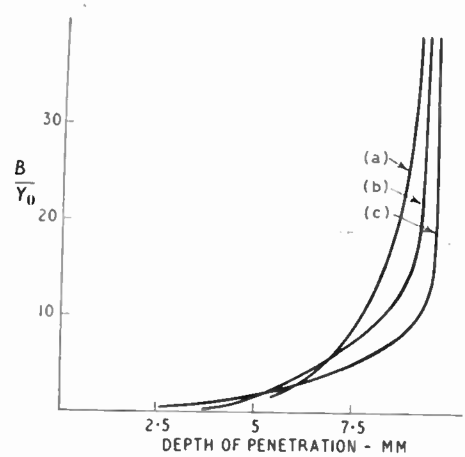


Fig. 3. Reactance of a post in X-band guide (WG 16) at 3·2 cm wavelength

(*a*) 1/16″ diameter, (*b*) 1/8″ diameter, (*c*) 1/4″ diameter.

tuned by the central plunger to 3·2 cm and having Q of about 500 one gave 80 per cent. modulation for a plunger movement of 0·12 mm at a mean penetration of 3·2 mm while the other gave the same modulation index for a plunger movement of 0·01 mm at a mean penetration of 5·6 mm. There are, however, some disadvantages to making the penetration too deep for the following reasons:

(*a*) A small gap between the moving plunger face and the opposite face of the guide may lead to a discharge should higher powers be used.

(*b*) The plunger is itself a tuned circuit absorbing more energy as its length within the guide approaches certain length. This energy is partly dissipated as heat and partly removed from the guide along the plunger. Very carefully designed chokes are required to reduce the latter loss.

(*c*) As the bandwidth of the filter is reduced (i.e. its Q is increased) and the plunger is made to penetrate considerable distance into the guide, it becomes increasingly important to place the plunger exactly half way between the two irises or posts, otherwise appreciable insertion loss may result.

The insertion loss due to slight inaccuracies and eccentricities inherent in the making of the modulator-filter unit and the reradiation loss via the plunger constitute an appreciable proportion of available power reducing the efficiency of this device well below the 50 per cent. mentioned in the previous paragraph.

3. Experiments on Filter Modulator

3.1. *Filter Characteristics*

In the design of the modulator it has been decided to make it as sensitive as was practicable to the movement of plunger but to make the filter of fairly low Q so as to reduce its sensitivity to frequency drifts which would result in amplitude changes. The eventual design employed a two-iris filter having Q of 300 at 9 kMc/s. Fig. 4 shows the relation between output and depth of penetration of the plunger. The absorption and radiation caused a somewhat unexpectedly large loss of power of 14 db, while further 4 db have been sacrificed to allow linear modulation up to a depth of 50 per cent. by tuning the filter to operate at a plunger penetration of 8·8 mm. At this setting a plunger movement of $\pm\cdot01$ mm resulted in 15 per cent. change in power output.

3.2. *Vibrator*

To convert voltage oscillations to plunger movement the Goodman Vibrator Model V47 was used. This vibrator can be used up to 10 kc/s, its impedance rising from 3 to 10 ohms over the range. A brass rod 0·07 in. diameter, 1 in. long, was attached to the vibrator to act as the plunger. The vibrator and filter were carefully aligned and clamped together to avoid any relative movement. The arrangement is shown in Fig. 5.

3.3. *Tuning*

The initial tuning was carried out by pulsing the klystron (CV 129) by a 2,000 c/s square wave applied to the grid and then adjusting the plunger penetration while a 300 c/s signal is applied to the vibrator. The purpose of pulsing is to be able to display the output obtained from a crystal on a cathode ray screen since the carrier cannot be so shown. The filter can be tuned visually now for optimum operation by observing the 300 c/s modulation envelope on the background of 2,000 c/s square wave.

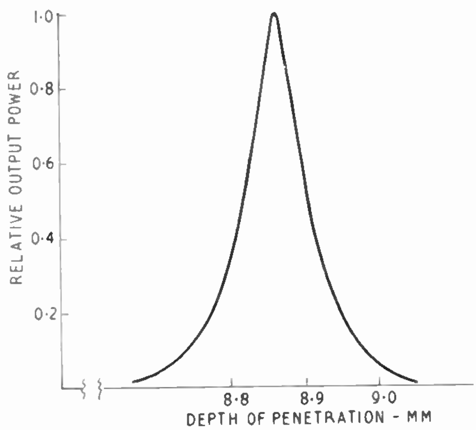


Fig. 4. Filter output versus depth of penetration of plunger in filter.

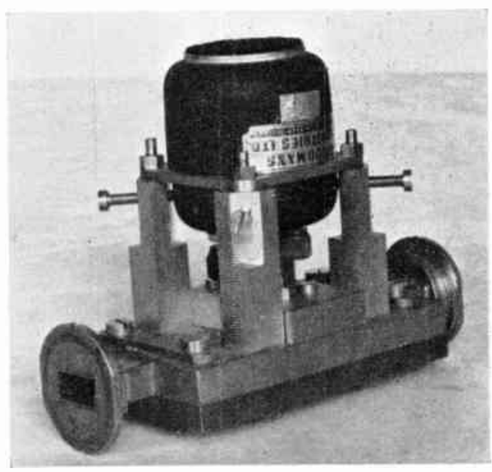


Fig. 5. Vibrator and filter assembly.

3.4. *Frequency Response*

The frequency response of the system has two characteristics. The first is the presence of many resonances in the vibrator and mountings. This is characteristic of the equipment used and additional developmental work is required to remove these.

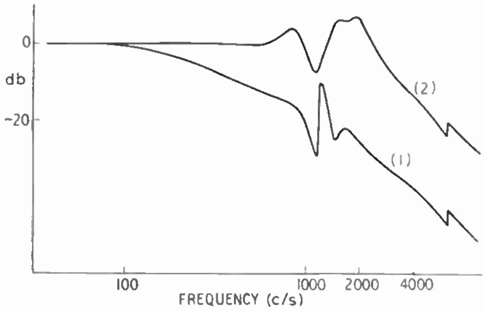


Fig. 6. Frequency response of modulator unit. (1) Without feedback. (2) With feedback.

The second characteristic of the response is the fall in modulation index as the frequency is increased for a constant input power. This is due to the fact that the power required to produce vibration of the plunger of a given amplitude increases with frequency. Fig. 6 shows the frequency response of the modulator unit. With suitable feedback the response can be improved as shown also in Fig. 6, but, of course, at the price of some reduction in depth of modulation.

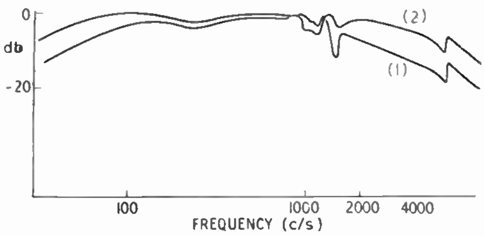


Fig. 7. Overall response. (1) Without feedback. (2) With feedback.

By applying the feedback across the whole system of modulator and receiver, considerable improvement can be obtained in overall performance as seen in Fig. 7. The schematic representation of the system is given in Fig. 8. It will be seen that the feedback is applied between the monitor receiver and the modulator, the final receiver being a wide band video-amplifier. The design of the amplifiers and correcting networks is given elsewhere.[10]

The system was tried out by applying various wave forms to it, and also by examining its reproduction of speech. A measure of the fidelity can be judged from Fig. 9.

4. Conclusions

The results obtained in the work just described show that this method of amplitude modulation of microwaves offers some advantages as compared with the other methods mentioned. It is essentially linear for low modulation depths and simple, in that no wave guide components involving transition from rectangular to round section nor precision cavities of any sort are required.

An improvement on the filter-modulator described might be obtained by replacing the moving plunger by a flexible diaphragm occupying one complete broad face of the filter, the tuning being achieved either by a central screw in the opposite face or by a post carried by the diaphragm itself. This would reduce the radiation loss, since the top and bottom of the filter would be completely closed. It would also remove the unwanted resonances by applying some of the extensive knowledge of microphone design. The diaphragm could be actuated by a moving coil or even perhaps by direct air pressure produced by speech or music.

A further possibility, which would, however, require a great deal more work, follows from the fact that this modulator gives a modulated reflected wave. Thus a receiving horn backed by such a modulator would return to source, radiating a u.h.f. signal of constant amplitude, a fraction of that signal but amplitude modulated, thus making it possible to communicate with a field operator who himself has no microwave generating equipment.

5. Acknowledgment

The authors wish to thank Professor F. M. Bruce, Head of Department of Electrical Engineering, The Royal College of Science and Technology, Glasgow, for providing the facilities to carry out the above described investigation.

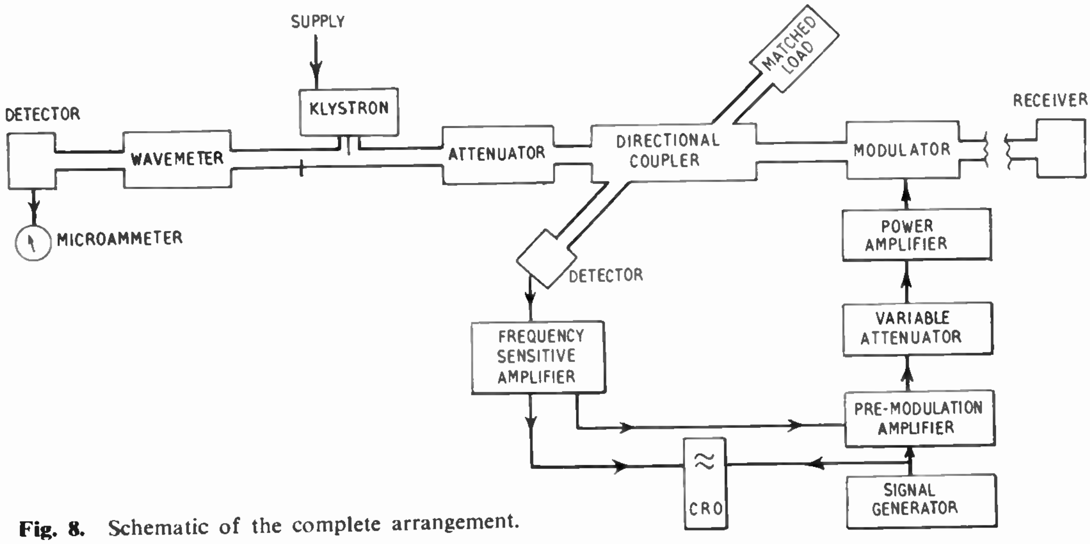
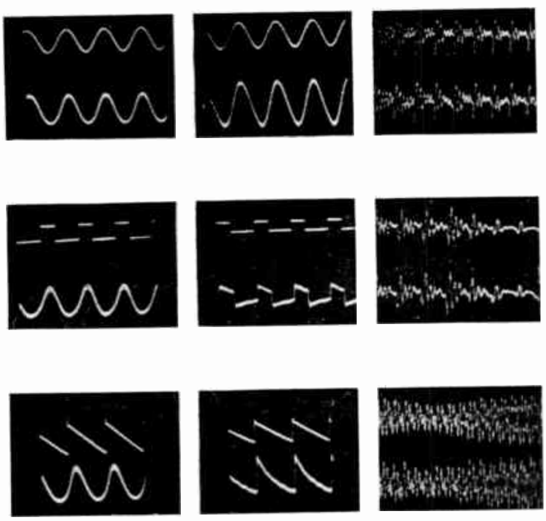


Fig. 8. Schematic of the complete arrangement.



(*a*) (*b*) (*c*)

Fig. 9. Traces showing response of the system to various signals. In each case, upper trace shows the input, lower trace the output.
(*a*) Response to various waveforms without feedback.
(*b*) Response with feedback.
(*c*) Response to speech and music signals.

6. References

1. J. R. Pierce, "Some recent advances in microwave tubes," *Proc. Inst. Radio Engrs*, **42**, p. 1735, 1954.
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4. G. H. B. Thompson, "Ferrites in waveguides," *J.Brit.I.R.E.*, **16**, p. 311, 1956.
5. W. W. H. Clark, W. M. Searle and F. T. Varl. "Ferrite microwave modulator employing feedback," *Proc. Instn Elect. Engrs*, **103**, Part B, p. 485, 1956.
6. H. G. Beljers, "Amplitude modulation of centimetre waves by means of ferrites," *Philips Tech. Rev.*, **18**, p. 82, 1956.
7. A. N. Bhattacharya, "Microwave modulation by variable circuit elements comprising waveguide or cavity resonators," *Indian J. Physics*, **23**, p. 175, 1949.
8. G. C. Southworth, "Principles and Applications of Waveguide Transmission." (D. Van Nostrand Co. Inc., New York, 1950.)
9. N. Marcuvitz, "Waveguide Handbook." (McGraw Hill, New York, 1951.)
10. J. Barbour, "Amplitude Modulation of Microwaves." Thesis for Postgraduate Diploma of Royal College of Science and Technology, 1956.

THE SUPPLY OF STUDENTS FOR TECHNOLOGY

DISCUSSIONS on the future of technological education have so far paid little attention to the source of supply of students to take advantage of the proposed new technical colleges and the new courses being introduced. The Association of Scientific Workers' conference on this subject in December was therefore a timely one; it dealt with science education in the schools, students for full-time courses, and students for part-time day and sandwich courses.

In discussing university training, Professor D. G. Christopherson (Imperial College) pointed out that anything done now would be unlikely to effect the supply of students for six years. However, the increase in the birth-rate immediately after the War would eventually increase the number of students available to universities by 30 per cent., provided the overall standard of attainment remained the same.

It would also be possible to increase the supply if women were encouraged to take up technology and were acceptable to industry.

By reducing the numbers who left school at 15 years, it would also be possible to increase the number of students available to universities by 5 per cent. per annum. Industry could help here by making grants to enable suitable candidates to stay at school to study science.

Regarding the supply of students for part-time and sandwich courses, Dr. J. W. Topping (Principal, Acton Technical College) gave statistics to show that it would be possible to produce 15,000 engineers and scientists a year as envisaged in the White Paper on Technical Education. He included in this figure all qualifications of Higher National Certificate level and above.

At present the universities have places to spare. This was largely due to the fact that the present university age group was the smallest for 100 years, and that accommodation has been considerably increased as a result of the post-war expansion.

In Dr. Topping's opinion, the supply of students for the Diploma in Technology would come either from those who had passed two or more subjects at the advanced level in science, or the top seven per cent. of the Ordinary National Certificate successes. These two sources should provide some 3,000 potential Diplomas in Technology. With 8,000 Higher National Certificates and 1,500 science and engineering degrees, this made a total annual output of 12,500. The increase in the university and technical college population due to the post-war increase in birth-rate would bring the total to 15,000 by 1962. It might also be possible to increase the percentage of the population taking courses of study. In 1954-55 out of the age group 18-20, only 90,000, or roughly 5 per cent., were attending full-time courses in universities and technical colleges; a similar number attended part-time day release courses and 160,000 or just under 10 per cent. took evening classes.

Mr. J. P. de C. Meade, an education and training officer in the electronics industry, indicated the problems of industrial training. In discussing the sandwich scheme, he pointed out the difficulty of firms working the one-sandwich-per-year scheme, where all the apprentices went to the college and all returned to industry for alternate periods of six months. With a large number of apprentices it was necessary to operate a pairing scheme, where one half of the apprentices were at college while the other half were in industry.

It was essential that industry should put training on a very much higher level of priority. Training tended to take second place and needed much greater attention by Management. It was important that every firm should train at least as many engineers as it required.

Similarly, industry must completely invert its ideas on part-time day release, or sandwich training release. The attitude at the present time was that it was a generous move by the employer to the apprentice. In actual fact it was an absolute necessity for the existence and survival of industry. The success of all these training schemes depended on employers releasing their apprentices. As a full four years training would cost at least £2,000 per apprentice, the success of the scheme was contingent to a large extent on industry's contribution.

NOISE AND ITS "SPECTRUM"

by

F. N. H. Robinson, D.Phil., M.A. †

SUMMARY

This paper explains the physical ideas underlying the familiar equation for shot noise in a temperature limited current. The mathematical analysis is based on Campbell's theorem, which is proved, and some consideration is given to the meaning of the averages which appear in expressions describing noise.

1. Introduction

The useful sensitivity of electronic apparatus is limited by noise. Two of the most important sources of noise are the fluctuations in electron emission in the valves, which cause shot noise, and the thermal fluctuations of voltage in the resistive elements which cause Johnson noise. The effect of these sources of noise on the performance of a given piece of apparatus is readily analysed using a number of well known formulae. One of the simplest of those formulae is that for the mean square shot noise current I_n^2 occurring within a bandwidth B in a temperature limited current I_0.

$$I_n^2 = 2e\, I_0\, B \qquad \text{.........(1)}$$

It is the purpose of this paper to examine the physical content of this equation and its relation to the fundamental random process which is responsible for the noise. This particular equation has been chosen because it can be discussed with the least introduction of extraneous ideas and in the hope that a clear understanding of this one equation will prove helpful in understanding other equations such as that for Johnson noise in a resistance R.

$$V_n^2 = 4\, kT\, R\, B \qquad \text{.........(2)}$$

2. Probability Expression for Shot Noise

Shot noise occurs because electrons have a finite charge *and* are not emitted at a uniform rate from thermionic cathodes. The emission of electrons is a random process and although the mean rate averaged over a long time will be independent of the time of averaging, the rate averaged over short intervals will show fluctuations about that mean.

In order to analyse the fluctuations we consider measurements of the total number of electrons n_i (or the total charge $q_i = en_i$) which are emitted from a temperature limited cathode in a time T. This measurement is repeated a great many (M) times, and n_i is the result of the ith measurement. We can then form the averages

$$\bar{n} = \frac{1}{M}\sum_{i=1}^{M} n_i \qquad \text{.........(3)}$$

$$\overline{n^2} = \frac{1}{M}\sum_{i=1}^{M} n_i^2 \qquad \text{.........(4)}$$

and

$$\overline{\Delta n^2} = \frac{1}{M}\sum_{i=1}^{M} (n_i - \bar{n})^2 = \overline{n^2} - (\bar{n})^2 \qquad \text{.........(5)}$$

For very large values of M these averages will tend to definite values which are not affected by consideration of the results of further measurements.

Clearly the average charge which flows in a time T is $e\bar{n}$ and the average current

$$I_0 = e\bar{n}/T \qquad \text{.........(6)}$$

while the mean square deviation of the results of individual measurements from that mean value is

$$\overline{\Delta I_T^2} = \frac{e^2}{T^2}\,\overline{\Delta n^2} \qquad \text{.........(7)}$$

We shall now show that

$$\overline{\Delta n^2} = \bar{n}$$

and so

$$\overline{\Delta I_T^2} = e\overline{I_T}/T \qquad \text{.........(8)}$$

To do this we subdivide every interval T into a great number N of equal intervals t each so small that the probability p of even one electron being emitted in any specified interval

* Manuscript received 31st October, 1956. (Paper No. 385.)

† English Electric Research Fellow, The Clarendon Laboratory, Oxford.

U.D.C. No. 621.355.

is small and proportional to:—

$$p = \nu t \ll 1 \qquad \text{.........(9)}$$

We now consider an interval T in which exactly n electrons are emitted. Thus in n of the intervals t an electron was emitted, while in $(N-n)$ no electron was emitted. The *a priori* probability of this happening in specified intervals is

$$p^n(1-p)^{N-n}$$

The intervals of course are not specified by the knowledge that n electrons were emitted in T and the number of possible combinations of intervals which would give exactly this result is

$$^NC_n = \frac{N!}{n!\,(N-n)!}$$

which is just the coefficient of x^n in the expansion of $(1+x)^N$. The total probability of exactly n electrons being emitted in T is thus

$$P(n, T) = {}^NC_n\, p^n\, q^{N-n} \qquad \text{.........(10)}$$

where we have written $q = 1-p$ ....(11)

If now we consider very many intervals T the various values of n will occur with a frequency proportional to $P(n, T)$ and so the average values $\overline{n}$ and $\overline{n^2}$ will be given by

$$\overline{n} = \sum_{n=0}^{N} n\, P(n, T) = \sum_{0}^{N} n\, {}^NC_n\, p^n q^{N-n} \qquad \text{...(12)}$$

$$\text{and } \overline{n^2} = \sum_{0}^{N} n^2\, {}^NC_n\, p^n q^{N-n} \qquad \text{.........(13)}$$

These two sums may be evaluated very simply as follows. From the binomial theorem we have

$$\sum_{0}^{N} {}^NC_n\, p^n\, q^{N-n} = (p+q)^N = 1^N = 1$$

$$\text{thus } \sum_{0}^{N} n\, {}^NC_n\, p^n\, q^{N-n} = p\frac{\partial}{\partial p}\sum_{0}^{N} {}^NC_n\, p^n\, q^{N-n} =$$

$$= p\frac{\partial}{\partial_p}(p+q)^N = pN\,(p+q)^{N-1} = pN \;\text{......(14)}$$

and similarly

$$\sum n^2\, {}^NC_n\, p^n\, q^{N-n} = p^2\frac{\partial^2}{\partial p^2}(p+q)^N + p\frac{\partial}{\partial p}(p+q)^N$$
$$= p^2N^2 - p^2N + pN$$

Thus

$$\overline{n} = pN$$
$$\overline{n^2} = (\overline{n})^2\left(1 - \frac{1}{N}\right) + \overline{n}$$

$$\text{and} \quad \overline{\Delta n^2} = \overline{n^2} - (\overline{n})^2 = \overline{n} - \frac{(\overline{n})^2}{N} \qquad \text{.........(15)}$$

As the number N of intervals t into which T is subdivided is made arbitrarily large the last term may be neglected and we have the result

$$\overline{\Delta n^2} = \overline{n} \qquad \text{.........(16)}$$

that is, the mean square fluctuation of the measurements of n over equal intervals is equal to the mean of all the measurements. This leads directly to equation (8) which gives the mean square fluctuation of the individual measurements of the current (each lasting a period T) from the long term average.

If we observe the current with an amplifier of bandwidth B we are in effect lumping together all those events which occur in a time $T \sim 1/B$ and so in a sense the output of the amplifier will at any instant be the average over the preceding period $T \sim 1/B$. The fluctuations observed in the current will therefore be expected to be of the order of

$$\overline{\Delta I_B^2} \sim e\, I_0 B$$

We now proceed to give a more quantitative derivation of this expression and one which throws some light on the meaning of the equation.

3. Noise Response of Amplifiers

We consider an instrument such as an amplifier which responds to the arrival of a single electron in the input circuit at time $t=0$ by giving a response (which may be a voltage, a c.r.t. deflection or a current, etc.) which is a function of time

$$\theta = f(t) \qquad \text{.........(17)}$$

We shall now prove Campbell's theorem†, which states that if the input consists of a random succession of electrons arriving at a mean rate ν (i.e. a temperature limited current $I_0 = e\nu$) then the mean response is

$$\overline{\theta} = \nu\int_{-\infty}^{\infty} f(t)\, dt \qquad \text{.........(18)}$$

† N. R. Campbell, *Proc. Camb. Phil. Soc.*, **15**, pp. 117-136 and 310-328, 1909. See also S. O. Rice, "Mathematical Analysis of Random Noise," *Bell Syst. Tech. J.*, **23**, p. 282, 1944.

and the mean square deviation of the response at any instant from the mean is

$$\overline{\Delta\theta^2} = \nu \int_{-\infty}^{\infty} f^2(t)\, dt \qquad \text{.........(19)}$$

Take an interval of time of length T so long that the responses due to events near the end of T which fall outside T are a negligible fraction of the total response. This is an approximation of the sort usually used in problems where it is desirable to be able to neglect "edge effects." Consider a particular interval of that length in which exactly n events occur. Number these events from 1 to n and let t_k be the time at which the kth event occurred. The response of the instrument at time t is then

$$\theta_n(t) = \sum_{k=1}^{n} f(t - t_k) \qquad \text{.........(20)}$$

If we examine a great many M intervals T the number having exactly n events occurring in them will be $M\,P(n,T)$ where $P(n,T)$ is given by equation (10). For a fixed value of t and for each interval having just n events, $\theta_n(t)$ will have a value which will depend on the times t_k of arrival of the events. These events are distributed at random over the interval and so if we average over this distribution, i.e. over all possible times of events in the interval we have

$$\overline{\theta_n(t)t_k} = \int_0^T \frac{dt_1}{T} \int_0^T \frac{dt_2}{T} \cdots \int_0^T \frac{dt_n}{T} \sum_{k=1}^{n} f(t - t_k) =$$

$$= \sum_{k=1}^{n} \int_0^T f(t - t_k)\, \frac{dt_k}{T} \qquad \text{.........(21)}$$

If the interval T is long (in the sense that we have supposed) then we may replace the limits on the integral by $\pm\infty$ and have

$$\overline{\theta_n(t)t_k} = \sum_{k=1}^{n} \frac{1}{T} \int_{-\infty}^{\infty} f(t_k)\, dt_k = \frac{n}{T} \int_{-\infty}^{\infty} f(t)\, dt \qquad \text{.........(22)}$$

We now average this expression over intervals in which other than exactly n events occurred and obviously the result is

$$\overline{\theta(t)} = \frac{\overline{n_T}}{T} \int_{-\infty}^{\infty} f(t)\, dt = \nu \int_{-\infty}^{\infty} f(t)\, dt \qquad \text{.........(23)}$$

since $\overline{n_T}/T$ is the mean rate ν of occurrence of events. The first part of the theorem has thus been proved.

To prove the second part we calculate $\overline{\theta^2}$. By analogy with (20) we have

$$\theta_n^2(t) = \sum_{k=1}^{n} \sum_{l=1}^{n} f(t - t_k)\, f(t - t_l) \qquad \text{.........(24)}$$

Averaging this over all arrival times t_k and t_l with n and t fixed

$$\overline{\theta_n^2(t)t_k,t_l} = \sum_{k=1}^{n} \sum_{l=1}^{n} \int_0^T \frac{dt_1}{T} \cdots \int_0^T \frac{dt_n}{T}\, f(t - t_k)\, f(t - t_l)$$

$$\text{.........(25)}$$

In this integral there are n terms in which $k = l$ and $n^2 - n$ terms in which $k \neq l$.
The terms in which $k = l$ give

$$\frac{n}{T} \int_{-\infty}^{\infty} f^2(t)\, dt$$

the terms in which $k \neq l$ give

$$\frac{n(n-1)}{T^2} \left[\int_{-\infty}^{\infty} f(t)\, dt \right]^2$$

when we make the passage from the analogue of equation (21) to that of (22).

We now average over all possible values of n and obtain

$$\overline{\theta^2} = \frac{\overline{n}}{T} \int_{-\infty}^{\infty} f^2(t)\, dt + \frac{\overline{n^2} - \overline{n}}{T^2} \left[\int_{-\infty}^{\infty} f(t)\, dt \right]^2$$

Now $\overline{n^2} = (\overline{n})^2 + \overline{n}$ from equation (15)

and so $\overline{\theta^2} = \dfrac{\overline{n}}{T} \displaystyle\int_{-\infty}^{\infty} f^2(t)\, dt + (\overline{\theta})^2$ using (6)

Thus $\overline{\Delta\theta^2} = \overline{\theta^2} - (\overline{\theta})^2 = \nu \displaystyle\int_{-\infty}^{\infty} f^2(t)\, dt$ ....(19)

which proves the second part of the theorem.

Expressing in terms of the mean current I_0 in the input circuit we have

$$\overline{\Delta\theta^2} = \frac{I_0}{e}\int_{-\infty}^{\infty} f^2(t)\,dt \qquad \text{.........(26)}$$

4. Fourier Transformation of Amplified Noise Equations

The characteristics of amplifiers are not generally specified by giving their response to the arrival of a single electron at zero time but rather by giving their gain as a function of frequency and so to make use of equation (26) we must find the relation between f(t) and the gain frequency curve $G(f)$. To make matters perfectly definite we shall regard θ as a voltage and $G(f)$ as the voltage produced at the output of the amplifier when unit current of frequency f flows in the input circuit.

The first step is to define $F(t)$ as the response of the amplifier to unit charge arriving at $t=0$ clearly

$$F(t) = \frac{1}{e}\,f(t) \qquad \text{.........(27)}$$

and so

$$\overline{\Delta\theta^2} = eI_0\int_{-\infty}^{\infty} F^2(t)\,dt \qquad \text{.........(28)}$$

If we apply a varying current $I(t)$ to the input of the amplifier the response at time t due to the current flowing at t_1 is

$$d\theta(t) = I(t_1)\,dt_1\,F(t-t_1)$$

and the response due to a prolonged application of the current is

$$\theta(t) = \int_{-\infty}^{\infty} I(t_1)\,F(t-t_1)\,dt_1 \qquad \text{.........(29)}$$

Now we can develop $I(t_1)$ as a Fourier integral

$$I(t_1) = \int_{-\infty}^{\infty} J(f)\,e^{2\pi i f t_1}\,df \qquad \text{.........(30)}$$

and so

$$\theta(t) = \int_{-\infty}^{\infty}\int J(f)\,e^{2\pi i f t_1}F(t-t_1)\,df\,dt_1 \qquad \text{.........(31)}$$

Now if we had specified instead of F the response of the amplifier $\varphi(f)$ to a current $J(f)$ of frequency f by giving the "gain" $G(f)$ we should have had

$$\varphi(f) = G(f)J(f) \qquad \text{.........(32)}$$

and

$$\theta(t) = \int_{-\infty}^{\infty}\varphi(f)\,e^{2\pi i f t}\,df = \int_{-\infty}^{\infty} G(f)\,J(f)e^{2\pi i f t}\,df \quad \text{....(33)}$$

Comparing (31) and (33) we see that

$$G(f) = -\int_{-\infty}^{\infty} F(t)e^{-2\pi i f t}\,dt \qquad \text{.........(34)}$$

Thus the "current gain" of the system is the Fourier Transform of the response to unit charge.

We now use Parseval's theorem† which states that if G_1 and F_1 and G_2 and F_2 are Fourier Transforms of each other then

$$\int_{-\infty}^{\infty} G_1(f)G_2(f)\,df = \int_{-\infty}^{\infty} F_1(t)F_2(-t)\,dt \quad \text{........(35)}$$

If in (7) we put $F_1(t)=F(t)$, $F_2(t)=F(-t)$ then $G_1(f)=G(f)$ and $G_2(f)=G(-f)=G^*(f)$.

The last result is a consequence of equation (34) and the fact that $F(t)$ is a real function.

We then find that:

$$\int_{-\infty}^{\infty} G(f)\,G^*(f)\,df = \int_{-\infty}^{\infty} F^2(t)\,dt$$

Now $G(f)\,G^*(f)$ is an even function of f since since $G^*(f)=G(-f)$ and so we can write this as

$$2\int_{0}^{\infty} G(f)G^*(f)\,df = \int_{-\infty}^{\infty} F^2(t)\,dt \qquad \text{.........(36)}$$

The fluctuations in the output voltage θ due to a temperature limited current I_0 in the input circuit are therefore

$$\overline{\theta^2} = 2eI_0\int_{0}^{\infty} G(f)\,G^*(f)\,df \qquad \text{.........(37)}$$

The derivation of equation (37) depends on the statistical properties of the noise current and the applicability of Fourier analysis to the response of amplifiers. Nowhere have we had to assume that a noise current possesses a spectrum and can be Fourier analysed. However, the result in (37) is of course exactly the same as that which would be obtained if we attributed to the shot noise a mean square current fluctuation in bandwidth df of

$$\overline{dI_n^2} = 2e\,I_0\,df \qquad \text{.........(38)}$$

† Titchmarsh, E. C., "Introduction to Fourier Integrals." (Oxford University Press, 1948.)

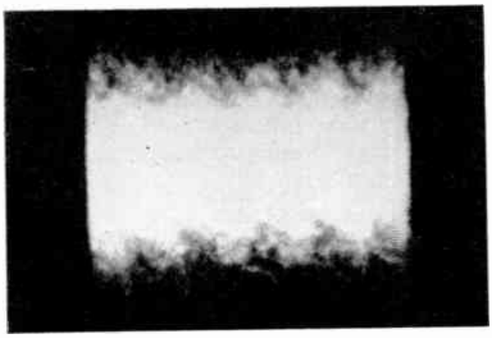

Fig. 1. Oscillogram of noise output of amplifier with unsynchronized time-base of frequency between bandwidth and its centre frequency.

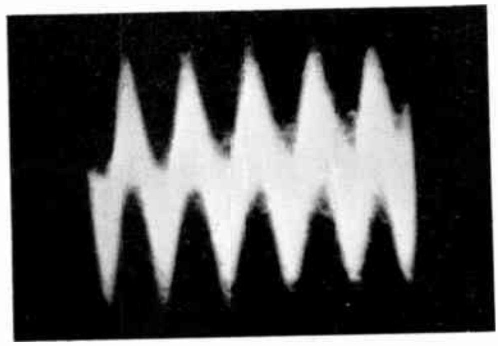

Fig. 2. Oscillogram of noise output of amplifier with time-base synchronized with the noise output itself.

with components of different frequencies being entirely uncorrelated. It is in this sense therefore that we should interpret equations such as (1).

5. Practical Illustration of Noise Averages

The averages which appear in the various equations describing noise are usually interpreted in a way which is most easily illustrated by giving an example. Suppose that a temperature limited current I_0 flows through a resistance R across the input terminals of an amplifier of bandwidth B and the voltage gain G. The fluctuating output voltage will have a mean square value

$$\overline{V_n^2} = 2\, e\, I_0\, R^2 G^2 B \qquad \text{.........(39)}$$

i.e., if we rectify the output using a square law detector and measure the resulting d.c. voltage with a long period meter we shall obtain the same result as if we had applied a sinusoidal voltage of r.m.s. value V_n to the detector.

There is, however, another way of considering the average which is useful when the noise is presented on a cathode-ray oscilloscope, and that again is most easily explained by an example. We suppose that the amplifier considered has a centre frequency f which is much greater than the bandwidth B. For example, it might be a 1 Mc/s amplifier with a bandwidth of 1 kc/s. If we observe on a c.r.o. the output for a time small compared with $1/B$ but large compared with $1/f$ we shall see a sine wave of arbitrary phase and amplitude V_n. Repetition of this measurement, as for example by using an unsynchronized repetitive time-base, will result in a superposed series of traces (because of the persistence of vision and the screen phosphor) with no apparent phase relation but an r.m.s. value given by (39). If the repetitive time-base is running at a frequency intermediate between B and f the noise will have the characteristic appearance shown in Fig. 1 which was taken with $f=100$ kc/s $B=2$ kc/s and a time-base repetition frequency of 20 kc/s. In this case one sweep of the spot covers about 5 cycles at the centre frequency.

We can, however, see a very different picture by synchronizing the time-base with the noise output of the amplifier. Because $B \ll f$ the noise voltage is approximately sinusoidal for some f/B cycles later (i.e. for longer than the sweep time). Each traverse of the screen starts at the same phase relative to the noise and so the resulting pattern is a diffuse sine wave (Fig. 2), diffuse because the amplitude of the noise wave will vary from sweep to sweep.

6. Conclusion

The formulae

$$I_n^2 = 2\, e\, I_0\, B \qquad \text{.........(1)}$$

for the shot noise current observed in bandwidth B is simply related by Campbell's theorem to the fact that the emission of electrons from a cathode is a random process. It should be interpreted to mean only that the output of an amplifier of bandwidth B excited by shot noise will consist at any instant of a sinusoidal voltage whose mean square value averaged over many instants can be calculated using (1). If the voltage is measured by a device with a time constant long compared with $1/B$, that is equivalent to averaging over many instants.

GRADUATESHIP EXAMINATION—NOVEMBER 1956—PASS LIST

These lists contain the results for *all* successful candidates in the November Examination. A total of 467 candidates entered for the examination which was held at 52 centres. This number included 158 candidates attempting all or parts of the examination in order to complete qualification for election to Graduateship or Associate Membership of the Institution.

The following candidates having completed the requirements of the Graduateship Examination, are eligible for transfer or election to Graduateship or higher grade of membership

United Kingdom and European Centres

ASLAND Greggar. (S) *Oslo.*
ATHANASSIADES, Elias. *Athens.*

BENNETT, Wilfred Dennis. (S) *Manchester.*
BOICE, Cyril John. *Manchester.*
BRACE, William James. (S) *Cardiff.*

CHANNING, Ronald Francis. (S) *London.*
COOPER, John Derek. *London.*

DEDMAN, William Leonard *London.*
DE RUYTER, Albertus Hermanus Maria. (S) *Delft.*

GREEN, Lawrence Young. (S) *London.*

HADJIEMETRIOU, Demetrious. (S) *Athens.*

NEIGHBOUR, Kenneth John. *London.*

RETTIE, Alister Brian. *London.*

SANDYS, Maurice Arthur. (S) *London.*
SENIOR, Eric. (S) *Manchester.*
SMITH, Charles Edward. (S) *London.*
SMITH, John Douglas. (S) *London.*
STICKLER, Gordon Alan. (S) *Cardiff.*

ZAIKOS, Demetrius. (S) *Athens.*

Overseas Centres

ACHUTHAN, Madras Gopolan. (S) *Madras.*

BASHYAM. R. (S) *Trichinopoly.*

CHACHAM, Shaul. (S) *Tel-Aviv.*

DHALL, Raj Kumar. (S) *Dehra Dun.*

GOGATE, Bhalchandra Damsdar. (S) *Delhi.*

IZZARD, Malcolm Ian. (S) *Durban.*

KAPOOR, Mulk Raj. (S) *Calcutta.*

MADAN, Amrit Lal. (S) *Calcutta.*
MALHOTRA, Bahri Jagmohanlal. (S) *Delhi.*

PINTO, Cyprian. *Bombay.*

SWAMINATHAN, Kashi Ramamurthu. (S) *Madras.*

The following candidates were successful in the Parts indicated

United Kingdom and European Centres

AKINYEMI, Isaac Olaonipekun. (3) (S) *London.*

BENTLEY, Edward Leslie. (1, 2, 3) (S) *London.*
BIRD, Gordon Joseph Alexander. (1) (S) *London.*
BONNER, John Stafford. (5) (S) *London.*
BOWEN, Kenneth. (1, 2, 3) (S) *London.*
BOWN, Kenneth Albert. (1) (S) *London.*

CHANDRA, Jagdish. (3) (S) *London.*

DIVECHA, Gautamrai Amritlal (4) (S) *London.*

EXARCHOS, Vladimir. (4) (S) *Athens.*

GALLIVER, Geoffrey Edward Lewis. (3) (S) *London.*
GARDIKIS, Dimitrios. (3) (S) *Athens.*
GEORGE, Julian. (4) (S) *London.*
GUPTE, Rajendra Keshavrao. (5) (S) *London.*

HALTON, Dennis Lewin. (3) (S) *London.*
HANCOCK, Harry James. (1) *Birmingham.*
HEWITT, Patrick John. (3) (S) *Birmingham.*

KARAMANOLIS, Ch. (3) (S) *Athens.*
KEANE, James. (5) (S) *Dublin.*
KENNY, Gerald (1) (S) *London.*
KOLLIAS, Spiros. (3) (S) *Athens.*

LARGE, Douglas. (2) (S) *London.*

MACKENZIE, Ian. (1) *London.*
MENKAL, Raymond. (1) (S) *London.*
MURPHY, Mathew. (3) (S) *Dublin.*

NATARAJAN, Ramakrishnan Iyer. (1) (S) *London.*
NEED, Richard John. (2) (S) *London.*
NICHOLS, Basil Hopes. (4) (S) *Newcastle.*

PAIS, Aloysius Francis. (2) (S) *Cardiff.*
PEVERETT, Anthony Michael. (4) (S) *London.*
PODLASKI, Jan. (3) (S) *Manchester.*

RICKERS, Denis. (2) (S) *London.*

SERELEAS, Christos. (3) (S) *Athens.*
STEPHEN, Sidney George. (2) (S) *Cardiff.*

TOMLINSON, Edward Rex. (1) (S) *Manchester.*
TOWNSEND, Brian Joseph. (2) (S) *London.*
TSAMOUSSIS, Efstratios. (5) (S) *Athens.*

WALES, Sydney Alfred. (4) (S) *H.M.S. Armada.*

ZIJDEMANS, Leendert Johannes. (2) (S) *London.*

Overseas Centres

AGARWAL, Durga Prasad. (3) *Delhi.*
ALTARATZ, Jacob. (3) (S) *Tel-Aviv.*
ARORA, Surendra Prakash. (1) (S) *Bangalore.*
AZAR, Yoram. (3) (S) *Tel-Aviv.*
BHACKA, Sam Jamshedji. (4) (S) *Bangalore.*
BHAGAT, Shiv Raj Huria. (4) (S) *Agra.*
BHATTACHERJEE, Amal Kumar. (4) (S) *Kanpur.*

CHATTOPADHYAY, Anil Baran (2) (S) *Agra.*

DAVIS, Trevor. (2) (S) *Woomera.*

FONG YAN, Alick. (1, 2) *Hong Kong.*

GOVINDARAGHAVAN, Doraiswamy. (5) *Delhi.*
GREENWOOD, Frank. (1, 2, 3) (S) *Adelaide.*
GURCHARAN SINGH SURIE. (2) (S) *Bangalore.*

HANDA, Jakdish Rai. (2) (S) *Agra.*

IBRAHIM, Tipu Mohamed. (2) (S) *Bangalore.*
INDER, James Haviland. (3) (S) *Auckland.*
ISRANI, Indur Kumar. (3) (S) *Bombay.*

JAIN, Sant Perkash. (3) (S) *Delhi.*
JOSHI, Devendra. (3) (S) *Bombay.*

KAMALJIT SINGH. (4) (S) *Delhi.*
KHATRI, Dindayal Tahilram. (3) (S) *Bangalore.*

LIPSCHITZ, Gerhard Fad. (2, 3) *Tel-Aviv.*
LUND, Hugh Forsyth (4) (S) *Durban.*

MALHOTRA, Chaman Lal. (1) (S) *Calcutta.*
MALHOTRA, Madan Mohan. (1, 2, 3) (S) *Agra.*
MANTEL, Juval. (1, 2, 3) (S) *Tel-Aviv.*
MATHEWS, Abraham. (4) (S) *Agra.*
MEHTA, T. Raj. (2, 3) (S) *Delhi.*
MITRA, Gobinda Lal. (1) (S) *Kanpur.*
MUKHERJEE, Samir Kumar. (3) (S) *Delhi.*

NARAYANA MENON, Pottekkat. (1, 2, 3) (S) *Bombay.*

PATTABIRAMAN, A. K. (3) (S) *Madras.*
PERERA, Prangagae Wimaladassa (1) (S) *Colombo.*
PISHARODY, A. P. Unnikrishna. (3) (S) *Bombay.*

QURESHI, Mohd Aslam. (1, 2, 3) (S) *Lahore.*

RAJAGOPAL, A. (2, 3) (S) *Bangalore.*
RAJENDRA, Nath. (4) (S) *Delhi.*
RUBEN, Moshe. (1, 2, 3) (S) *Tel-Aviv.*
RUPRAI, Balwant Singh. (2) (S) *Bombay.*

SANKARA RAO, Nagaraja. (4) (S) *Madras.*
SESHA, Venkatarama Adi. (2) (S) *Delhi.*
SHAHI, Brij Lal. (3) (S) *Delhi.*
SIDDIQUI, Tausif Ahmed. (3) (S) *Karachi.*
SOOD, Omkar Nath. (2, 3) (S) *Dehra Dun.*
SURI, Sham Lal. (1) (S) *Bombay.*

UPPAL, Kanwal Krishnan. (2) (S) *Delhi.*

VARMA, Sarvottam. (3) (S) *Agra.*
VENKITACHALAM, Y. (4) (S) *Bombay.*
VIRDI, Harbans Singh. (1) (S) *Lucknow.*
VIRINDER SINGH. (3) (S) *Bangalore.*

(S) denotes a registered Student.

VARIATION OF CABLE LOSS WITH STANDING WAVE RATIO *

by

E. G. Hamer, B.Sc.(Eng.)(Member) †

SUMMARY

The effects of power losses in mismatched transmission lines are considered, and formulae and nomograms are derived for the increased effective attenuation and power loss in the cable. Consideration is also given to the reduction in the standing wave ratio between the receiving and sending end and a nomogram derived.

LIST OF SYMBOLS

n = voltage attenuation along cable.

$= \dfrac{\text{sending end incident voltage}}{\text{receiving end incident voltage}}$

$= \dfrac{\text{sending end reflected voltage}}{\text{receiving end reflected voltage}}$

Z_0 = characteristic impedance of cable (assumed resistive).

K = reflection coefficient $= k|\varphi$

E_I & E_R = r.m.s. value of incident and reflected voltage at receiving end of the cable.

S = standing wave ratio.

1. General Considerations

Cables of the coaxial and balanced twin types are commonly used as radio frequency connections at v.h.f. and u.h.f. In certain instances the power loss caused by the cable must be kept to a minimum, or when measurements of power are being made the losses caused by the cable must be accurately known. If a means of measuring standing wave ratio is available, such as a slotted line or pair of directional couplers the attenuation of the matched cable may be measured‡: and if S_{oc} is the standing wave ratio with one end of the cable open or short-circuited, then

Attenuation of the matched cable in db

$$= 20 \log_{10} \sqrt{\left(\frac{S_{oc}+1}{S_{oc}-1}\right)}$$

This attenuation is that to be expected when the cable is in use, and matched so that the standing wave ratio is 1 and there is no reflected energy from the load. If, however, the load is mismatched to the cable, there will be a standing wave caused by reflected energy and the losses and hence the effective attenuation will be increased. It is of importance to know the magnitude of these extra losses, so that they may be allowed for; or alternatively at what s.w.r. the extra losses have a reasonably small value.

Assume that a steady state condition has been reached, then there will be four voltages to consider:

(*a*) The sending end voltage

(*b*) The receiving end voltage

(*c*) The incident wave voltage

(*d*) The reflected wave voltage

The r.m.s. value of the latter two will vary along the length of the line owing to the line losses.

The sending end power will be the difference between the incident power, and the reflected or returned power at the sending end; and similarly at the receiving end.

It is important to distinguish between the voltage and power of the generator, load, and the two travelling waves.

2. Attenuation of Mismatched Cable

The power output at the receiving end of the cable is:—

$$P_{out} = \frac{E_I^2 - E_R^2}{Z_0} \quad \text{where } E_R = kE_I$$

$$= \frac{E_I^2}{Z_0}(1 - k^2)$$

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† Westinghouse Electric Corporation, Baltimore, Md., U.S.A.

U.D.C. No. 621.315.212/3: 621.317.341.3.

‡ E. G. Hamer, "Slotted line techniques," *Electronic Engineering*, **23**, pp 466-470, December 1951.

The power input at the sending end is:—

$$P_{in} = \frac{(nE_I)^2 - (E_R/n)^2}{Z_0} = \frac{E_I^2}{Z_0}(n^2 - k^2/n^2)$$

$$\text{Attenuation of cable} = \frac{\text{power input}}{\text{power output}} = \frac{n^2 - k^2/n^2}{1 - k^2}$$

If the cable is matched at the load end $k = 0$ and $S = 1$ and the attenuation of the matched cable $= n^2$

$$\frac{\text{attenuation of mismatched cable}}{\text{attenuation of matched cable}} = \frac{1 - k^2/n^4}{1 - k^2}$$

$$= M \quad ....(1)$$

and as $K = \dfrac{S-1}{S+1}$

(S = standing wave ratio at load.)

$$M = \frac{(S+1)^2 + (S-1)^2/n^4}{4S}$$

For values of s.w.r. less than 2 and $n^2 > 2$ (i.e. a matched cable of attenuation greater than 3 db)

$$M = \frac{(S+1)^2}{4S} \quad ....(2)$$

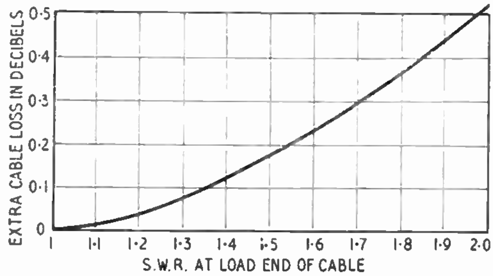


Fig. 1. Extra loss in cable due to mismatch for s.w.r. less than 2, cable attenuation greater than 3 db.

Figure 1 is a graph showing the approximate correction in decibels (as derived from equation (2)) which must be added to the attenuation expressed in db of the matched cable for different values of the s.w.r. at the load.

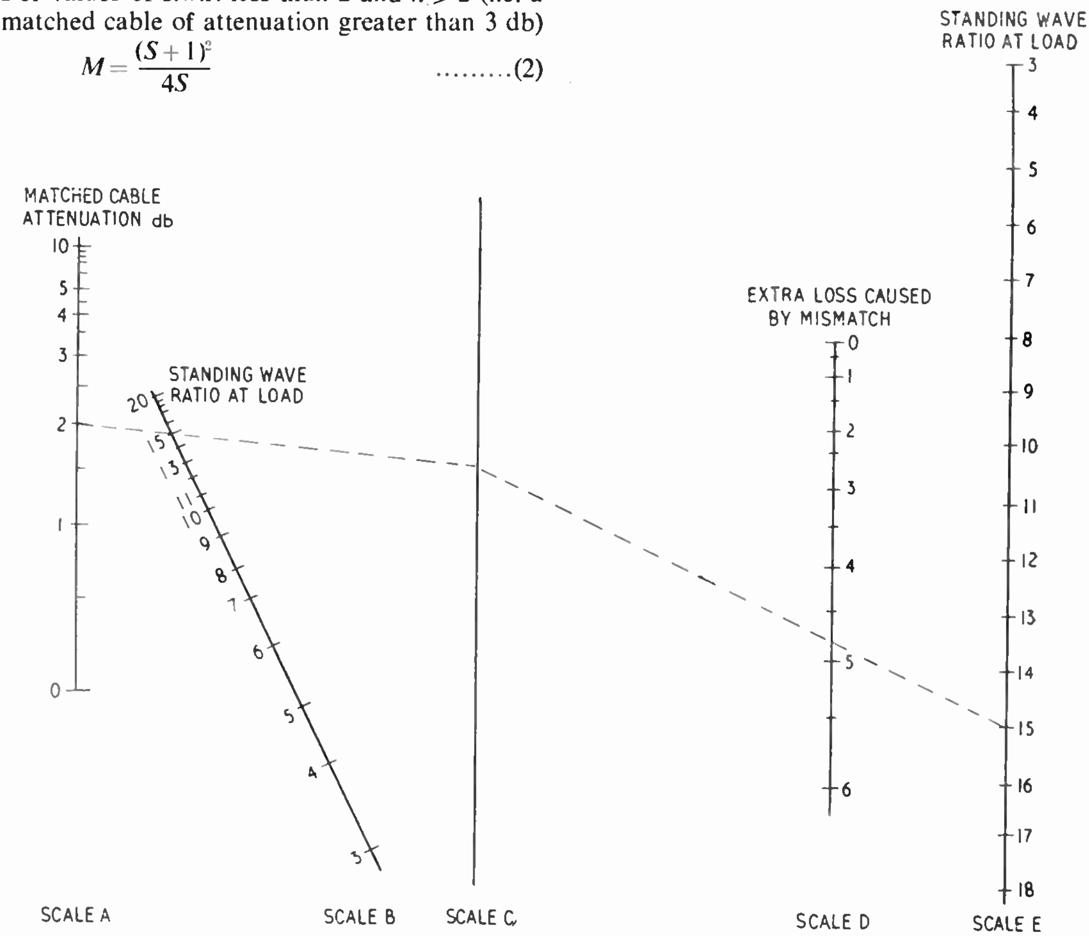


Fig. 2. Nomogram for additional loss due to mismatched cable.

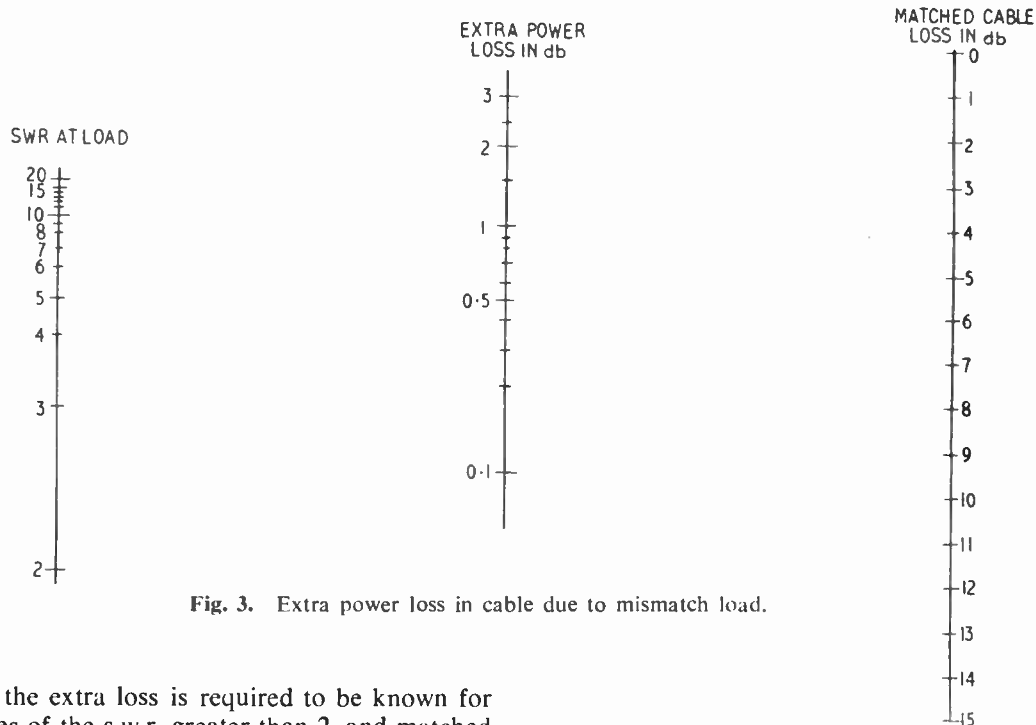


Fig. 3. Extra power loss in cable due to mismatch load.

If the extra loss is required to be known for values of the s.w.r. greater than 2, and matched cable attenuation of less than 3 db, the formula of equation (1) must be used and this can be conveniently used in the form:—

$$\frac{\text{attenuation of mismatched cable}}{\text{attenuation of matched cable}} = \frac{1}{1-k^2} - \frac{k^2}{1-k^2} \cdot \frac{1}{n^4} \quad \text{.........(3)}$$

Figure 2 is a nomogram for the solution of equation (3). To use this nomogram first the matched cable attenuation (Scale A) is joined to the standing wave ratio at the load (Scale B) and the line produced to obtain an intersection on Scale C. This intersection is then joined to the s.w.r. at the load (Scale E) and the extra attenuation caused by the mismatch is shown on Scale D.

In the example shown the load s.w.r. = 15 and the matched cable attenuation = 2 db giving an extra loss of 4·75 db.

3. Power Loss in Mismatched Cable

In some cases where high power equipment is in use the extra power being dissipated in a cable due to a mismatch may be of importance, and we have

$$\text{Power lost in cable} = \frac{E_I^2\,(n^2 - k^2/n^2)}{Z_0} - \frac{E_I^2\,(1 - k^2)}{Z_0}$$

Under matched conditions $k = 0$ and the matched power loss is

$$\frac{E_I^2}{Z_0}(n^2 - 1)$$

$$\frac{\text{power loss in mismatched cable}}{\text{power loss in matched cable}} = 1 + k^2/n^2 \quad \text{...(4)}$$

and Fig. 3 is a nomogram for equation (4) showing the extra power loss expressed in db for a mismatched cable.

It will be seen that the additional power loss, and increased attenuation in proportion to those for the matched cable is considerably reduced for cables with a large attenuation. This is caused by the attenuation of the cable causing a marked reduction in the standing wave ratio as the measuring point is moved further away from the load towards the generator.

4. Reduction of S.W.R. by Cable

Quite short lengths of cable or waveguide can reduce a large s.w.r. to a much smaller value.

If S_m = standing wave ratio at sending end of cable

k_m = reflection coefficient at sending end of cable

k = reflection coefficient at load.

$$S_m = \frac{1+k_m}{1-k_m} = \frac{1+k/n^2}{1-k/n^2} = \frac{n^2+k}{n^2-k} \qquad \ldots\ldots\ldots(5)$$

and equation (5) is expressed in the form of the nomogram of Fig. 4. From this it can be seen that a cable of 6 db attenuation will reduce a load s.w.r. of infinity (i.e. open or short circuit), to a value of between 1·5 and 2. Hence the importance can be seen of low loss connections between load and measuring equipment when measuring large standing wave ratios.

The effect of cable attenuation is to reduce large s.w.r.'s to much smaller values, a disadvantage when measurements are being made, but an advantage when connecting a generator to a variable load. The increased losses may be neglected for cables with a large initial attenuation due to the fact that the original mismatch is greatly reduced as we proceed from the load to the generator. The extra power loss on the cable is usually negligible except in a short length close to the load.

Appendix: Alternative Derivation of the Cable Attenuation Expression

The fraction of the *incident* power flowing into the line which is available at the receiving end is $1/n^2$; of this, a fraction k^2 is reflected back to the source. Thus the fraction of the power input which is absorbed in the receiver is

$$\frac{1-k^2}{n^2}$$

The fractional power reflected back to the *sending end* is

$$k^2/n^2$$

and the fraction reaching the *sending end* is

$$k^2/n^4$$

The power input at the sending end is:

$$1-(k^2/n^4)$$

The impedance the generator or source will see depends on the amount and phase angle of the reflected power, and this may determine the absolute magnitude of the power which can be delivered by the generator to the sending end of the cable.

Then

$$\frac{\text{Power in}}{\text{Power out}} = \frac{1-(k^2/n^4)}{1-(k^2/n^2)} = \frac{n^2-(k^2/n^2)}{1-k^2}$$

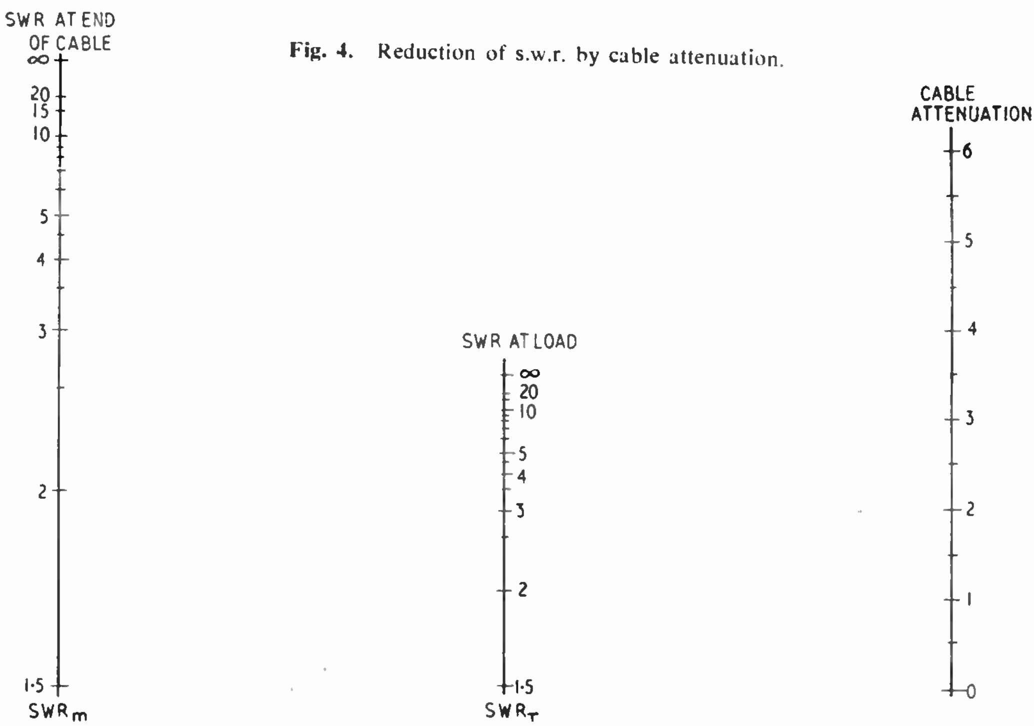


Fig. 4. Reduction of s.w.r. by cable attenuation.

SOME ASPECTS OF TRANSISTOR PROGRESS*

Report of the Discussion at an Institution meeting in London on September 26th, 1956.
In the chair: Rear Admiral Sir Philip Clarke, K.B.E., C.B., D.S.O.

N. J. Golden: Experiments have been carried out in the United States in which the common base alpha cut-off frequency was measured and compared to that predicted from equation:

$$\alpha = G \cosh \theta / (G \theta \coth \theta + Y_n)$$
$$\cong \operatorname{sech} [W(1 + j\omega\tau_p)^{\frac{1}{2}} / L_p]$$

The approximation is valid for alloy structure at moderate injection levels ($N_p \ll P_n$). When $\omega\tau_p \gg 1$ it was found that

$$f\alpha_{co} \cong \frac{\sqrt{6}\, D_p}{2\pi\, W^2}$$

followed from this equation. Careful metallographic section permitted measurement of W. Quantitative agreement between the theoretical expression and the experimental value for cut-off frequency was poor; the measured values being often factors of 2 or 3 lower. A qualitative theory was evolved to explain this.

If the alloy transistor geometry is considered wherein the emitter is made smaller than the collector in order to reduce the effects of surface recombination on the current gain, the hole flow pattern is more or less strictly isomorphic (for $W/L_p \ll 1$) to the electric field between two parallel conducting discs of different diameter whose centres are collinear. The "fringe field" then presents carrier paths much longer than W which will cut off at appropriately lower frequencies.

With regard to temperature limitations on transistor operation the main effect of lowering the ambient temperature is a decrease in current gain. This can be understood readily since the bulk lifetime decreases with temperature as predicted by Shockley and Reed.† Hence insofar as the current gain is controlled by bulk recombination, it would be expected to decrease.

Measurements of surface recombination velocity as a function of temperature with the present technique of surface preparation show S increasing as the temperature decreases. Hence the surface-controlled part of the current gain will also decrease.

Aside from this effect, in American experience operation from −55° C to +85° C has not shown any peculiarities.

G. I. Hitchcox (*Member*): In electrochemistry it is often necessary to measure very small currents, or small voltages generated in very high resistance sources. There is at present no wholly satisfactory electrometer which combines adequate electrical performance with stability, reliability, and reasonably low cost. Can Dr. Loeb suggest any possible application of transistors, or devices such as the fieldistor and the dielectric amplifier, to electrometry?

J. N. Barry: Undoubtedly one of the chief factors affecting the rather slow appearance of commercial transistorized equipment is that of economics. This is particularly important in the domestic broadcasting field, and is likely to apply for some time to come. In this connection it would be interesting to know whether this factor might be affected even more adversely in the design of some of the more recent high-frequency transistors mentioned in the paper. It would appear, at first sight, that the production costs of such types as the drift and diffused base transistor might, for some long time, be appreciably higher than the better established alloy junction types.

Although much work has been done on the production of experimental types having better high-frequency characteristics, the results achieved have sometimes entailed limitations in other directions. Thus I believe it is correct to say that the collector voltage rating of such types as the "well" and "surface barrier" transistor is relatively low, of the order of 4 volts. If this is so, would not these devices become relatively unattractive in many possible applications, for instance in video amplifiers for television use?

It is noted that no mention is made in the paper of any advance in the design of power transistors. In this field conflicting design requirements usually arise, e.g. the large

* H. W. Loeb, *J. Brit.I.R.E.*, **16**, pp. 515-528, September 1956.

† "The statistics of recombination," *Phys. Rev.*, **87**, p. 835, 1952.

collector junction area required for large power handling capacity gives rise to lower "α" cut-off frequencies and larger collector/base capacitances, thus reducing high-frequency performance of the device. Could the author give any information regarding the possible exploitation of some of the recent types mentioned in the paper (e.g. the diffused base transistor) for power applications at high frequencies. Such a device, if feasible, would be particularly attractive if not unduly restricted on peak voltage rating.

In the concluding section of the paper some comparisons are made between the use of transistors in switching circuits and domestic entertainment equipment. I feel that the conclusions drawn may be somewhat misleading, for the following reasons.

In the switching field the use of transistors offers a number of additional advantages when compared to the broadcast receiver field. The most important of these are:—

(*a*) Comparatively large tolerances on transistor parameters can be acommodated.

(*b*) When acting as a switch, power dissipation in the transistor is very low.

(*c*) For many purposes the speed of operation of currently available transistors is adequate.

(*d*) There is usually a more favourable economic balance in this field due to the very suitable characteristics of transistors for switching purposes.

In addition, present junction transistors will perform more of the functions of cold cathode tubes at a comparable cost, but have the considerable advantage of an operating speed at least an order higher. Present experimental work indicates that in fully electronic telephone exchanges of the future transistors are likely to be employed to the exclusion of cold cathode tubes.

W. C. R. Withers: I notice that the author does not mention the development of photo-transistors. I wonder if he could say what limits their response to the low audio frequency range and whether in view of their high sensitivity and small size they are likely to outdate conventional photo-voltaic and photo-emissive types.

L. Nelson Jones (*Graduate*): Could Dr. Loeb please comment on the use of *n-p-n* junctions, as so far he has only spoken of *p-n-p* types for the improved types of construction described.

As the greater mobility of electrons in the *p*-type base of *n-p-n* transistors shortens the transit time compared to corresponding *p-n-p* types I am still a little puzzled at the relative rarity of the *n-p-n*. One particular advantage of the *n-p-n* at present is that its supply voltage polarity is the same as for the thermionic valves, but as transistors become of wider use this advantage will reduce.

J. J. Robinson (*Associate Member*): To an ever-increasing extent I am concerned with the replacement of manual operations by electronic methods in civil aviation applications which at the same time call for improved dependability in order to meet the twin requirements of maximum speed with maximum safety. In this respect I am hoping the transistor will make a valuable contribution particularly in switching networks. Has the author any substantial data on the question of transistor reliability both as regards those already marketed and those still undergoing laboratory tests?

F. Oakes (*Associate Member*): The *p-n-i-p* transistor has been said to provide great advantages over other transistor types, with respect to performance as well as suitability for large-scale production. Does the recently developed diffused base transistor provide even greater advantages, thus rendering the *p-n-i-p* type obsolete?

Wing Commander E. C. Seeley (*Member*): Will Dr. Loeb please tell us whether, in the course of his work, he has encountered anything significant in regard to temperature, moisture, or pressure effects which may impede the progress of development or production of transistors.

The object of my question is to ascertain whether transistors were likely to be more or less troublesome than thermionic valves under tropical, arctic and high altitude conditions.

This question is one which is causing much concern in aircraft radio circles. Cooling, heating and pressurization can be fairly easily achieved in large passenger aircraft but in military aircraft the problem is not simple. In order to keep cooling, heating and pressurization systems to minimum weight their efforts are generally confined to the aircraft cockpits. The essential size of the cockpit dictates that many radio systems must be relegated to other parts of the fuselage. The less radio components are affected by temperature and pressure changes, the less is the problem of ensuring reliability.

M. Venner: Many of us have experienced the difficulties of stabilizing transistor circuits to cope with increases in ambient temperatures (I_{co}, etc.). I have often wondered whether, given that some circuits can be adequately cooled, there is a minimum temperature below which transistors cannot be used.

The specifications issued by manufacturers sometimes give a bottom limit but I feel that in these cases they are stating "our transistors will work at any temperature in this range" and not "our transistors do not work at all below this temperature."

It has been stated that below $-55°$ C the current gain (α) falls off so much that the transistor is virtually useless. This is the cause of the limited temperature range rather than that at these low temperatures the connections to the semi-conductor wafer fell off!

G. Roman: The limit arising from one process for the temperature below which a transistor would cease working could be estimated by calculating the temperature at which the number of ionized impurities has fallen to a small fraction of its room-temperature level.

The ionization energy of antimony in silicon is about 0·04 eV, in germanium 0·026 eV. It can easily be shown that the temperatures at which only 5 per cent. of the total number of impurities are ionized are 150° K for silicon, 100° K for germanium.

The state of 5 per cent. ionization was assumed quite arbitrarily and therefore may be disputed. It seems, however, that it is an ultimate limit especially if we remember that long before this temperature is reached some phenomena associated with ionization will prohibit the use of the transistors because of circuit consideration. (Increase of base resistance, increase of Early effect, decrease of the mobility of carriers etc.)

P. G. Briggs (*Graduate*): A pair of junction transistors had been left in an oven where the temperature was well in excess of 200° C. Of course they ceased to operate as transistors at this temperature, they would conduct a current but not operate as an active element. However, when these transistors were cooled to room temperature they could be again operated as good transistors.

F. C. Lye (*Graduate*): What is the effect of irradiation on transistors, particularly with regard to their use in nucleonic equipment? Considerable work is no doubt being carried out on this aspect. The question is asked in view of several recent papers on the subject of changes of crystal structure particularly in dielectrics when exposed to irradiation but I have found no reference to the change of crystal structure in semi-conductors. This seems a very important field of research particularly with regard to the use of transistors in such equipments as Geiger counters and radiation monitors.

AUTHOR'S REPLY

Dr. H. W. Loeb: In reply to questions which referred to specific advantages and disadvantages of particular transistor types it should be stressed that developments have reached a stage where, for a considerable number of important application fields, several types of structure have been proved feasible. Except possibly in the case of the low-frequency alloyed junction triode, insufficient manufacturing experience has been accumulated as yet to enable an economic assessment to be carried out with any degree of confidence. For example, if a medium frequency (alpha cut-off frequency ~ 3 Mc/s), low power, low voltage transistor is required, almost all the types referred to would be functionally adequate, but it is probable that in the end one or two structures will become preferred types—on economic grounds. Hence such questions as whether the diffused base transistor makes the swept intrinsic types obsolete, whether *n-p-n* types will gain ground at the expense of *p-n-p*'s, or even whether the transistor will completely replace the cold cathode tube in low-speed counting circuits will be decided by the relative costs of manufacture. For the various manufacturing techniques proposed so far, at any rate, differences in complexity between the processes pertaining to particular types will matter less than the degree of control which can be achieved and the overall process yield. Until transistor types which have so far been made under laboratory conditions only go into mass production it is virtually impossible to assess their economies of manufacture to the order of accuracy required for such comparisons. It is quite correct to say that, particularly in the earlier designs, higher operating techniques were achieved at the cost of a restriction to low operating voltages. In the later structures which incorporate some form of drift base or swept intrinsic layer, the opposite is true, i.e., rather higher collector voltages are necessary before the optimum frequency response is obtained. Generally, higher frequency of operation is coupled with lower limits for the maximum power dissipation. Here, also, the diffused base type of structure is likely to widen the range over which compromises can be achieved.

The development of power transistors can be traced out in a manner similar to that applied to the h.f. transistor to indicate how design theory and fabrication methods progressed together. Apart from the problem of providing a sufficiently low thermal resistance between the collector junction and a heat sink to prevent the junction temperator from exceeding a safe value, which is in the region of 80° C for germanium transistors, means had to be found for reducing the current dependence of the transistor small signal parameters.

Temperature limitations can arise from two causes. The first is the cessation of the device to work as a transistor at all, which may be brought about by mechanical changes (melting of alloying material, e.g. indium which melts at 155° C, failure of solders, etc.) or, as pointed out during the discussion, by the temperature dependence of some physical process involved in transistor action like ionization of impurity atoms, or change from extrinsic to intrinsic conduction. The second limitation arises from the temperature dependence of the various transistor parameters and is to some extent relative since it will depend upon a particular application and upon the steps taken to minimize the effects of such changes—for example, the use of compensating diode circuits. As was pointed out during the discussion, it has been established that satisfactory operation of germanium alloyed junction transistors is possible over a temperature range from −55° C to +85° C.

Other ambient influences such as pressure and humidity can be excluded by suitable encapsulation of the device.

A number of reliability estimates have been published which indicate that the failure rate even of point-contact transistors is very low, of the order of 0·1 per cent. per 1000 hours. For junction types figures of 0·05–0·01 per cent. have been quoted. As regards such estimates it must be remembered that we are still at a somewhat early stage for adequate assessments which will have to be based upon a long-term analysis of a large statistical population, and that as production methods improve, reliability will increase further. Even the figures already obtained compare favourably with those for ordinary vacuum tubes.

Junction photocells will undoubtedly benefit from the developments in transistor technology. At present their outputs are higher by an order of magnitude than those of vacuum photocells, on the other hand junction capacitance and diffusion effects combine to limit the frequency capabilities; typical values for frequencies at which response begins to fall are 100–200 kc/s.

Finally, as regards d.c. amplification, both direct coupled and "chopper"-type amplifiers have been described. Extensive compensation against drift occasioned by temperature changes has to be used in these applications and input impedances are generally low, i.e., the amplifiers are essentially current amplifiers. It is conceivable that future electrometer applications may make use of other solid-state amplifying principles which are more easily adapted to the necessary high impedance input requirements.

IMPLICATIONS OF PHASE PRECOMPENSATION IN A TELEVISION TRANSMITTER ON THE SHAPE OF THE RADIATED SIGNAL*

by

A. van Weel, Dr.Techn.Sc.†

SUMMARY

The smears after black-to-white transitions due to phase errors in the receiver can be compensated by an overshoot introduced by a phase-precompensating network in the video-frequency section of the transmitter. In a negative-modulation system this overshoot can only be accommodated in the available modulation space by raising the maximum-white level from 10 per cent. to at least 20 per cent. which is equivalent to a loss of transmitted signal power of 30 per cent. Curvature of the modulation characteristic makes a complete compensation impossible.

With a positive-modulation system, the output stage of the transmitter should be capable of delivering pulses, reaching to 22 per cent. over and above the maximum-white level.

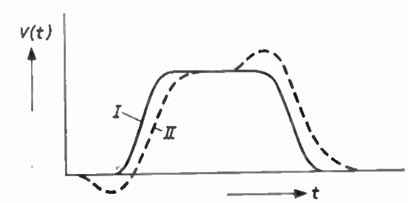
1. Introduction

Both the vestigial-sideband filter of a television transmitter and the selective intermediate-frequency amplifier of a television receiver introduce pronounced phase non-linearities into the signal path.[1-5] The phase distortion caused by the vestigial-sideband filter of the transmitter is considered to be inevitable, and has therefore to be compensated in the video-frequency section of the transmitter. As regards phase distortion caused by the receiver, two possible solutions exist, one solution being precompensation in the video-frequency section of the transmitter[3]; the other is the design of a receiver which is inherently free from phase distortion.[5]

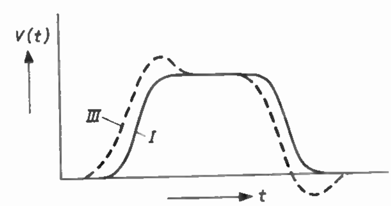
The influence of this kind of phase distortion on the signal shape is shown qualitatively in Fig. 1 (*a*), in which curves I and II represent the responses to a square wave of networks with and without phase distortion respectively. This effect causes smears after a sharp black-to-white transition in the television picture.

To compensate for this distortion, the signal fed to the modulator stage has to be distorted in the opposite direction by a phase-equalizing network. The output voltage of such a network will therefore have a shape as depicted in Fig. 1 (*b*), curve III. The rounding-off of the trailing edge of curve II is clearly compensated by the overshoot of curve III; this does, however, not imply that curves II and III are exactly symmetrical with respect to curve I. Fig. 1 only aims at giving a general qualitative picture of the signal shapes.

The signal according to curve III is modulated on the radio-frequency carrier in the output stage of the transmitter. This means that the output stage should be capable of delivering a peak signal equal to the nominal



(*a*)



(*b*)

Fig. 1. (*a*) Square wave with and without phase distortion (curves II and I respectively).
(*b*) Square wave with and without phase predistortion (curves III and I respectively).

* Manuscript received 22nd December 1956. (Paper No. 387.)
† Philips Research Laboratories, N. V. Philips' Gloeilampenfabrieken, Eindhoven, Netherlands.
U.D.C. No. 621.397.62.

maximum level plus the overshoot. On the other hand, the lowest carrier level should be sufficiently high to accommodate the undershoot in the trailing edge of the square wave.

In this paper we will calculate the amount of this overshoot and discuss the consequences for the radiated signal as well as for the demands put on the transmitter output stage.

2. Calculation of the Overshoot caused by the Phase-precompensating Network.

Data on phase-precompensating networks, as far as the author knows, have only been published for the former N.W.D.R. transmitters in Germany. The curve, which is obtained by adding the group-delay curves of a vestigial-sideband filter and a receiver, and which, accordingly, determines the phase characteristic of the precompensating network in the video-frequency section of the transmitter, was

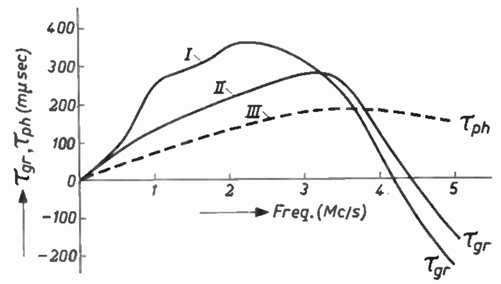


Fig. 2. Group-delay characteristics of phase-pre-compensating networks used in N.W.D.R. television transmissions (curves I and II). Characteristic III gives the phase delay corresponding to curve II.

reproduced in Fig. 8 of a previous paper by the author[4], having appeared originally in a N.W.D.R. publication.[6] The group-delay characteristic of the corresponding pre-compensating network is given in Fig. 2 as curve I. More recent data were put at the author's disposal by N.W.D.R. engineers during joint experiments with the Hanover television transmitter in February 1956. Curve II in Fig. 2 is the group-delay characteristic of the correction network according to these data. Obviously, this correction holds for a combination of a vestigial-sideband filter and receiver with different properties as compared with the combination for which curve I was deduced. The calculations in this article are based on curve II.

The phase-delay characteristic (curve III) of the correcting network was derived from the group-delay curve by graphical integration. In order to make it easy to work out the corresponding transient distortion, we used Fig. 3 to transform the phase-delay distortion

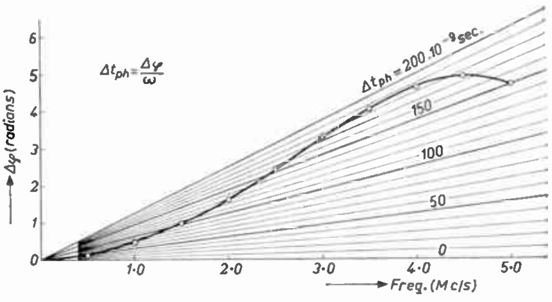


Fig. 3. Graph for deriving the phase-angle distortion characteristic corresponding to curve III of Fig. 2.

curve into a phase-angle distortion curve. Fig. 3 consists of a pencil of straight lines, each representing the phase angle versus frequency characteristics for the case of a constant phase delay. The small circles in the figure are values transferred from curve III in Fig. 2; the resulting curve thus indicates the phase shift occurring at different frequencies.

However, we are not interested in the *absolute* phase shift, but only in departures from a linear phase versus frequency characteristic; the angle between such a straight characteristic and the frequency axis merely gives a certain shift of the zero point of the

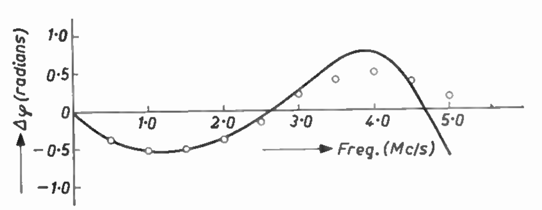


Fig. 4. Phase-angle distortion characteristic corresponding to the phase-delay characteristic III in Fig. 2. (The small circles give the sinusoidal function $\Delta\varphi = -0{\cdot}55 \sin \omega\tau$, with $\tau = 185 \times 10^{-9}$ sec.)

time axis. We chose this reference characteristic for $\Delta t_{ph} = 160 \times 10^{-9}$ sec; the maximum positive and negative deviations of the given phase-delay curve with respect to this straight line are roughly equal and the (absolute) magnitude of the largest deviation is therefore minimum. The resulting phase-angle distortion characteristic is given by the heavy line in Fig. 4.

A reasonable approximation to this phase-angle distortion curve is provided by a sine function, which is indicated by the small circles in Fig. 4. The transient distortion corresponding to this sinusoidal phase distortion can easily be calculated because in this case the maximum phase deviations amount to not more than 0·5 radian, allowing the approximations $\sin \Delta\varphi = \Delta\varphi$ and $\cos \varphi = 1$ to be used. The transient distortion is given in the lower part of Fig. 5 for a frequency band extending to $f_m = 5$ Mc/s.

Curve II in the upper part of Fig. 5 represents a sine-integral function, which is the response to a step-function signal of a distortion-free low-pass filter with the same bandwidth. Curve III, which is the sum of curves I and II, gives the response of the predistorting network to a sine-integral input signal. The overshoot amounts to 40 per cent. in this case.

If the amplitude characteristic of the video-frequency channel does not fall abruptly to zero at the maximum frequency, but extends to higher frequencies with a gradual fall-off, the overshoot of the input signal can be smaller or even be nil. In such a case we should be obliged, when working out the transient distortion curve, to take into account the properties of both the phase-correction network and the video-frequency channel in the frequency band above 5 Mc/s. This would involve data of the system that are uncertain; it could only give a fine structure on the transient response shown in Fig. 5, because the amplitudes of components of higher frequencies are very small. Therefore we confine ourselves to the distortion curve I of Fig. 5 and conclude that the amount of overshoot, present in the output signal of the phase-precompensating network is at least 32 per cent.

It may be mentioned that the transient response of a phase-precompensating network in accordance with the group-delay curve I in Fig. 2 would show a substantially larger amount of overshoot.

3. Implications for a Television Signal with Negative Modulation

We will firstly discuss the implications of the overshoot for a television signal complying with the so-called "Gerber" standard (the European 625-line standard), because the calculations were based on a phase-precompensating filter designed for this standard. Afterwards a television standard with positive modulation will be considered.

In Fig. 6 (*a*), the shape of the signal under the Gerber standard is depicted; it can be seen that in this negative-modulation system the maximum-white level is at 10 per cent. of the full modulated signal. This 10 per cent. is a minimum value, fixed with a view to the possibility of intercarrier-sound reception; a higher percentage for the maximum-white level is not forbidden in the Gerber standard, although it obviously decreases the modulation space and therefore the signal-to-noise ratio.

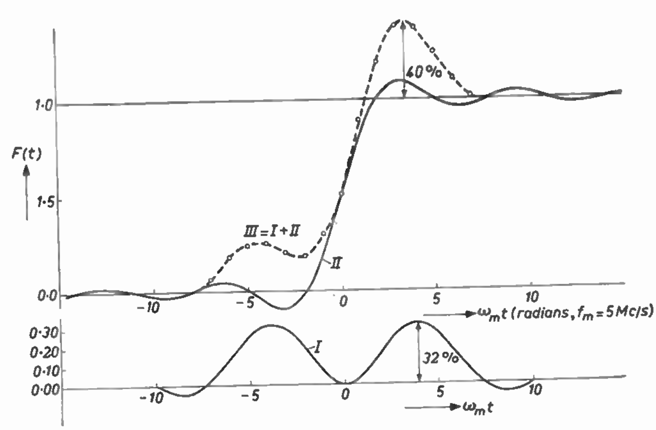
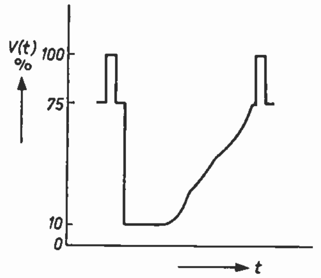
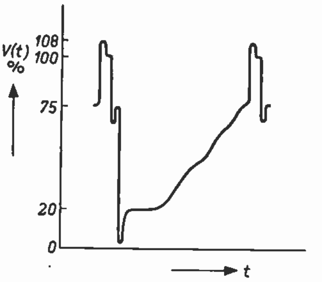


Fig. 5. Transient distortion caused by the phase-precompensation (curve I). Sine-integral function (curve II). Distorted transient response (curve III = I + II).



(*a*) Undistorted signal shape; maximum-white level at 10%.



(*b*) Shape of signal that has undergone predistortion to compensate phase errors; maximum-white level at 20%.

Fig. 6. Television signal with negative modulation.

With the maximum-white level at 10 per cent., the maximum amount of overshoot that can be modulated in the picture signal is $(10/65) \times 100 = 15$ per cent. From this figure it follows that with a phase-precompensated signal the maximum-white level must be substantially higher than 10 per cent.; as a matter of fact, it has to be at $(32/132) \times 75 = 18$ per cent. at least. However, modulation down to absolute zero level meets with practical difficulties which we will discuss in more detail in the next section. For the moment we take these difficulties into account by an upwards rounding-off of the percentage calculated, and conclude that phase-precompensation in the transmitter for the phase errors of the receiver entails a maximum-white level of at least 20 per cent.

As mentioned before, phase-precompensation is in any case necessary for the phase errors of the vestigial-sideband filter of the transmitter. The magnitude of these phase errors is roughly equal to the magnitude of the phase errors of the receiver. Consequently the transient response of the phase-equalizing network in the case where only the phase errors of the sideband filter are compensated will have an overshoot of about 16 per cent. Such a signal could just be modulated on a carrier with a maximum-white level at 10 per cent.; however for the same reason as above, the maximum-white level will have to be at 12·5 per cent. at lowest.

The modulation space for the picture signal proper amounts to $75 - 20 = 55$ per cent. and $75 - 12{\cdot}5 = 62{\cdot}5$ per cent. in these two cases. We may conclude, therefore, that compensation of the receiver's phase errors in the transmitter entails a lowering of the signal-to-noise ratio with a factor of $62{\cdot}5/55$ in amplitude, i.e. one decibel. Formulating this result a little differently, we may say that with phase-precompensation, the power of the transmitter has to be increased by at least 30 per cent. to achieve the same signal-to-noise ratio. A consequence for the design of receivers based on phase-precompensation is that the amplification of the video-frequency section has to be increased by 15 per cent.

Of course, overshoot also occurs after upwards transitions, which means that short overshoot pulses will be present on top of the synchronization signals. This overshoot will amount to $(32/100) \times 25 = 8$ per cent. in the case of full precompensation, and to 4 per cent. if only the phase errors of the sideband filter are compensated.

The shape of the signal at the output of the phase-precompensating network is shown in Fig. 6 (*b*) for the case of full compensation. We draw attention to the fact that this is the signal shape at the input of the modulator stage. The signal at the detector in the receiver is again normal without overshoot or smears.

4. Influence of a Curved Characteristic Modulator

The modulator characteristic of a television transmitter is usually rather non-linear as a consequence of the necessity to use grid modulation. An example of such a modulator characteristic is given in Fig. 7 (full line). The non-linearity can be counteracted by inserting a gamma-correcting device in series with the modulator stage proper; the dotted line in Fig. 7 shows the influence of such a gamma correction.

In order to illustrate the consequences of the curvature of the modulator characteristic for the phase-compensation problem, two modulating step-function signals, each with 32 per cent. undershoot, are given in the lower left-hand part of this figure. The signal I corresponds to a black-to-maximum-white transition, with the maximum-white level chosen so as to coincide with 20 per cent. r.f. output voltage. The signal shape of the output voltage can be constructed from the modulator characteristic; curve I′ gives the envelope of the output signal derived from the dotted modulator curve, thus taking some gamma correction into account. The relative undershoot of this envelope amounts to 12/55 = 22 per cent., which is 10 per cent. less than the prescribed value. Consequently a smear of 10 per cent. will be present in the v.f. output signal of the receiver.

Curves II and II′ of Fig. 7 give corresponding results for a maximum-white level at 25 per cent.; the relative undershoot of the envelope of the output voltage is in this case 13/50 = 26 per cent., which still falls 6 per cent. short with respect to the nominal value of 32 per cent. With smaller step-function signals, the phase compensation will be better; however, the smears caused by phase distortion are most annoying with the largest transitions, i.e. from black to maximum-white, and for these transitions the compensation is only partial.

We thus conclude that, due to the curvature of the modulator characteristic, a complete compensation of the phase errors of vestigial-sideband filter and receiver cannot be achieved in the v.f. section of the transmitter even if the maximum-white level is chosen at 20 per cent. of the peak output voltage.

5. Implications for a Television Signal with Positive Modulation

We have no data available on phase-equalizing networks for systems with positive modulation, and can therefore only extrapolate the results of the preceding sections. Fig. 8 (*a*) gives the standardized signal shape. In order to be able to accommodate an overshoot of 32 per cent., the bottom of the synchronization signals would have to be not lower than $(32/132) \times 30 = 7$ per cent. To avoid modulation down to absolute zero, this bottom level has to be raised to 9 or 10 per cent. At the maximum-white level side, the output stage of

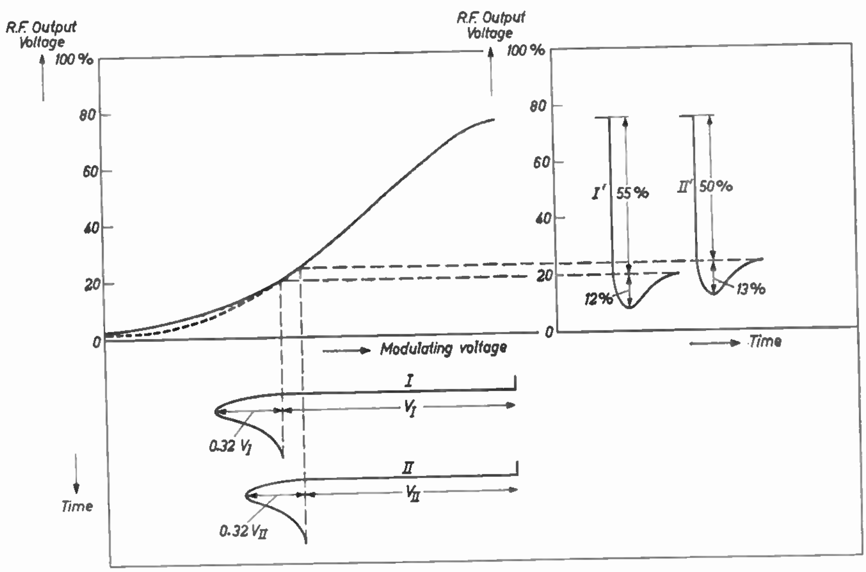


Fig. 7. Modulation characteristics of a television transmitter with and without gamma correction (full line and broken line respectively). Modulating signals (I, II) and envelope of output signals (I′, II′) for the maximum-white level of 20% (I, I′) and 25% (II, II′) respectively.

(*a*) Undistorted signal shape; bottom of synchronization pulses at 2%.

(*b*) Shape of signal that has undergone predistortion to compensate phase errors; bottom of synchronization pulses at 9%.

Fig. 8. Television signal with positive modulation.

the transmitter must be able to deliver an overshoot peak of $0 \cdot 32 \times 70 = 22$ per cent. above the nominal 100 per cent. level. The modified signal is depicted in Fig. 8 (*b*). For the case, where the phase-precompensating network only compensates for the phase errors of the vestigial-sideband filter of the transmitter, the corresponding figures are 6 per cent. and 11 per cent. respectively.

6. References

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. . . Radio Engineering Overseas

A selection of abstracts from European and Commonwealth journals received in the Library of the Institution. All papers are in the language of the country of origin of the journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

621.317.34

Investigations on helical lines.—K. Habener. *Nachrichtentechnische Zeitschrift,* **9**, pp. 581-584, December 1956.

The effect of inhomogeneities in helical lines is investigated. Inhomogeneities such as kinks, bends and variations in pitch are treated as lossless or lossy quadripoles. The equivalent four-terminal network circuits are derived from measurements and the reflection coefficients are calculated.

621.317.789.029.64: 621.372.832.43

A microwave power monitor.—J. Swift. *Proceedings of the Institution of Radio Engineers, Australia,* **17**, pp. 424-428, December 1956.

The coaxial power monitor described in this paper consists basically of a directional coupler and a detector. The instrument operates from about 300 to 4,000 Mc/s and in addition to facilitating the adjustment of microwave power sources, provides a rapid and simple means of measuring wide ranges of r.f. power. An r.f. load may be rapidly matched to a transmission line if the power monitor is used in reverse and with minor alterations the device can also be used as a wideband attenuator.

621.372.2

The gyrator, an electric network element. B. D. H. Tellegen. *Philips' Technical Review,* **18**, pp. 120-124, October 1956.

The author considers the electric networks having one or more terminal pairs that can be built up from conventional network elements, namely resistors, coils and capacitors. The set of relations between terminal voltages and currents determined by such a network has properties that correspond to the linear, constant and passive nature of such networks and, furthermore, has the property of reciprocity. It has been demonstrated that, conversely, these properties are sufficient for any set of relations possessing them to be realizable by a network. The writer reverses this line of reasoning and poses the question: What (ideal) network elements must be introduced in order to make it possible to realize all linear constant passive systems. There are no grounds for including reciprocity amongst the properties imposed. It is then shown that, apart from the conventional elements—resistor, coil, capacitor and the ideal transformer—a new network element has to be introduced; this new element has been christened the ideal gyrator. Networks containing the new element generally lack the property of reciprocity. A brief sketch is given of some of the properties of the gyrator.

621.372.43

Negative impedances produced with transistors and their treatment in conjunction with the network synthesis.—H. Ebel. *Nachrichtentechnische Zeitschrift,* **9**, pp. 513-519, November 1956.

Networks for producing linear negative impedances and their properties are discussed. The use of transistors in amplifier circuits with feed-back permits a simpler design and leads to the design of technically simple impedance converters. The compensation of the loss in a filter by compensating the resistive losses in its coils is given as an example. A second example is a description of a line amplifier with two negative impedances having transmission properties similar to a symmetrical "go and return' repeater.

621.372.54/55

Filters and delay-equalizers for television transmission on cables.—H. Keil. *Nachrichtentechnische Zeitschrift,* **9**, pp. 469-475, October 1956.

The single-sideband transmission of television signals on coaxial cables offers a number of problems to the filter designer. These are: the suppression of one sideband, the design of filters for carrier-frequencies and adjustable equalizers and the construction of filters for the separation of the power supply from the television signal. Solutions for these problems are discussed. By using potential analogue methods and punch-card computers, the considerable work involved in designing delay equalizers for terminal and intermediate stations can be performed in a relatively short time.

621.375.4

The design of wideband amplifiers using transistors.—G. Meyer-Brotz and F. Felle. *Nachrichtentechnische Zeitschrift,* **9**, pp. 498-503, November 1956.

Formulae are stated which permit the calculation of the whole frequency response for the absolute value of amplification in a multiple stage amplifier. The approximation of the exact frequency responses of the four-terminal network parameters by means of simple circle diagrams produces approximate agreement between calculations and measurements. However, the trend of the gain response and its relationship to the data and the operating points of the transistors can be seen easily.

621.385.1.032.269.1

High "perveance" electron guns.—R. Hechtel. *Archiv der Elektrischen Ubertragung,* **10**, pp. 535-540, December 1956.

After defining the concept of "perveance" the paper points out the significance of this quantity for the generation of very short electromagnetic waves. Subsequently, the various ways of approach towards electron guns of high perveance are delineated. The type of electron guns mostly in favour today is that devised by Pierce. Pierce's simple theory which yields useful results only for small values of perveance up to approximately $1\mu A/V^{3/2}$ was amplified and applied to an electron gun having a perveance of $5\mu A/V^{3/2}$. The calculated characteristics of this gun could be verified with high accuracy by experiments.

621.385.16

Spatial harmonics in electron beams.—R. MULLER. *Archiv der Elektrischen Ubertragung*, **10**, pp. 505-511, December 1956.

It is well known that the propagation in periodic structures cannot be described by a single wave, but only by a sum of waves with different phase velocities, the spatial harmonics. The same argument is valid for the space-charge-waves in electron beams with periodic structure. The periodicity in an electron beam can be caused by a special shape of the electron path (mäander line, cycloid) or by periodic variation of the d.c.-conditions (e.g., velocity jumps). The consideration of space harmonics in electron beams leads to a common treatment of some known microwave amplifiers as well as to new types of amplifiers and oscillators.

621.385.832

Current form in electromagnetic deflecting coils. G. LUKATELA. *Elektrotehniski Vestnik*, **24**, pp. 271-275, September/October 1956.

The author treats the problem of linearity of the time base in terms of the linearity of the magnetizing current. The equivalent circuit of the electromagnetic deflecting coil with the amplifier belonging to it is analysed by Laplace transforms and the explanations are completed by quantitative data. At the end the equation is deduced, which makes it possible to achieve the linearity of the magnetizing current in the deflecting coil.

621.385.832:621.317.351

A blue-screen c.r.t. as a new equipment for recording non-periodic responses.—W. DIETRICH. *Nachrichtentechnische Zeitschrift*, **9**, pp. 504-507, November 1956.

A special blue-screen c.r.t. is used for recording single events containing frequency components of up to approximately 15 kc/s. The electron beam leaves a black trace on the screen which may be preserved for any length of time or may be erased within a few seconds. This tube is particularly suitable for recording non-periodic responses. The circuit and the operation of an oscilloscope fitted with such a valve is described. Some examples for application are explained.

621.385.832 : 681.142

The technology of electrostatic cathode ray storage tubes. F. CHOFFART. *Onde Electrique*, **36**, pp. 815-821, October 1956.

The construction and manufacture of guns, and their performance is described. The method of deposition of the dielectric of the target, the manufacture of fine structure grids having a mesh of 20 to the mm, and the assembly of targets are also discussed. A description is given of the general assembly: gun, target, glass envelope and the precautions taken.

621.394/6:620.16

Experiences gained from vibration tests.—G. WEINMANN and A. HOLZ. *Nachrichtentechnische Zeitschrift*, **9**, pp. 584-589, December 1956.

Communication equipments contain a large number of joints which frequently have a bad contact or unwanted contact with other parts, even after careful optical inspection. Intermittent faults are particularly difficult to localize. Various organizations are aiming at an improvement of reliability in their communication links and have started large scale investigations and developed test methods for detecting faulty contacts during the early stage in production. The paper describes the magnitude of the problems and the experience gained from low-frequency and carrier frequency equipment. Test methods are also reported.

621.395.625.3

The development of a miniature battery-operated tape recorder.—W. R. NICHOLAS and A. D. HILDYARD. *Proceedings of the Institution of Radio Engineers, Australia*, **17**, pp. 367-372, November 1956.

The widespread use of magnetic recording in the production of broadcast programmes has produced a demand for a small self contained battery-operated tape recorder. The general requirements for such a recorder are discussed, followed by an outline of the development of both the mechanical and electrical design of a suitable machine which employs a spring driven motor.

621.395.625: 681.84.081.48

A magnetodynamic gramophone pick-up. II. Frequency characteristics. N. WITTENBERG. *Philips Technical Review*, **18**, pp. 173-178, November 1956.

The author puts forward a number of arguments for extending the frequency response to above 15 kc/s. In the course of the discussion on the frequency characteristic for the upper register, a method is described of separating static and dynamic tracing losses, these being a wave-length and a frequency effect respectively. With records of synthetic resin, static tracing loss exhibits a cut-off frequency of about 24 kc/s where the groove has a diameter of 24 cm and of about 12 kc/s for the inner-most groove; dynamic tracing loss, on the other hand, is connected with a groove/needle resonance at about 18 kc/s. Mention is also made of reproduction at very low frequency, the effect of the electrical load on the pick-up, and non-linear effects.

621.397.813

Frequency dependent graduation equalization of television signals.—H. SCHONFELDER. *Archiv der Elektrischen Ubertragung*, **10**, pp. 512-526, December 1956.

An electrical equalization of graduation errors is ordinarily linked with a raising of the "noise" level in the dark portions of the image. Since on the other hand with a visual judgment of the received picture, linearity of the graduation is of importance for large areas only, there is a possibility of reducing greatly this increase in picture noise for the dark portions of the picture by equalizing the graduation merely for the low frequencies. Signal distortion appearing with such a frequency-dependent graduation equalization can be considered as visually acceptable with a suitable choice of the cut-off frequency (approx. 3 Mc/s, i.e., that frequency above which the equalization of the graduation is practically nil), while the signal-to-noise ratio in the dark portions is improved 1.7 and 2.4 times, respectively (with a noise spectrum emphasizing the high frequencies). It must be decided on the merits of each case, however, whether with use of the method for making disturbance less perceptible the image errors which are likely to occur can be neglected.