A MEMBER'S RESPONSIBILITY

IN relation to total membership only a comparatively few members have opportunity to take active office in the Institution. Every member, however, on his election, undertakes to support the objects of the Institution, and one of these purposes is, of course, to contribute to the advancement of knowledge. This may be done in several ways, of which the most important is to stimulate thought and to share ideas at meetings, and to present papers or topics for discussion.

The arrangements for the presentation of papers or discussions at Institution meetings are now shared by the Programme and Papers Committee, the local Section Committees, and the Specialized Group Committees. Broadly speaking, the main Programme Committee endeavours to cater for the general and the widest range of interest; the local Section Committees bear in mind principally the need of members in their own areas, although for obvious reasons, papers presented are not necessarily connected with local manufacturing interests; then as their names indicate, the Group Committees endeavour to see that the specialized interests of groups of members are also included in Session meetings.

In this latter connection, four Specialized Groups are at present in operation, covering Electronics in Biological Research Computers, Radar and Navigational Aids and Audio Frequency Engineering. Other Groups will include Television Engineering.

Recent Institution meeting arrangements have included a number of half-day symposiums; these have enjoyed wide appeal and further symposiums have therefore been planned for next Session. So far arrangements have been made for a Symposium on Frequency Synthesis Techniques to be held on 25th May, and in the Autumn there will be a meeting on Instrumentation in Nuclear Power Stations, and another on Component Reliability.

These Symposiums give opportunity for deliberating and discussing a subject from a much broader view and might perhaps be classed as miniature Conventions. Their success has prompted some local Sections to adopt the same practice during their programmes.

In the task of planning meetings and obtaining papers, the various Committees rely largely on the help and co-operation of members. Offers of papers are always welcomed, and it is particularly helpful if such offers can be accompanied by a synopsis of the proposed contribution. Many members may not be able personally to submit a paper, but their co-operation is cordially invited in suggesting authoritative people qualified to write a paper, or submitting ideas on a subject which should be included in a programme. Bearing in mind the ever-widening scope of radio and electronics, it is especially felt that senior members of the Institution may do much to encourage members of their staff to submit papers.

Every member, wherever he may be situated, is therefore invited to give support to those Committees already mentioned by contributing or sponsoring a paper. The responsibility of members in this connection was well expressed in an earlier editorial (September 1949).

"Members cannot better serve the Institution than by contributing to radio progress through the pages of the *Journal*. Scientific work is never complete until it has formed the subject of a paper. On such activity all our future depends; it is the life blood of the practising engineer and requires donation from many, rather than from the few."

G. D. C.

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OBITUARY

It is with considerable regret that the Council records the death of **James Nelson,** one of the founders of the Institution, and the third President. Soon after the first world war Mr. Nelson was a founder member of the Radio Association. This society did not meet the ideals of many engineers, and in a letter circulated to the technical press in 1925 he suggested "that a really technical society is wanted in Britain." In the letter, published in *Experimental Wireless and the Wireless Engineer* (August 1925) it was stated:

"It is intended that the Society shall be called The British Institute of Radio Engineers".

The editorial view of *Experimental Wireless* stated, "If this is to be a British rather than an international institution, why use the international 'radio' instead of the British 'wireless'? We would sooner see it called The Institution of Wireless Engineers, or The Wireless Institution."

This recommendation appeared to have been accepted by the Interim Committee which was appointed, and confirmed at a meeting held at the Hotel Russell, London, on 31st October, 1925, Mr. Nelson being elected Chairman.

Thereafter Mr. Nelson was actively associated with the Institution in various offices. He served as Chairman until 1931 when he was elected a Vice-President; five years later he became President of the Institution. In 1942 Mr. Nelson was elected an Honorary Member.

For much of his professional career Mr. Nelson was associated with British Insulated Cables Ltd. During the 1914-18 war he served as an officer in the R.N.V.R. and in 1930 he joined the Central Electricity Authority as transmission engineer for the Central England area. In 1945 he returned to B.I.C.C. with whom he remained until his retirement in 1953.

Mr. Nelson died in Weston-super-Mare on the 31st January last in his eightieth year.

The Council has also learned with regret of the deaths of the following members and has sent messages of sympathy to their relatives.

Frederick Lonsdale Channell had been concerned with the manufacturing side of the radio industry since shortly after the 1914-18 war. He held various appointments on the production side of the industry until 1946, when he went to South Africa and set up in business in this field. During 1958 his health deteriorated and he returned to England early last year; he died on 20th May at the age of 63 years. Mr. Channell was elected an Associate in 1933 and was transferred to Associate Member in 1941.

Alexander Mackintosh Cowie, who died on 29th May last at the age of 62, commenced his technical career in the Royal Navy in 1912. He was awarded the Distinguished Service Medal in 1919 for wireless services during the war, and subsequently held appointments with the Royal Indian Marine and with the Government of Iraq as Wireless Supervisor. In 1937 Mr. Cowie resigned to found a private radio and television college in Glasgow. He was elected an Associate Member in 1938.

Peter Gareth Lovell died suddenly on 30th October, 1959, at the age of 33 years after an attack of coronary thrombosis. After earlier experience as technical assistant with Philips Electrical Ltd., Mr. Lovell joined the Plessey Company in 1952 as a project engineer, and was subsequently promoted to group leader, a position he held at the time of his death. First registered as a Student in 1953, Mr. Lovell completed the Graduateship Examination in that year, gaining the Audio Frequency Engineering Prize. He was transferred to Associate Membership in January 1957.

Herbert Owen Sampson was in charge of technical operations at the B.B.C.'s London television studios. He had been with the Corporation since 1936; recently he had been concerned with experimental work on electronic control of lighting at the new television centre. Mr. Sampson, who was elected an Associate Member in 1958, was 44 years of age at the time of his death on 25th May, 1959.

Microwave Television Mobile Relay for Outside Broadcasting†

by

J. POLONSKY, I.E.G., E.S.E.R.‡

A paper read on 4th July 1959 during the Institution's Convention in Cambridge.

Summary : A brief account is given of the principal qualities required in mobile links. These include transmission of picture and sound without degradation of the quality and stability in time of the technical performances, and ease of operation as regards transport, installation, monitoring and maintenance. The essential causes of distortion introduced in the transmission by a microwave link are reviewed and the problem of cross-talk between the picture and the sound channels and the transmission of a colour television programme are dealt with in some detail. A short description is given of an equipment operating in the band 6400-6900 Mc/s.

1. Introduction

The increasing length of television programmes, together with the growing interest of viewers in the day's events presented on direct transmission, have urged television producers to make increasing use of outside broadcasting. The importance of mobile relays, whose role is to provide a microwave connection between outside broadcast vans and the central Control Room, has grown in consequence.

In addition to permanent microwave links, which form the nervous system of the national network of studios and television transmitters, mobile relays generally provide short connections of at most a few tens of miles, in the case of normal broadcasts, and up to 50 to 150 miles in the case of special broadcasts from points relatively distant from a permanent link.

But it should be pointed out that even for so-called short range (from 5 to 30 miles) one or two intermediate repeaters may be used in order to ensure adequate clearance of the first Fresnel zone. This case is a frequent one, especially in large cities.

Another current application of mobile microwave relays consists in setting up temporary connections over long distances extending up to 300 miles and more. Such links are quite frequently used in countries where the installation of the permanent link system is behind the development of the transmitter network.

In the course of this paper, mobile links will be considered mainly as means of connection between an o.b. van and the central control. In conclusion, a few words will be said on the use of these equipments for long distance connections.

Reporting points are often somewhat distant from high fidelity cables for the transmission of the audio signal. It has therefore been found useful in mobile relays to transmit the picture and audio signals by multiplexing.

2. Essential Qualities Required from a Mobile Microwave Relay

Taking into account the special conditions of utilization, it would seem useful to analyse the qualities of this equipment from the following points of view:

(a) Technical performance (tolerance on the degradation of the signals transmitted).

(b) Ease of use as regards: transport, installation, monitoring and maintenance.

(c) Reliability.

2.1. *Technical Performance*

The C.C.I.R. has prepared recommendations for permanent long distance television microwave links, and tolerances for a distance of 1,500 miles have been defined. For short distance there exist at the present day only the

specifications prepared by national administrations. Furthermore, these specifications develop rapidly with technical advances made in this field and with the more and more stringent tolerances imposed by the needs of exploitation; in particular, the arrival of colour television may be mentioned.

In any case, these relays are part of the video equipment preceding the central control, and it is therefore right to expect from them negligible degradation of quality with respect to that allowed for long distance microwave links. It can be asserted that in the present state of technique this requirement can be fully met.

As an illustration, Table 1 is given below, which indicates the essential specifications for the picture and audio signals in the microwave link type TM-110B—which is the subject of this paper—these being given in comparison with the C.C.I.R. tolerances for long distance links. It should be noted that C.C.I.R. recommendations concern only links transmitting video signals in monochrome without audio.

2.2. *Ease of Use*

In cases where visibility is assured between the o.b. van and the receiving aerial at the central control, the transmitter alone is mobile; the receiver can be installed as a fixed station with an orientatable aerial. In a case where intermediate links are found to be necessary, the transmitter and the receiver have to be transportable and capable of installation under somewhat precarious conditions.

The ease of use of the mobile relay will be examined from the following points of view: transport of the equipment, its installation. monitoring and maintenance.

Table 1 : Specifications for Television Links

	Tolerances for a TM-110B Relay	C.C.I.R. Tolerances for Long Distances†
LUMINANCE CHANNEL:		
Video frequency response ...	6 Mc/s $\pm$ 0.5 db 10 Mc/s $\begin{array}{l}+0.5 \text{ db}\\-3 \text{ db}\end{array}$	6 Mc/s $\pm$ 3.5 db 10 Mc/s $\begin{array}{l}+1 \text{ db}\\-6 \text{ db}\end{array}$
Video time delay	6 Mc/s $\pm$ 5 mμsec 10 Mc/s $\pm$ 50mμsec	6 Mc/s $\pm$ 400 mμsec 10 Mc/s $\pm$ 500 mμsec
Differential gain	0.5 db	2 db
Variation of synchronizing signal	<5%	+10 to –30%
Low-frequency distortion ...	<3%	$\pm$10%
Rise time	60 to 70 mμsec	100 to 140 mμsec
Overshoot	<4%	20%
Signal/noise ratio for 30 miles ...	6 Mc/s 54 db 10 Mc/s 47 db	6 Mc/s 41 db 10 Mc/s 37 db
Audio cross-talk in video channel	>60 db	not specified
CHROMINANCE CHANNEL:		
Differential phase at 4 Mc/s ...	<1°	not specified
Differential gain at 4 Mc/s ...	<0.5 db	,,
AUDIO CHANNEL:		
Audio frequency response ...	40 – 10,000 c/s $\pm$ 1 db 10 – 14,000 c/s $\begin{array}{l}+1 \text{ db}\\-3 \text{ db}\end{array}$	not specified
Audio distortion	<1%	,,
Audio noise level	60 db	,,

† C.C.I.R. Warsaw 1956.

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2.2.1. Transport

It is necessary to separate the transmitter from the receiver and to design each in the form of packs of low weight and small size; the aerial and its support have to be light, small and easily dismantled.

The equipment has to be rugged and must not demand special precautions for its transport.

2.2.2. Installation

Particular attention has to be given to the design of the head, as well as of the part which has to be operated without shelter. A judicious compromise has to be made between conditions which are often mutually contradictory, and more especially :

(a) lightness and ease of handling on the one hand, solidity and rainproofness on the other,

(b) ease and speed of erection and orientation of the aerial and its mechanical stability,

(c) speedy starting up of the complete installation and verification of its correct performance with a minimum of external equipment.

With this in mind, it is useful to have available in the transmitter section of the equipment internal test signals for the picture and for the sound, and monitoring gear at the transmitter and at the receiver for verifying that the equipment is in proper order.

2.2.3. Monitoring in operation

A simple video monitor with built-in oscilloscope, a head-set or loud speaker for sound, must be available for easy monitoring of the operation, both at the transmitter and at the receiver.

It is very useful to have available at the transmitter a means of monitoring the picture detected at s.h.f. and for comparing it against the transmitter input signal.

Taking into account the small amount of space in an o.b. van, the number of monitoring and measuring equipments has to be kept at a minimum, for bringing the link into operation and for monitoring.

2.3. *Reliability*

For countries which have only one television programme, an interruption in the relay between the o.b. van and the Central Control causes a break in the transmission of the programme over the whole of the national network and sometimes even of the international system, as in the case of Eurovision. In order to avoid such troubles, it is usual to duplicate the normal relay by a stand-by link. Nevertheless, it frequently happens that transmissions have to be made without any stand-by equipment. The degree of reliability of operation of a microwave link is not measured by the frequency of the breakdowns but more particularly by the stability in time of its technical performance.

The equipment has therefore to be carefully designed with this in mind. Negative feedback for stabilizing the performance of the equipment is to be employed wherever possible. Components have to be of high reliability. In order to facilitate maintenance or servicing of the equipment it has to be readily accessible, and monitoring is to be so designed as to permit easy location of a fault.

3. Principal Distortions introduced in the Signals by Microwave Links

3.1. *Functional Block Diagram of a Link*

The block diagram of a microwave relay operating on s.h.f. and transmitting by multiplexing the picture and audio signals is generally as shown in Fig. 1.

The transmitter consists essentially of :

(1) a video amplifier,

(2) a picture/sound multiplex-unit generally composed of a 6 to 12 Mc/s oscillator, frequency-modulated by the audio channel and added to the picture signal.

(3) a reflex klystron of about 100 mW to 1 W, operating as an s.h.f. oscillator, frequency modulated by the complete video signal (picture signal combined with the audio sub-carrier),

(4) a directional aerial.

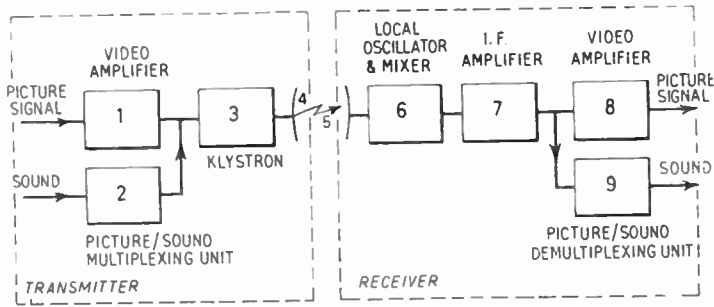The receiver consists essentially of :

(5) a directional aerial,

Fig. 1. Block diagram of a microwave relay.

(6) a s.h.f. local oscillator (reflex klystron) and—a crystal mixer coupled to the receiving aerial and to the local oscillator,

(7) a wide-band i.f. amplifier, with limiter and discriminator,

(8) a video amplifier for the picture channel.

(9) a picture/audio demultiplier consisting of a channel separator, and a sound channel (amplifier, limiter and discriminator of the sub-carrier and a.f. amplifier).

After a brief consideration of the main distortions introduced in the various parts of the relay, the distortion due to the i.f. part of the receiver will be gone into in greater detail. In particular, consideration will be given to the problem of cross-talk between picture and audio channels, as well as to the effects of variation of the differential gain and of the differential phase on the chrominance signal, in the case of the transmission of colour television.

3.2. *Nature of Distortions Introduced by the Various Parts of the Mobile Relay*

(1) *In the video amplifiers* at the transmitter and at the receiver, attention must be given more especially to:

(a) good transmission of the d.c. component and of the low video frequencies, or their correct restitution,

(b) protection of the signal from hum and noise, more particularly at points where the video level is low,

(c) reduction of phase distortions caused

more particularly by the rejector circuit of the sound sub-carrier at the receiver,

(d) stability of the frequency and phase response.

(2) *The non-linearity of the frequency modulation characteristics of the klystron* is liable to introduce the following defects: compression or expansion of the synchronizing level, which varies with the average shade of the picture, distortion of the gamma of the luminance signal, cross-talk between the picture and audio channels, and lastly, differential gain on the chrominance signal.

(3) *In the s.h.f. section of the transmitter,* mismatching of the s.h.f. components causes phase distortion. This problem is particularly important when the transmitter klystron is at a long distance from the aerial, through a waveguide and without a ferrite isolator.

(4) *In the receiver s.h.f. section and in the mixer,* any mismatch increases the noise figure.

(5) *The i.f. part of the receiver* is the main seat of distortions, which are in particular:

(a) distortion of the video frequency response,

(b) distortion of the time delay at the higher frequencies of the video spectrum,

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(c) differential gain and differential phase in the chrominance channel.

(d) cross-talk between the picture and audio channels.

(6) *In the i.f. discriminator,* the main defects are of the same kind as those met with in the transmitter klystron, in particular: distortion of the synchronising signals and of the gamma, introduction of differential gain and intermodulation between picture and sound channels.

It should be noted that in sections of the equipment handling only the audio channel, attention should be directed to obtaining a good performance at the receiver second limiter (in the sound sub-carrier spectrum), in order to eliminate any spurious amplitude modulation of the signal caused by the picture channel, before the audio signal is demodulated by the discriminator.

3.3. *Examination of the Distortions Introduced in the I.F. Section of the Receiver*[1]

3.3.1. Nature of i.f. distortions

The main defects introduced in the i.f. section, enumerated in section 3.2.5. are closely related to the imperfections of the i.f. amplifier, and in particular to:

distortion of the i.f. response,

i.f. time delay distortion,

imperfections in the i.f. limiter.

An ideal i.f. amplifier should have a constant frequency response, of the maximally flat type, in a spectrum not less than

$$2F_{\max \text{ (video)}} + \Delta F_{\max \text{ (picture + sound)}}$$

corrected by all-pass network for equalizing the i.f. time delay. ($\Delta F_{\max}$ is the sum of the frequency deviations by the picture and sound signals.)

This amplifier should be followed by a perfect limiter having a short time-constant with respect to the maximum video spectrum frequency to be transmitted. For a modulation index greater than 1, the effects resulting from these three causes of distortion, enumerated above, are difficult to separate. But, for a low modulation index, which is the case for the higher frequencies of the video spectrum for the chrominance sub-carrier and for the audio sub-

carrier, it is possible with a fair approximation by a vectorial analysis to determine the effects produced by each one of these causes of i.f. distortion after discrimination.

3.3.2. Effects of distortion of the i.f. response (phase distortion assumed to have been corrected)[2]

Figure 2 shows two typical frequency responses of a receiver i.f. amplifier. Curve 1 (full line) is of the "maximally flat" type: curve 2 (dotted line) corresponds to a small time-delay distortion curve.
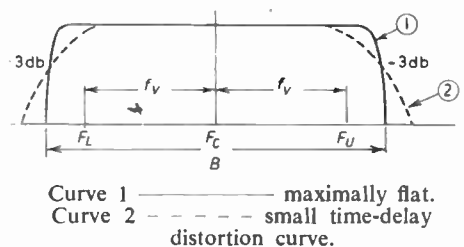

Curve 1 ——————— maximally flat.
Curve 2 – – – – – small time-delay distortion curve.

Fig. 2. Typical frequency responses of a receiver i.f. amplifier.

Let F_c represent the intermediate frequency corresponding to the carrier signal,

F_U the i.f. corresponding to the upper modulation band,

F_L the i.f. corresponding to the lower modulation band,

m_f the modulation index in the transmitter.

Let G_c, G_U, and G_L be the i.f. gains corresponding respectively to F_c, F_U and F_L.

For a low modulation index, m_f does not exceed 0·3, and it can be shown that the distortion of the i.f. response (curve 2, Fig. 2):

(a) causes a change in the modulation index which becomes:

$$m'_f = \frac{G_U + G_L}{2G_c} . m_f$$

and (b) introduces a spurious amplitude modulation:

$$m'_a = \frac{G_U - G_L}{2G_c} . m_f$$

Figure 3 is a vectorial representation of the composite i.f. signal. Both symmetrical sidebands U and L are shown in Fig. 3 (a); they can be resolved into M_{FM}, 90 deg. out of phase

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and M_{AM} in phase with the carrier. The latter corresponds to a spurious amplitude modulation and can be resolved into two vectors U'' and L'', while the vector M_{FM} represents the resultant of frequency-modulated sidebands equal and symmetrical with respect to the vector 90 deg. out of phase with the carrier (Fig. 3 (b)).
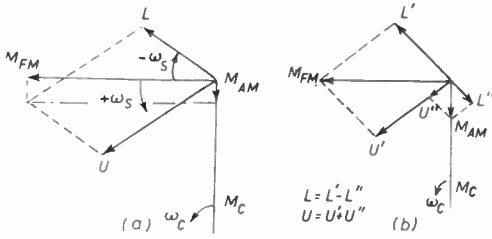

Fig. 3. Vectorial representation of composite i.f. signal.

The i.f. limiter is capable of eliminating the amplitude modulation, while the change in the modulation index causes, after discrimination, distortions of the video frequency response and causes a differential gain α_1. (The latter appears through the effect of the variation of the value of F_c depending on the luminance signal.)

The differential gain can be calculated by

$$\alpha_1 = 20 \log_{10} \Delta \left| \frac{G_{\max} + G_{\min}}{2G_c} \right| \frac{F_{c\ \max}}{F_{c\ \min}}$$

3.3.3. Effects of the variation of the i.f. time delay (the frequency response is assumed to be constant).

It can be shown that the variations of the i.f. time delay can affect:

(a) the video frequency response.

(b) the phase of the detected video signal,

(c) the differential gain and the differential phase in the chrominance signal,
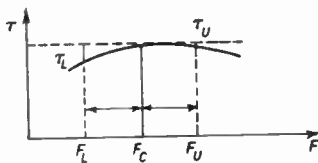
(d) the cross-talk between the picture and sound channels.

Figure 4 shows a typical i.f. time delay curve $\tau = f\ (F)$ (the time delay for F_c being taken as reference).

According to whether the curve of Figure 4 is symmetrical or not with respect to F_c, the effects of variation of the time delay on the demodulated signal will be different.

(1) Case of a symmetrical curve (Fig. 5 (a))

Let us assume $\tau_L = \tau_U$; we have $\theta_L = \theta_U = \theta$, and the amplitude M_1 at the receiver becomes $M_1 = M_0 \cos \theta$.

The variation of τ introduces a phase shift θ on the demodulated video signal, which is variable with the video frequency. There results a time delay on the demodulated video signal.

But, there is neither a variation of the modulation index nor any spurious amplitude modulation.

(2) General case of an asymmetrical time delay distortion curve with respect to the carrier (Fig. 5 (b))

In this case, the following defects appear:

(a) a phase shift, after demodulation, which varies with the video frequency: $\theta = \frac{1}{2} (\theta_U + \theta_L)$, which appears as a distortion of the time delay curve. θ_U and θ_L are referring to θ_C.

(b) a change in the modulation index from m_f to m''_f:
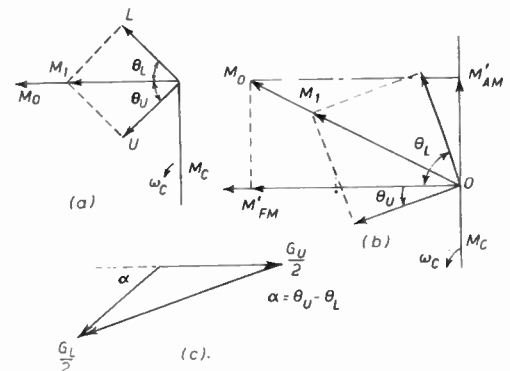
$$m''_f = m_f \cos \frac{1}{2} (\theta_U - \theta_L)$$

Fig. 5. Vectorial representation of effect of time-delay on demodulated signal for

(a) symmetrical time-delay curve;

(b) asymmetrical time-delay curve.

(c) i.f. amplitude and phase distortion.

After demodulation, this fact causes distortion of the amplitude-frequency curve and a differential gain if F_c moves along the i.f. characteristic α_2.

$$\alpha_2 = 20 \log_{10} \Delta \left| \cos \frac{\theta_U + \theta_L}{2} \right| \begin{matrix} F_{c \ max} \\ F_{c \ min} \end{matrix}$$

(c) a spurious amplitude modulation:

$$m''_a = m_f \sin \tfrac{1}{2} (\theta_U - \theta_L)$$

This can be eliminated by the i.f. limiters.

(d) a differential phase $\Delta\varphi$ of the video frequency signals producing the two sidebands F_U and F_L, the value of $\theta = \tfrac{1}{2}(\theta_U + \theta_L)$ varying with the position of F_c i.e., with the amplitude of the luminance signal.

$$\Delta\varphi = \Delta \left| \frac{\theta_U + \theta_L}{2} \right| \begin{matrix} F_{c \ max} \\ F_{c \ min} \end{matrix}$$

Differential gain and differential phase are particularly troublesome for the transmission of colour television pictures in the N.T.S.C. system. The combined effects of i.f. amplitude and phase distortion can be graphically represented by Fig. 5(c). Picture intermodulation on sound is mainly due to the differential phase of the sound sub-carrier as a function of the position of F_c.

Intermodulation can be roughly expressed by:

$$\frac{S}{I} = \left(\frac{\Delta f}{f} \right) \cdot \frac{57°}{\Delta\varphi_{sc}}$$

where S is the nominal audio level after demodulation

I is the disturbing picture signal in the spectrum from 30 to 15000 c/s

$\Delta f/f$ is the a.f. modulation index for 100 per cent. sound modulation at a.f. reference frequency

$\Delta\varphi_{sc}$ is the differential phase of the sound sub-carrier.

(For example:

for $\dfrac{\Delta f}{f} = 100$ and $\Delta\varphi_{sc} = 6°$,

$\dfrac{S}{I} = 1000$, or 60 db)

The introduction of a large amount of pre-emphasis in microwave links improves notably the quality of the transmission of colour television pictures and reduces cross-modulation between the video and audio channels.

3.4. *Defects Introduced by the Limiters*

The i.f. limiters act as non-linear amplifiers. Their role is to eliminate the spurious amplitude modulations before the discriminators.

However, as mentioned in Section 3.2, it would be wrong to think that the limiters can eliminate all the defects introduced by the variation in the i.f. response.

Without going more deeply into details, it can be said that imperfect limiters generally introduce the following defects:

(a) a reduction, but not a suppression, of the spurious signal amplitude modulation,

(b) a limitation which decreases as the video frequency increases, taking into account the time constants of the limiter circuits; this effect is particularly marked on the sound sub-carrier situated at the end of the video spectrum. This fact is one of the main causes of cross-talk between picture and sound.

3.5. *Conclusion*

In order to obtain a high quality i.f. amplifier, it is necessary to provide:

(1) Either a curve meeting the " maximally flat " requirements, corrected by a time-delay equalizing network,

or (2), a frequency response, as shown by curve 2 of Fig. 2, with a low time-delay distortion.

The second solution is more economical and can be applied in links with demodulated repeaters, as in the case of mobile links. In this case, it is possible to effect a partial video correction after detection.

In the case of repeaters without video demodulation, curve 1 of Fig. 2 is recommended in preference.

(3) The differential gain and phase on the chrominance sub-carrier and cross-talk between picture and sound can be kept sufficiently small if, for the picture channel, a pre-emphasis circuit is used at the transmitter, ensuring a substantial reduction of the frequency deviation for low-frequency luminance signals (as now recommended by the C.C.I.R.).
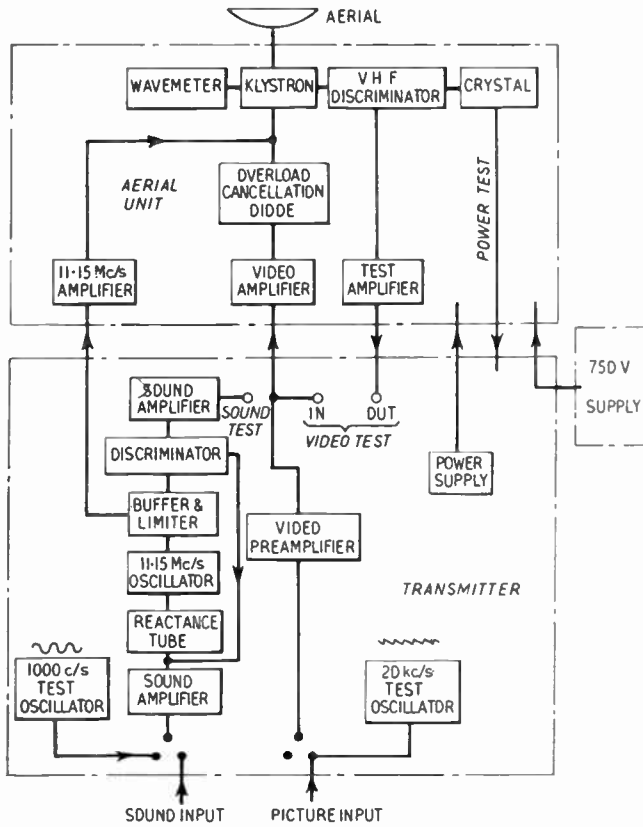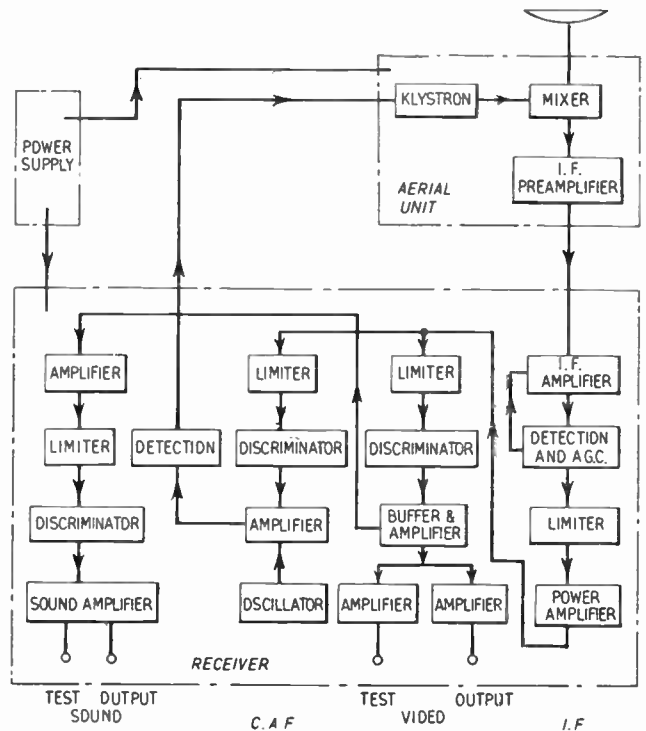
Fig. 6. (above)
Transmitter block diagram.

Fig. 7. (right)
Receiver block diagram.

Fig. 8. (above)
Transmitter receiver ; aerial, head, transmitter or receiver unit, and power supply unit.

(4) It is most important to reduce as much as possible the frequency excursion of the luminance signal along the i.f. curve, caused either by the variation of the mean value of the picture signal, or by frequency instabilities in the transmitter and receiver klystrons. An automatic frequency control in the receiver allows this problem to be solved. It is referred either to the mean value of the video signal in the case of repeaters with pre-emphasis. or, preferably, to the peak value of the synchronizing signals in the case of repeaters using the nominal frequency deviation of the luminance signal.

(5) It is also strongly recommended that the i.f. level be kept constant at the limiter inputs. in spite of field fluctuations. This is the role of the i.f. automatic gain control.

4. Brief Description of the Mobile Relay Type TM-110B

4.1. *Origin and Principal Characteristics*[3]

The original equipment, predecessor of the TM-110B, known at the time as TM-110, was designed in 1953 and was used for the first time in Eurovision for the coronation of H.M. Queen Elizabeth II and on a larger scale in 1954 for the transmission over the Eurovision Network of the International Football Championships. To-day, about 250 links of this type are in operation in many countries.

This link operates in the band 6400 to 6900 Mc/s and is generally used for mobile o.b.; it is also used, however, for temporary or semipermanent long distance links, which may reach up to 400 or even 600 miles (for instance. the Stockholm-Göteborg-Malmöe and the Prague-Warsaw links).

Link TM-110B differs from the TM-110 chiefly in that it meets the standards laid down for this type of equipment for colour television on the N.T.S.C. system.

The specification of the TM-110B relay is summarized in Table 1.

4.2. *Brief circuit description*

Figures 6 and 7 show respectively the block diagram of the transmitter and of the receiver. It is not intended to give in this paper a detailed description of the link. A few precautions taken in the design of the TM-110B will, however, be briefly indicated.

(1) As far as possible, circuits are used in which negative feed-back is introduced in the video and audio i.f. circuits as well as in the power supplies, in order to improve performance and especially stability with time.

(2) The receiver a.f.c. and a.g.c. circuits have received special attention so that the i.f. amplifier and limiter signals operate in the optimum ranges of frequency and levels.

(3) The s.h.f. components have been designed for small values of the v.s.w.r., in order that the klystron modulation characteristics shall remain linear and that the receiver noise figure shall be small.

(4) A high value has been chosen for the audio 11·15 Mc/s—so that the sub-carrier rejection circuits shall introduce practically no phase distortion in the upper part of the video spectrum of the picture signal. The relay is thus capable of transmitting with practically no phase distortion a signal of 7 to 8 Mc/s. It can also transmit a 10 Mc/s bandwidth by the addition of a phase corrector.

(5) A pre-emphasis of 11 db is applied at the transmitter to improve the performance of the differential gain and differential phase in the chrominance signal, as well as the crosstalk between the picture and audio channels.

(6) The a.f.c. is effective over a range of 15 Mc/s each side of the carrier and the a.g.c.

Table 2
Principal Characteristics of TM-110B Relay

S.H.F. band: 6400 to 6900 Mc/s

Transmitter power: 1 watt

Aerial gain and beam width at half-power points:

aperture 1 m: $G = \simeq 35$ db; $\pm 3°$

aperture 2 m: $G = \simeq 41$ db; $\pm 1° \ 30'$

I.f. response: 28 Mc/s at 3 db points about 115 Mc/s.

Receiver noise figure: less than 15 db

Nominal s.h.f. deviation by the picture signal: ± 4 Mc/s with pre-emphasis.

Nominal s.h.f. deviation by the sound subcarrier: ± 1 Mc/s

Video band: 10 Mc/s

Sound sub-carrier: 11·15 Mc/s

keeps constant the input voltage to the limiter for fading of about 25 db.

4.3. *Engineering Design* (*Figs.* 8, 9, 10 *and* 11)

In order to meet the requirements detailed in section 2.2. both transmitter and receiver are divided into three units: two are under cover close to the operator, while the third is sealed and associated with the aerial.

In order to ease maintenance and any servicing that might be required, the components and, in particular, the number of valves in the aerial unit have been cut down to a minimum. The cable interconnecting the head and the transmitter or receiver can extend to some 160 yards. The arrangement in the units has been designed to ensure full accessibility of all the component parts.

The aerial normally used consists of an off-set parabolic reflector of perforated sheet metal 1·15 metre in diameter; it is quickly and rigidly secured to the supporting unit by means of screws and knurled nuts. The transmitter or receiver r.f. heads are housed in weatherproof cases. The base carries an arrangement consisting of a transparent sight for rapid orienta-

tion. The sealed cover is fitted with a perspex window which passes the energy radiated or received by the horn located on the axis of the paraboloid. An inspection window which can be hermetically sealed provides access to the control and monitoring panel of the aerial unit. This window is chiefly used when the parabolic dish is being orientated. During normal operation, all these controls are brought, by means of a change-over switch, to the units kept under cover (in the o.b. van for instance). The unit and its aerial are carried on a tripod, each leg of which rests, through a ball joint, on a removable rubber sole-plate. The height of the tripod is adjustable and it carries an elevation and azimuth orientation device graduated in degrees, provided with clamping means. An incorporated spirit level facilitates the verification of the reference plane and so avoids the possibility of errors during orientation from maps; this is particularly useful in the case of the transmitter.

The receiving parabolic dish can be orientated by a single operator, means for monitoring the receiver grid current and sound received on the head-set being close at hand on the receiving head.
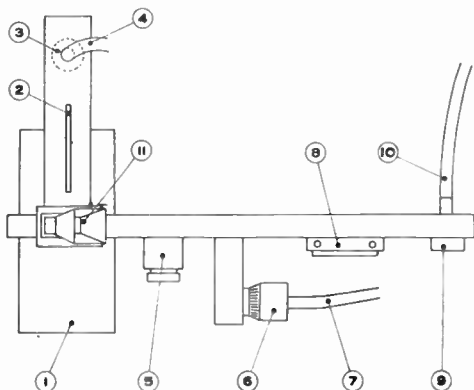

Fig. 9. (a) Transmitter microwave circuit.

Key:
1. Klystron.
2. Power test attenuator.
3. Power test crystal.
4. Power test output.
5. Test discriminator cavity.
6. Wavemeter.
7. Wavemeter control shaft.
8. Frequency deviation test attenuator.
9. Test discriminator crystal.
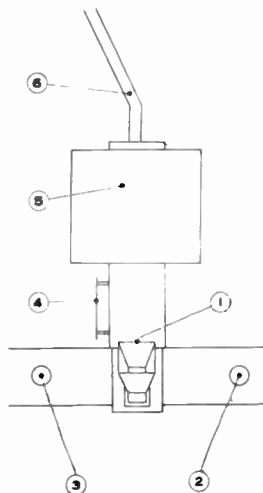10. Output toward test amplifier.
11. Horn.

Fig. 9. (b) Receiver microwave circuit.

Key:
1. Horn.
2. Mixer crystal.
3. Mixer crystal.
4. Local oscillator output control attenuator.
5. Local oscillator klystron unit.
6. Klystron cavity control.

It should be noted that the equipment includes a built-in interphone which can be connected to the telephone network, thus securing telephonic communication between the operators at the transmitting and receiving aerials. This arrangement is particularly useful for connections at extreme range, as well as for the better co-ordination of the installation and of starting up at the transmitter and at the receiver.

4.4. *Built-in Monitoring Arrangements*

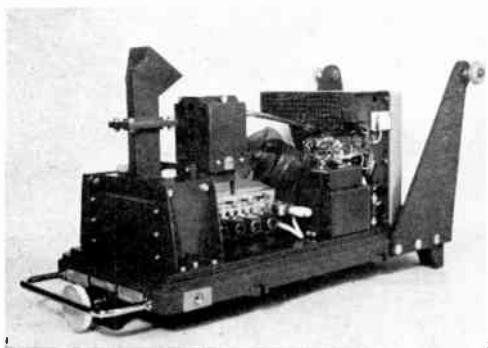In order to make the starting up of the link

Fig. 10. Receiver head.

easier (orientation and control of quality) the transmitter has been provided with internal test signals (saw-tooth at line frequency for the picture and a sinusoidal signal at 1000 c/s for the sound). By means of these calibrated signals, quick adjustments can be made, without the need of external generators.

An s.h.f. discriminator coupled to the waveguide and followed by a video amplifier provides easy control of the picture at the transmitter output. By means of a switch, the output picture can be compared with that at the input, on the same monitor. A s.h.f. wavemeter provides means for monitoring the transmitter frequency and the frequency deviation. In addition, the transmitter is provided with a s.h.f. power monitor.

In addition to the video and sound outputs, the receiver is provided with independent outputs for monitoring the picture and sound signals. Built-in meters are available for all the principal controls.

In order to verify that the installation is in proper order, as well as for purposes of control in operation, it is only necessary to have available at the transmitter and at the receiver a picture monitor with profile oscilloscope and a loudspeaker or head-set for sound. The same test equipment can be used for quick servicing. Naturally, for more thorough maintenance, a video signal generator, wobbulator, etc., are needed.

5. Conclusion

In this paper the author has endeavoured to present a brief analysis of a few of the technical
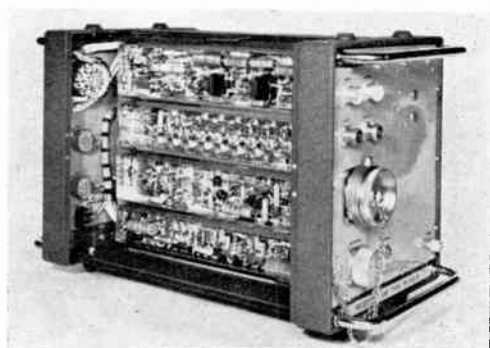

Fig. 11. Receiver unit.

and engineering problems, which arise in connection with the design and development of mobile television microwave relays. As frequently happens, the role of the design engineer is to find a judicious compromise between technical performance, ease of operation, reliability of operation and the initial cost price of the equipment.

The TM-110B relay which has been described by way of example has supplied much valuable information, particularly as regards to the possibilities of connecting in cascade a number of links, with successive demodulations for long distance (300 to 600 miles for instance).

This technique would appear to be of special interest in a case where the microwave beam has to transmit multiplex picture and audio signals. The same technique has been applied in the TF-120 link for long-distance and permanent connections.[4] In this equipment tolerances have been markedly tightened up compared to those acceptable for the TM-110.

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This applies more particularly to video amplifiers, the modulation curve of the transmitter klystrons and the i.f. discriminators. Against this, the i.f. part in demodulating repeaters raises less problems than in the case of repeaters operating on the frequency converter technique without demodulation.

A few words should be said regarding the problems of transistorization of mobile microwave links. Advances made in the field of semi-conductors constantly raise the following problem for the engineer: When should a radio equipment be transistorized, and to what extent? This question does not always find an easy answer.

Our experience has shown that in the range of professional television equipment, transistorization has to be applied after some careful thought. Total transistorization offers great advantages; on the other hand, partial transistorization should be looked upon with a certain amount of suspicion; the resulting advantage in weight and dimensions is frequently slight; and there is risk of loss of quality and reliability.

6. Acknowledgments

7. References

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2. M. M. McWhorter and J. M. Pettit, " The design of stagger-tuned double-tuned amplifiers for arbitrarily large bandwidth," *Proc. Inst. Radio Engrs,* **43**, pp. 923-31, August 1955.
3. J. Polonsky, " Light-weight equipment for radio relays on centimetric wavelengths," *Onde Electrique*, **33**, pp. 932-9, November 1954.
4. J. Polonsky and E. Safa, " Large capacity microwave relays for the 7000 Mc/s band." *Onde Electrique*. **35**, pp. 986-1003. November 1957.

# of current interest . . .

Orkney Television Station

The B.B.C.'s Orkney television station, at Netherbutton, went on to full-power service on 17th December when a permanent installation replaced the temporary low-power installation which was brought into operation a year ago. The station operates on the same frequencies as before (Channel 5: vision 66·75 Mc/s, sound 63·25 Mc/s) and the transmissions are vertically polarized. It uses a directional aerial providing an effective power of from 4 to 14 kilowatts.

The new Orkney station extends the B.B.C.'s Television Service to the whole of the Orkney Islands and to a large part of Caithness, including Thurso, an area which includes over 40,000 people. Work on the permanent installation at the B.B.C.'s television station at Thrumster, near Wick, is progressing and this station will be on full-power operation on 1st March.

The Higher Award in Technology

Detailed arrangements for the award of "Member of the College of Technologists" (M.C.T.) have now been published by the National Council for Technological Awards. This award, which was announced in the April 1959 issue of the *Journal,* follows the Diploma in Technology and is given to those who complete a substantial programme of work of value to industry carried on jointly in industry and a college. Lord Hives, Chairman of the National Council and of the Court of Governors of The College of Technologists,* has stated that it intended that the new award should provide a qualification in the field of technology equivalent to the Ph.D. degree of a university.

The principal features of the award as set out in a memorandum issued for the guidance of applicants and of colleges and industrial organizations concerned are:

(1) A programme of work must be satisfactorily carried out jointly in industry and at a college.
(2) The programme may be concerned with any technological aspect of industrial activity. Examples are—research, development, design and production.
(3) The programme must be one which is likely to result in a useful contribution to technological knowledge. It must also require industrial experience and academic study over a substantial period.
(4) The period of study which The College of Technologists has in mind is three years, provided that a high proportion of this time is spent on work *acceptable* for Membership of the College.

A high standard is demanded for candidates for Membership of The College. Those who may register include holders of the Dip. Tech. and others with equivalent qualifications.

M.Sc. Course in Information Engineering

Suitably qualified students (preferably with some industrial experience) can qualify for the degree of M.Sc. on satisfactory completion of a 12-months' course in the Electrical Engineering Department at the University of Birmingham. The next session runs from 1st October, 1960 to 30th September, 1961, and subjects available cover Communications, Radar, Computers and Control Systems, with some degree of choice to suit individual requirements. Engineers who can be seconded from industry for a period of five weeks only can sometimes be admitted to courses of lectures on individual topics.

Conference on "Solid State Microwave Amplifiers"

The Electronics and Radio Spectroscopy Groups of the Institute of Physics are arranging a Conference on the above subject to take place at the University of Nottingham from 6th-8th April, 1960.

Aspects to be covered include:
(a) Masers.
(b) Parametric amplifiers.
(c) Related resonance studies.

Accommodation will be available in one of the University Halls of Residence. Further particulars and application forms are available from The Secretary, The Institute of Physics, 47 Belgrave Square, London, S.W.1.

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* Since this announcement was made Lord Hives has retired and has been succeeded by Sir Harold Roxbee Cox, who has been vice-chairman of the National Council since its foundation in 1955.

GRADUATESHIP EXAMINATION—NOVEMBER 1959—PASS LISTS

This list contains the results of *all* successful candidates in the November Graduateship Examination. A total of 343 candidates entered for the Examination which was held at sixty-five centres.

In November 1959 the examination was held for the first time under the new entry regulations wherein the examination was divided into two separate sections, Section A and Section B. Candidates are required to pass all parts of the Section, or those from which they are not exempt, at one sitting. All candidates entering for Section B had previously qualified for exemption or had passed Section A.

LIST 1 : The following candidates have now completed the Graduateship Examination and thus qualify for transfer or election to Graduateship or a higher grade of membership.

ABERCROMBIE, Frank Leonard (S) *Weston-super-Mare.*
ABRAHAM, V. T. (S) *Kottayam, Kerala State.*
AHMAD, S. Amin Uddin *Rawalpindi.*
AKINYEMI, Isaac Ola (S) *London, N.4.*

BILSBOROUGH, Gordon *St. Albans.*
BOWEN, Joseph Alfred Edward (S) *Crawley, Sussex.*
BROOKS, Peter John (S) *Hamilton, Ontario.*
BUNTING, Derek Henry Stanley (S) *Seven Kings, Essex.*

CAPNISTOS, Emman (S) *Athens.*
CHAPMAN, Derrick James Kenneth (S) *Kingsbridge, Devon.*

DOGRA, Yash Pal (S) *Bangalore.*
DUCKWORTH, John (S) *Bournemouth.*
DUGGAL, Jagmohan Sarup (S) *Bombay.*

EXTON, Harold (S) *Bird, Manitoba.*

FISHER, John Alexander (S) *Nottingham.*

HAIGH, Fred Ellison (S) *Newcastle upon Tyne.*
HENDERSON, Douglas Ross (S) *Codsall, Staffs.*

ISAAC, Ponnazhath Varghese (S) *Shertally, Kerala State.*

KEMP, Paul Courtney (S) *Plymouth.*
KERVELL, Michael George (S) *Cambridge.*

LAMBERT, Keith Walter (S) *Cardiff.*
LEE CHING CHIN (S) *Hong Kong.*

MASSINGHAM, Richard Peter (S) *Bury St. Edmunds.*
MURPHY, Matthew (S) *Tallaght, Co. Dublin.*

PANCHAPAKESAN, Ramachandra (S) *Madras.*
PODLASKI, Jan (S) *Manchester.*
PULLINGER, Gordon Patrick *Birmingham.*

RENGARAJALU, Subbarayalu (S) *Secunderabad.*
RUBEN, Moshe (S) *Kiryat Yam, Israel.*

SHUKLA, Ratan Prakash (S) *Poona.*
SPARKES, Joseph Thomas (S) *Liverpool.*
SRIVASTAVA, Flt. Lt. Uma Shankar *Bangalore.*
SRIVASTAVA, Flg. Off. Vishwanath Sahai *Bangalore.*
STEELE, Michael (S) *Hitchin, Herts.*
STRAND, Roy Charlesworth *Chislehurst.*
SUBRAMANIAM, Vasudeva Ayyar (S) *Trivandrum.*

TEMBE, Shivram Ramchandra (S) *Thana, Bombay State.*
TIMMS, Anthony Francis (S) *Solihull.*

VAN DEN HEUVEL, Teunis *Christchurch, New Zealand.*
VASUDEVAN, V. (S) *Kanpur.*

LIST 2 : The following candidates have now satisfied the requirements of Section A and are eligible to enter for Section B.

BAKKER, Auke (S) *Johannesburg.*
BAMFORD, Thomas Arthur (S) *London, N.22.*
BARTOLO, Anthony Joseph (S) *Hamrun, Malta, G.C.*
BHASIN, Karam Chand (S) *New Delhi.*
BHOWAL, Kamal Kanti (S) *Calcutta.*
BUTT, Saeed Ahmad (S) *Lahore.*

CHAN, Peter Kiu (S) *London, N.7.*
CHAN, Tai-Yuen (S) *London, N.5.*
CHATURVEDI, Ram Nath (S) *Farrukhabad, U.P.*
CHOPRA, Sushil Chander (S) *New Delhi.*

DANDEKAR, Shridhar Kashinath (S) *Bombay.*
DAS, Mukunda Madhale (S) *Calcutta.*
DUTTA, Subal Chandra (S) *Bangalore.*

FAY, John William (S) *London, W.14.*
FRANCIS, Lionel Robert (S) *Cardigan.*
FRASER, William Morrison (S) *Weston-super-Mare.*

GUA ROY, J. K. (S) *Jodhpur.*
GUPTA, Makhan Lal (S) *Allahabad.*

HALPIN, Robert Joseph (S) *London, N.8.*
HANCOCK, Harry James (S) *Great Malvern.*
HO KWOK KI (S) *Hong Kong.*
HOBBS, Francis John (S) *London, W.9.*
HOSANGDI, Rabindranath Radhakrishna (S) *Bombay.*
HOWARD, Trevor Neville (S) *Pietermaritzburg.*
HURST, Leslie (S) *Manchester.*
JOSHI, Ganesh Bahirao (S) *Bombay.*
JOSHI, Rajaram Sivasanker (S) *Bangalore.*
KASTURI RANGAN, Ganapathi (S) *Ernakulam.*

KOHLI, Suraj Parkash (S) *Ambala, India.*

LAWSON, Alan John William (S) *Montreal.*
LEUNG CHEUK SING (S) *Hong Kong.*

MacKINNON, Terance Charles Flynn (S) *Chester.*
MALHOTRA, Prem Kumar (S) *Poona.*
MANTOVANI, Kenneth Paul (S) *London, N.W.8.*
MENON, Venugopalan V. (S) *Poona.*
MEHDI, Capt. Ch. Ghulam. *Quetta.*
MOHAN RANGA RAJ, Goodali Venugopal (S) *Bangalore.*
MORGAN, Bernard John (S) *London, N.1.*

NEWMAN, Martin Myer (S) *London, N.16.*
NIGAM, Surendra Kumar (S) *Bangalore.*

ONDRICH, Milan (S) *Scarborough, Ontario.*
ONWUEGBUZIA, Stephen Onuora (S) *Lagos.*

PILLAI, P. K. Narayana (S) *Secunderabad.*

RAE, Alexander Watson (S) *Bishopbriggs, Lanarkshire.*
RESHEF, Amos (S) *Tel-Aviv.*
REYNOLDS, John Alexander *Cardiff.*
RING, Hans Chanan (S) *Ramat Gan, Israel.*
ROPER, Russell (S) *Middlesbrough.*
RUTTEMAN, Maarten Hendrik (S) *Toronto, Ontario.*

SLANEY, Vernon Howard (S) *Worksop, Notts.*
SUBBARAO, Addepalli (S) *Kurnool, India.*

UNNIKRISHNAN KARTHA, B. (S) *Ernakulam.*

WADHAWAN, Pran Nath (S) *Bangalore.*
WALLACE, David John (S) *Hayes, Middlesex.*
WILLIAMS, David James (S) *Auckland, New Zealand.*

LIST 3 : The following candidate who has already satisfied the Graduateship Examination requirements, has succeeded in passing an additional specialist subject.

BALASUBRAMANYAM, Tiruvarur Rama *Madras* (Television).

(S) denotes a Registered Student.

A Transductor Regulator for Stabilized Power Supplies †

by

A. N. HEIGHTMAN, ASSOCIATE MEMBER‡

A paper read on 3rd July 1959 during the Institution's Convention in Cambridge.

Summary: A new form of transductor regulator is described which is principally intended for use in voltage-stabilized power supplies delivering output currents of ~1 A. Better efficiency and reliability is obtained than is usually possible with the conventional series-valve regulator, and the complete power supply can also be physically smaller. The transductor circuit is unusual in that full-wave operation is obtained with a transductor having only one core and, in the simplest case, only one winding. A definite limit exists, however, to the range of regulation that can be obtained. The relatively slow response of the transductor generally necessitates the incorporation of a valve regulator to deal with rapid disturbances ; such a regulator, working under Class-B conditions for high efficiency, is also described. Details of a complete power supply embodying both regulators are given, and a simpler power supply using only the transductor regulator is briefly discussed.

1. Introduction

The series-valve regulator has for some time been the standard type used in stabilized h.t. power supplies, and has been extensively treated in the literature. Benson[1] provides a large number of references on the subject. Whilst satisfactory for moderately low output currents it is less so where outputs in the region of 1A are required, as for the supply of television studio equipment, for example, with which the author is particularly concerned.

A principal disadvantage of the series valve regulator is the relatively low efficiency, resulting from its high minimum voltage drop, and the large heater power consumption of the high-emission valves that are designed for regulator service. Apart from the waste of input power, the heat generated by the power unit adds considerably to the problem of adequately ventilating the usually confined space of equipment areas.

Although no less reliable inherently than other electronic devices, high-current power supplies using series-valve regulators may require additional maintenance effort. This arises from the

need to use valve types for which replacements are readily available and as there are few, if any, in this category capable of regulating currents in the region of 1A singly, it is necessary to use a number of smaller valves connected in parallel. For a typical 250 V, 1A supply five or six double-triode valves may be used in this way. Consequently frequent checking is essential to ensure that the load is being uniformly shared amongst the valves, as one or two low in emission may cause over-dissipation and premature failure in the remainder. The situation can be improved somewhat, for example by providing individual bias adjustments for each valve, or by including a resistor in series with each valve with the object of reducing the effects of differing anode impedances. These steps involve circuit complication, however, and in the second case, an additional power loss. The number of valves necessary can be reduced by passing part of the load current through a shunt resistor, with some loss of sensitivity, and Neale[2] shows that barretters may be used for this purpose where the load current is substantially constant.

The number of valves to be accommodated, with proper regard to their ventilation, contributes much to the bulk of the conventional power unit, especially where valve rectifiers are used. A physically smaller power unit would undoubtedly be welcomed by those whose business is

the planning of television studio installations.

The belief that it should be possible to devise a new type of regulated power supply, superior to the conventional type in respect of efficiency and reliability, and, of smaller size, led to an investigation of alternative regulating methods. In the course of the work the transductor regulator[3] described here was evolved. The reliability, robustness and low power losses, where suitable core material is used, were features which prompted particular attention to this type of element.

The transductor is included in the power supply rectifier circuit, and is driven by a small control valve and an amplifier similar to that required for a series-valve regulator. It stabilizes the output voltage in respect of slow changes of mains voltage and load current, and rapid variations are dealt with by a supplementary "fast regulator",[3] employing valves operated under Class-B conditions for high efficiency. A description has been given previously[4] of a stabilized power supply unit using both magnetic and valve regulators, designed for high frequency a.c. input supplies.

The voltage-amplifier stages, common to both regulators, are not dealt with at length here, although special features are noted where necessary.

A description of the transductor circuit and its mode of operation are given, with an analysis the results of which may be used for design purposes. The fast regulator is described, and finally details of a complete stabilized power supply incorporating both regulators are presented.

2. Some Properties of Transductors with "Rectangular-Loop" Core Material

The term "transductor" is used to describe the basic electromagnetic component of the magnetic amplifier, consisting of one or more cores having one or more windings, some encompassing several cores. Here the transductor has only one core, and in the simplest case, only one winding, and the term is used in preference to "saturable reactor" to distinguish its properties from those most generally associated with the latter.

The core material used for the regulator transductor is of the type developed for magnetic amplifier use: this has a B-H characteristic that is very nearly rectangular, a high permeability and a high saturation flux density.[5] The properties of transductors using such core material, as circuit elements, differ from those resulting from the use of ordinary core materials of moderate permeability, and in which saturation is approached only gradually with increasing magnetizing force. In the latter case the component may be used as an inductive impedance whose effective value may be smoothly changed by variation of the steady magnetization of the core. A familiar form of this type is the "swinging-choke", used in rectifier circuits to obtain a more nearly constant output voltage with varying load current; it also finds application in regulators for a.c. supplies. Benson[1] describes some of these applications, and provides further references to the literature.
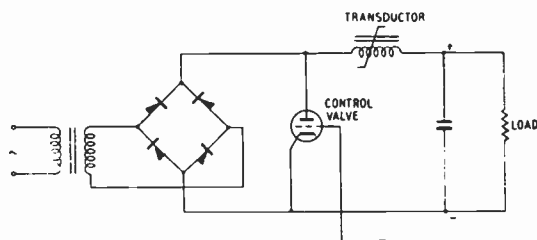

Fig. 1. Basic transductor regulator circuit.

When the core material is of rectangular loop material, however, the transductor behaves rather as a switch, because the onset of saturation is abrupt, and very high ratios of unsaturated to saturated impedance can be obtained. Another property, implied by the description of the B-H loop shape, is a very high remanent flux density, which for many purposes may be assumed to be equal to the saturation flux density, so that the core, once saturated, will remain substantially in that condition indefinitely, until a reverse magnetization is applied.

2.1. Core Structure

The rectangular loop alloys are characterized in that the special properties are only obtained for magnetization in certain preferred directions, as a result of the domain or crystal orientation introduced in manufacture. The properties of the materials are therefore more fully realized in wound-tape cores, a preferred direction being along the tape. The materials are available in

laminated form, special lamination shapes being used in which a comparatively slight deterioration of properties occurs.

The use of toroidal cores is advocated for the regulator transductor described here, although they are not essential, and useful results were obtained with laminated cores in early experiments.

3. The Transductor Regulator

Figure 1 is a basic circuit diagram of the regulator. As can be seen, the transductor is included in the power supply rectifier circuit in the path between the full-wave bridge rectifier and the reservoir capacitor. The load is shown simply as a resistor shunting the reservoir. The purpose of the valve is to control the magnetization of the transductor, and is referred to as the "control valve". Its grid is fed from earlier stages similar to those required for a series-valve regulator.

3.1. *Description of a Cycle of Operation*

A cycle of operation is described with the aid of the voltage, current, and flux-waveforms of Fig. 2, and the *B-H* diagram of Fig. 3. For clarity, some details of the waveforms observed in practice have been omitted. The numbered
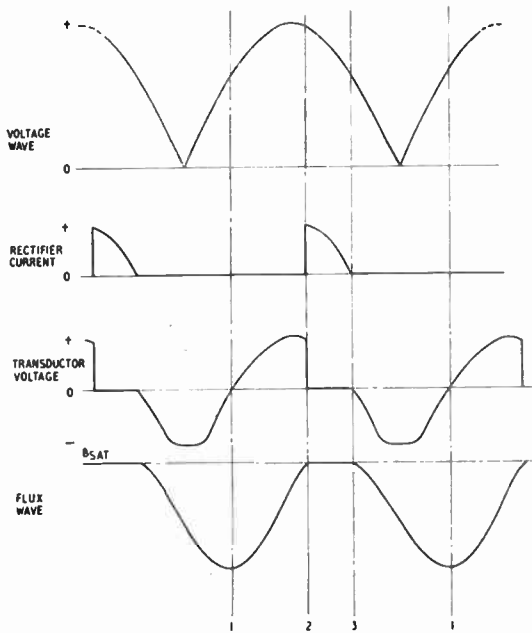

Fig. 2. Voltage, current, and flux waveforms.

points in the two figures represent the same instants in time. As full-wave rectification is employed, the cycle of operation repeats every half-cycle of the a.c. input wave.
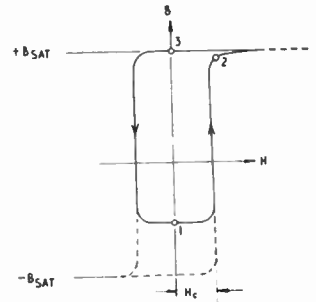

Fig. 3.
B-H diagram.

At the start of a half-cycle, under steady-state conditions, the a.c. input wave passes through zero and increases positively. At the instant indicated at (1) the input reaches a voltage equal to that at the reservoir capacitor, and the appropriate pair of rectifier arms of the bridge begins to conduct, applying the input voltage to the transductor and reservoir. At this time the transductor is not saturated, as shown by the *B-H* diagram, so that only the small magnetizing current of the transductor flows into the reservoir, and negligible charge is acquired. In order to support the impressed voltage the transductor core flux must change, until at (2) saturation is reached. The interval between the instants (1) and (2) is called the "preconduction period".

With the core saturated, and the transductor impedance very small, full rectifier conduction occurs, the core being driven further into saturation. The reservoir charge increases until at (3) the reservoir voltage is equal to the instantaneous input voltage, and rectifier conduction then ceases as the input voltage, proceeding in its cycle, falls below that at the reservoir. As the magnetizing force falls to zero the core flux returns to the remanent point (3).

With the rectifier branch non-conducting the path containing the transductor, control valve and reservoir becomes effective. It will be seen that the reservoir voltage is applied to the transductor via the valve, and that the polarity of this voltage at the transductor terminals is opposite to that applied from the rectifier bridge when conducting. The valve grid has some steady potential enabling it to conduct. Thus from the instant (3) the core flux begins to fall from the

remanent value. The fall continues until rectifier conduction recommences at (1) in the succeeding half-cycle; the valve current then being drawn from the rectifier. The interval between the instants (3) and (1) is called the "control period".

It is convenient to assume that the flux change occurring during the control period depends simply on the valve current; this point is further discussed in Section 3.7. Thus if the valve current is zero the core remains at the positive remanent flux level; if the valve current is at a maximum the core will reach negative saturation.

The charge acquired by the reservoir in each half-cycle, and hence the mean output voltage, depends on the time for which full-conduction of the rectifiers can occur. If the transductor core has remained at saturation during the control period, then when the instantaneous a.c. input voltage reaches equality with the reservoir voltage, full-conduction occurs at once. Under this condition the rectifier circuit behaves substantially as it would without the transductor, and a maximum relative output voltage is obtained. An additional voltage drop is caused by the winding resistance of the transductor, but in a suitable design this resistance will be only about 1 per cent of the load resistance, so that the drop is quite small. With suitable core structure the saturated reactance will have a negligible effect on the mean output voltage.

If, on the other hand, the core is at negative saturation at the instant that the rectifier is able to conduct, the transductor will prevent full-conduction whilst the core moves towards positive saturation under the influence of the applied e.m.f. The charge then received by the reservoir in each half-cycle will have its minimum value, resulting in a minimum relative mean output voltage.

The d.c. output for a given a.c. input is thus a function of the control valve current and hence of its grid potential. The output voltage can therefore be held constant whilst the a.c. input and load current vary, by suitable variations in grid potential.

The regulator may be regarded as a special case of the *half-wave* self-saturated magnetic amplifier.[6]

3.2. *Transductor Volt-Second Storage*

It has been seen that conduction of the recti-

fiers is prevented whilst the core flux moves towards saturation. The change of flux is necessitated by the application of the input e.m.f. via the rectifiers, for to satisfy Kirchhoff's second law the transductor terminal voltage must equal the applied e.m.f. From Faraday's law:

$$e = N \frac{d\varphi}{dt} . 10^{-8} \qquad \dots\dots\dots(1)$$

where e is the e.m.f. in volts, N the number of turns cut by the flux φ, in maxwells†. Here, however, we are interested in the *change* of flux, for it is this, rather than the rate of change, which is influenced by the control valve. The change of flux can be shown to be proportional to the integral of the applied voltage, for from eqn. (1),

$$\frac{e . 10^8}{N} dt = d\varphi,$$

and if the flux change is $\Delta\varphi$, from φ_1 to φ_2 say, in the interval t_1 to t_2

$$\Delta\varphi = \int_{\varphi_1}^{\varphi_2} d\varphi = \frac{10^8}{N} \int_{t_1}^{t_2} e.dt \qquad \dots\dots\dots(2)$$

This is more useful practically in terms of the flux density B, and the core area A, and since $\Delta\varphi = \Delta BA$, we have, rearranging,

$$\int_{t_1}^{t_2} e.dt = \Delta BAN . 10^{-8} \qquad \dots\dots\dots(3)$$

in which B is in gauss and A in cm.

If now $-\Delta B$ is the change in flux density which has occurred in the control period, a flux density of $B_{SAT} - \Delta B$ exists in the core at the end of the period, where B_{SAT} is the saturation density. Thus the same flux excursion, ΔB, must take place during the following preconduction period before saturation is reached and rectifier current can flow. It can therefore be stated that the flux excursions in the control and preconduction periods are equal, and from eqn. (3), that the integral of the transductor terminal voltages is also the same for the two periods. Of course, neither the form of the voltages nor their duration of application need be the same.

† A list of symbols is given in the Appendix.

The transductor has the greatest effect upon the circuit when the maximum flux change, from $+B_{SAT}$ to $-B_{SAT}$, occurs. Then $\Delta B = 2B_{SAT}$. Introducing the symbol S to represent a voltage-time integral, eqn. (3) becomes

$$S = \Delta BAN.10^{-8} \qquad \ldots\ldots(4)$$

and the maximum value of the integral for a given transductor,

$$S_t = 2B_{SAT} \; AN.10^{-8} \text{ volt-seconds.} \quad \ldots\ldots(5)$$

This quantity is referred to as the "storage" of the transductor.

3.3. Effect of Circuit Parameters on Input/Output Relationships

Whilst the circuit action is in fact extremely complicated, involving non-linear elements and complex functions of current and voltage, an analysis has been performed with the aid of a number of simplifying assumptions, the results of which are useful for design purposes. The steady-state relationships between the a.c. input voltage, d.c. output voltage, and load current, and how they depend on the rectifier circuit resistance and transductor storage, are developed in what follows. The further relationship between the effective storage and the control valve grid potential, or anode current, has not yet been satisfactorily established for the present circuit. The problems involved have been discussed, in relation to the half-wave magnetic amplifier, by Dornhoefer[6] and Lehmann[7]. Fortunately, in a degenerative regulating circuit of high loop gain, the grid potential automatically takes the value appropriate to the external conditions.

The transductor B-H characteristic is first of all assumed to be perfectly rectangular, and the magnetizing current to be zero. These assumptions are not unreasonable, particularly where toroidal wound tape cores are used; in a suitable design the magnetizing current will be about 1 per cent of the maximum mean load current, and will contribute a negligible charge to the reservoir. The transductor may therefore be regarded as a switch, closing the rectifier circuit on saturation.

The reservoir voltage is assumed to be constant; this proves to be a tolerable simplification,

for the effect of the ripple voltage which exists in practice is to increase the accumulation of charge early in the conduction period and to reduce it later, thus the average charging rate is not greatly different from that implied in the assumption. The actual reservoir capacitor-load resistance time-constant should not be less than about $5/f$, where f is the supply frequency, although the exact value is not at all critical.

The rectifier forward resistance is also taken to be constant, i.e. not a function of current. This would represent the practical case fairly closely where semi-conductor rectifiers of very low forward resistance are used, with current-limiting resistors in series. The case of non-linear forward resistance is discussed in Section 3.6.

The circuit series inductance is taken to be zero, although in practice it exists in the leakage inductance of the mains transformer and the saturated inductance of the transductor. As Moullin[8] has noted, in connection with the ordinary rectifier circuit, neglect of the inductance does not lead to large errors in calculations relating to the mean output current, for whilst its effect will be to retard the initial rise of rectifier current, it will also tend to prolong the flow as the rectifier ceases to conduct.
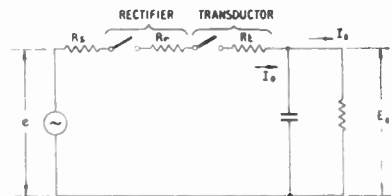

Fig. 4. Equivalent rectifier circuit.

The circuit is reduced to that of Fig. 4 in which R_s is the a.c. supply source resistance, R_r the total resistance of the two rectifier arms active in one half-cycle of the supply, and R_t is the transductor winding resistance. The rectifiers and transductor are otherwise represented by switches, as shown. Figure 5 shows a half-cycle of the a.c. input wave, taken for convenience as the cosine wave, $e = E_p \cos \theta$. The switch representing the rectifier is closed when e exceeds E_0, the reservoir voltage, at the angle $-\alpha$, so that the rectifier circuit e.m.f. is

$$e_I = E_p \cos \theta - E_0; \; (E_p \cos \theta > E_0), \quad \ldots\ldots(6)$$

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and the rectifier switch opens again at the angle α, and we have

$$\cos(-\alpha) = \cos\alpha = \frac{E_0}{E_p} \qquad \ldots\ldots\ldots(7)$$

Now the average rectifier circuit current must equal the mean output current since the *average* reservoir current must be zero, (the charge lost must equal the charge gained in each complete half-cycle, in the steady-state). The output current, I_0, is therefore given by[4]

$$I_0 = \frac{E_a}{R} \qquad \ldots\ldots\ldots(8)$$

where E_a is the average rectifier circuit voltage, and R the sum of R_t, R_r and R_t. The average value of e_t is given by

$$\frac{1}{\pi}\int_{-\alpha}^{\alpha} e_t \, d\theta = \frac{1}{\pi}\int_{-\alpha}^{\alpha} (E_p\cos\theta - E_0)d\theta$$

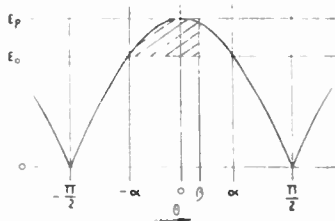$$= \frac{2}{\pi}(E_p\sin\alpha - \alpha E_0) \qquad \ldots\ldots\ldots(9)$$

Fig. 5. Rectifier circuit voltage wave.

From this must be subtracted the average voltage lost whilst the transductor proceeds towards saturation, and saturation will occur when the integral of e_t reaches the amount equivalent to the flux change of the control period. The integral is represented by the shaded area in Fig. 5. If this is S volt-secs, and the angular frequency $\omega = 2\pi f$ radians/sec, the integral is ωS volt-radians, and the corresponding average value of e_t is

$$\frac{1}{\pi}\int_{-\alpha}^{\beta} e_t \cdot dt = \frac{\omega S}{\pi} \qquad \ldots\ldots\ldots(10)$$

subtracting this quantity from eqn. (9), E_a is obtained directly:

$$E_a = \frac{1}{\pi}\left[2(E_p\sin\alpha - \alpha E_0) - \omega S\right] \ldots\ldots\ldots(11)$$

Substituting for E_p from eqn. (7) and for E_a from eqn. (8) we can obtain:

$$I_0 = \frac{1}{\pi R}\left[2E_0(\tan\alpha - \alpha) - \omega S\right]\ldots\ldots\ldots(12)$$

Finally, in the more useful form:

$$\frac{I_0 R + 2fS}{E_0} = \frac{2}{\pi}(\tan\alpha - \alpha) \qquad \ldots\ldots\ldots(13)$$

This is related to the peak input voltage E_p by the angle α, $\cos\alpha = E_0/E_p$, or in terms of the equivalent r.m.s. value, E_{ac}, $\cos\alpha = \sqrt{2}.E_0/E_{ac}$. We can therefore put

$$\frac{E_{ac}}{E_0} = \frac{1}{\sqrt{2}\cos\alpha} \qquad \ldots\ldots\ldots(14)$$

The desired relationship is that of $(I_0 R + 2fS)/E_0$ and E_{ac}/E_0, which cannot be solved directly. However, a graphical solution can be obtained by plotting $1/(\sqrt{2}\cos\alpha)$ against $2(\tan\alpha - \alpha)/\pi$ over a range of appropriate values of α, as has been done to obtain the curve of Fig. 6.

Equations (13) and (14), with the aid of Fig. 6, may be used for calculation of the quantities of interest.

3.4. Limitation of Regulator Range

A fundamental limitation exists in the present circuit of the value of the control period integral, and hence of the preconduction period integral, S, in eqn. (13).

During the control period the left-hand terminal of the transductor (Fig. 1) is negative with respect to the other, and if the transductor voltage exceeds the difference between the reservoir voltage and the instantaneous a.c. input voltage applied via the rectifiers, the rectifiers in two arms of the bridge will conduct. This therefore represents a limit to the control period voltage swing, and also to the integral, whose maximum value is shown as the shaded area in Fig. 7. In practice not all the area shown is available, for the voltage excursion cannot reach zero as the control valve will require a certain minimum anode voltage, shown by the dotted line, and the
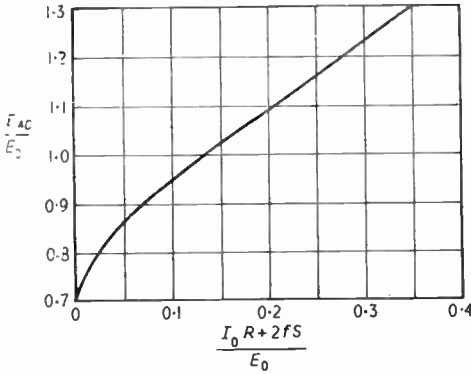
Fig. 6. E_{ac}/E_0 as a function of $(I_0R+2fS)/E_0$

area lost will be that below this line. The loss may be negligible where a pentode is used, but in the case of a triode it can result in a significant restriction of control range. Assuming that the loss is negligible, the integral is given by:

$$\left(\int e_c.d\theta\right)_{max}=2\left[E_0(\pi/2-\alpha)-E_p\int^{\pi/2}\cos\theta.d\theta\right]...(15)$$

which results in

$$\left(\int e_c.d\theta\right)_{max}=2E_0(\tan\alpha-\alpha+\pi/2-\frac{E_p}{E_0})\quad......(16)$$

If the maximum storage of the transductor, S_t, exceeds the integral defined by eqn. (16) then the transductor flux density cannot reach $-B_{SAT}$ in the control period. The effective transductor storage will therefore be less than S_t, and in fact will be that defined by eqn. (16). To determine what limitation is thus imposed on the variables of eqns. (13) and (14), eqn. (16) can be put in terms of the effective storage, S_e:

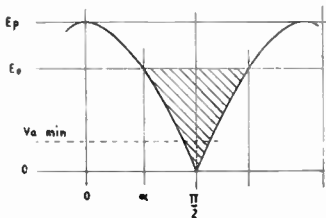$$\omega S_e=2E_0(\tan\alpha-\alpha+\pi/2-\frac{E_p}{E_0})$$

Fig. 7. Limiting control period integral.

or

$$S_e=\frac{E_0}{\pi f}(\tan\alpha-\alpha+\pi/2-\frac{E_p}{E_0})\quad......(17)$$

Now rearranging eqn. (13)

$$S=\frac{E_0}{\pi f}(\tan\alpha-\alpha)-\frac{I_0R}{2f},$$

and putting $S_e \geqslant S$ we find the limitation is

$$\frac{E_p}{E_0}\leqslant\pi/2\left(\frac{I_0R}{E_0}+1\right)\quad.........(18)$$

A maximum storage is required in each half-cycle when E_0 is to be held constant despite a maximum value of E_{ac} and a minimum value of I_0. Putting $I_0R=0$ to represent either the ideal case of zero R, or the practical case of zero load current, the maximum value of E_p/E_0 for which regulation can be obtained is, from eqn. (18):

$$E_p/E_0=\pi/2,\text{ or }\frac{E_{ac}}{E_0}=\pi/2\sqrt{2}\quad.........(19)$$

Under the same conditions, $I_0R=0$, but with the transductor saturated throughout each half-cycle, $(S=0)$, the minimum value of E_p/E_0 obtains, which is $E_p/E_0=1$, for clearly the output voltage will reach the peak value of the input wave. The maximum possible range of relative input voltage for which regulation is obtained is thus between the limits of $E_p/E_0=\pi/2$ and $E_p/E_0=1$. Expressed as a percentage variation about the arithmetic mean of $\pi/2$ and 1 this is ±28 per cent. In practice a range of about ±15 per cent can be achieved with constant load current; with adverse load current variation the range is rather less. When I_0R increases from zero both the maximum and minimum values of E_p increase (for constant E_0), but the minimum value does so more rapidly so that the range becomes less than the ideal.

From Fig. 7, or less readily from eqn. (16) it can be seen that as E_p increases with E_0 constant the available control period integral, S_e, falls, and from eqn. (18) it can be seen that as I_0R increases so must E_p if S_e is to be fully utilised. The maximum value of S_e therefore occurs when $I_0R=0$, and we then have, from eqn. (18), that $E_p/E_0=2/\pi=\cos\alpha$. Substituting for this value of α in eqn. (17) we find the maximum value

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of S which can be used with advantage, which is:

$$S_{max} = 0 \cdot 106 \ E_0/f \qquad \ldots\ldots\ldots(20)$$

If the transductor is designed to have a storage S_t equal to S_{max} then the maximum range of regulation can be obtained. In particular E_0 can be held constant at low load currents.

In using the foregoing equations for design purposes the following points may be noted. The given information as to external conditions will generally be the maximum and minimum load currents and the percentage variation of the a.c. supply. The minimum transductor storage will be required under the condition of minimum a.c. input and maximum load current, so that by putting $S = 0$ and the maximum value of I_0 in eqn. (13) the minimum transformer secondary voltage, E_{ac1}, can be obtained. Then from the percentage a.c. input variation the maximum value, E_{ac2} follows, which inserted in eqn.(14) enables a value of $2(\tan \alpha - \alpha)/\pi$ to be obtained. From this, and the minimum value of I_0, the maximum value of S can be calculated. The restriction defined by eqn. (18) should be noted.

If inferior core material is used, or if the rectifier reverse current (Section 3.8) is high or uncompensated, the condition of $S = 0$ may not be achieved, and it will be necessary to assume some fraction of the maximum value of S, as experience may dictate.

3.5. R.M.S. and Peak Rectifier Circuit Currents

The r.m.s. value of current needs to be known in connection with the mains transformer and transductor design. In the present case the small conduction angle forced upon the rectifier by the transductor action leads to r.m.s. currents higher than normal. Calculations of this current proves to be difficult, for its waveform is considerably modified by the presence of circuit inductance, and by the exact rectifier characteristics, which it has hitherto been permissible to to neglect for the reasons given in Section 3.3. The r.m.s. current value can be obtained with reasonable accuracy for the condition of $S = 0$, for then the conduction angle is greatest, and the effect of inductance least. The instantaneous current is

$$i = (E_p \cos \theta - E_0)/R \qquad \ldots\ldots\ldots(21)$$

and the r.m.s. current therefore:

$$I_{rms} = \frac{1}{R}\left[\frac{1}{\pi}\int_{\theta_1}^{\theta_2}(E_p \cos \theta - E_0)^2 d\theta\right]^{\frac{1}{2}} \qquad \ldots\ldots\ldots(22)$$

When $S = 0$, $-\theta_1 = \theta_2 = \alpha$ and eqn. (22) reduces to

$$I_{rms1} = \frac{E_p}{R}\left[\frac{1}{\pi}(\alpha - 3 \sin \alpha \cos \alpha + 2 \alpha \cos^2 \alpha)\right]^{\frac{1}{2}} \tag{23}$$

Moullin[8] gives a convenient approximation; adapted for conformity it is

$$I_{rms1} = 1 \cdot 16 \ I_0(1 - \frac{E_0}{E_p})^{-1/4} \qquad \ldots\ldots\ldots(24)$$

A reasonable approximation when S is not zero has been found to be:

$$I_{rms2} = I_{rms1} \ E_{p2}/E_{p1} \qquad \ldots\ldots\ldots(25)$$

in which I_{rms1} and E_{p1} are the values corresponding to the condition of $S = 0$, and I_{rms2} and E_{p2} to $S \neq 0$.

The peak current is similarly affected by circuit inductance and rectifier characteristic. When $S = 0$ the peak current is in practice close to the theoretical value,

$$I_{p1} = (E_{p1} - E_0)/R \qquad \ldots\ldots\ldots(26)$$

No satisfactory method of deriving the peak current has yet been found for $S \neq 0$. It rises rather more rapidly with S than does I_{rms}, but is unlikely to reach the value $i_{p2} = (E_{p2} - E_0)/R$, for under conditions of the maximum value of E_p the transductor prevents conduction until the peak of the wave is passed.

3.6. Approximations for Non-linear Rectifier Characteristic

So far it has been assumed that the rectifier forward resistance is constant. Two methods have been found practically useful for dealing with the case of a non-linear characteristic.

Firstly, the rectifiers may be operated in the circuit prior to the addition of a transductor, and from the measured value of E_{ac} the value of R determined from eqns. (13) and (14). From this total rectifier circuit resistance R_r can be deduced. R_r, so found, will be an apparent average value, and will vary with the load current. The appropriate load currents should

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therefore be drawn, and in subsequent calculations the corresponding value of R_r used.

An alternative method, where a graph of the rectifier characteristic is available, is to make a straight line approximation to the curve. The slope of the line $\delta e/\delta i$ is taken to be R_r. The line will meet the axis, not at the origin, but at a forward voltage V_r, say, so that the fictitious rectifier will not pass current until the applied e.m.f. exceeds V_r. With this assumed characteristic, therefore, rectifier conduction will not commence until $e = E_p \cos 0 = E_0 + V_r$. E_0 is therefore replaced in eqns. (6) to (20) by $(E_0 + V_r)$ where this approximation is made.

Both methods have given results sufficiently accurate for normal design purposes, although the second method has had only limited experimental confirmation.

3.7. *Control Valve Current*

A property of the rectangular-loop core material is that only a very small flux change occurs until the magnetizing force H reaches the value H_c, half the width of the static B-H loop; then the rate-of-change of flux is only limited by the external e.m.f. available to maintain the magnetizing current. This serves to demonstrate the significance of the voltage-time integral in dealing with this type of core material. If the e.m.f. is high, so that the flux change can be rapid, the width of the loop is increased[7] as a result of the increased eddy current losses.

The valve current during the control period is given approximately by

$$I_c = \frac{10.Hl}{4\pi N} \qquad \dots\dots(27)$$

where H is in oersteds, l is the effective core length in cm, and N the number of turns. To obtain the value of the minimum current H is put equal to H_c for the static loop. The maximum value is obtained by taking the value of H corresponding to the onset of saturation in the dynamic loop for the working frequency, twice the supply frequency in the present case.

The valve is allowed to conduct throughout the cycle, although its current is only useful in the control period. During this period the valve anode potential is lower than in the remainder of the cycle, and the grid potential is set so that an appropriate current is drawn. Through the re-

mainder of the cycle the anode voltage is higher and since the grid potential is unchanged the anode current is also higher, so that the average current will be greater than given by eqn. (27) for the control period. Where a pentode is used the difference may be slight, but in the case of a comparable triode the average current will be approximately twice as large. For further consideration of anode circuit conditions it may be noted that Fig. 7 represents the anode voltage waveform in the limiting condition.

It would be possible to reduce the current to about one half in the case of a pentode, or one quarter in the case of a triode, by arranging for the valve to be cut off excepting during the control period, though this may not be justifiable in view of the circuit complication involved, as a maximum current of about 10mA only is required to control a transductor for a 1A 250V power supply.

3.8. *Effect of Rectifier Reverse Current*

It can be seen from Fig. 1 that besides the path for transductor reverse current provided by the control valve, an additional path exists in the reverse impedance of the rectifiers, and current flowing by this path will produce an undesired flux change in the transductor. The practical effect of this is that the minimum a.c. input voltage for which regulation is obtained is higher than expected.
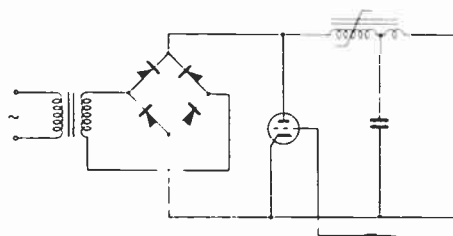

Fig. 8. Transductor regulator with rectifier reverse current compensation.

Rectifiers having a low reverse current should be used, such as are available for magnetic amplifier service, and a further reduction in reverse current can be secured by operating the rectifiers at less than their rated maximum peak inverse voltage. If, having taken these steps, a loss of regulation range remains, the residual reverse current can be neutralized by applying a forward current bias to the transductor. This can be

readily done by passing the load current through a few additional turns wound on the transductor core, so that the circuit is as shown in Fig. 8. The bias should be determined in relation to the maximum load current; the fact that the bias will be reduced at lower currents will generally be of no consequence, as it is at maximum load current that the effect of the rectifier reverse current is most pronounced.

Positive feedback is introduced by the arrangement of Fig. 8, and will actually aid the regulating action; if overdone, however, instability will result. The four rectifier arms should preferably be similar in respect of reverse current, as differing reverse currents in alternate half-cycles may induce a tendency towards half-wave rectification.

A simple method of measuring reverse current in a bridge rectifier is given by Walker and Martin.[10] Whilst the relationship is not clear, it appears that the effective reverse current in the present circuit is about half that measured by this method.

The reverse impedance in the case of selenium rectifiers includes a capacitive component, and is a function of the immediately previous forward current flow.[11]

3.9. Effect of Ripple Component at the Control Valve Grid

It is important to keep to a minimum any ripple voltage which might reach the control valve grid, as its effect is to cause non-uniform rectification of alternate half-cycles, and if the ripple is large, half-wave rectification may result. The mechanism is as follows.

If, during a control period, the grid potential is raised as a result of the presence of the ripple component, then a correspondingly increased flux change will occur in the transductor. In the following conduction period, therefore, a smaller charge will be received by the reservoir, and hence in the following control period the ripple component will be less than in the previous control period. In consequence a reduced flux change will occur, so that in the next conduction period a greater charge will be received by the reservoir. The pattern repeats in subsequent cycles until the development of the condition is limited by non-linearity in the control valve

or the preceding amplifier. In the extreme case conduction is completely suppressed in alternate half-cycles. The process can be started by some initial asymmetry in the rectifier bridge either in the forward or reverse characteristics.

Related effects can be caused by a.c. components other than ripple, for example those arising from cyclic variations of current in the load apparatus, at or near the supply frequency or its second harmonic. In general it appears that such load modulation causes little trouble unless very heavy. An effective cure is to include a resonant filter in the control valve grid circuit tuned to the appropriate frequency. A single filter can serve for both supply frequency and second harmonic if tuned intermediately.

3.10. Voltage-Doubler Circuit

The regulator may be adapted for use with a voltage-doubler rectifier system, as shown in Fig. 9. As this system consists, in effect, of two half-wave rectifiers with their outputs connected in series, to obtain the full-wave transductor operating conditions of the bridge circuit it is
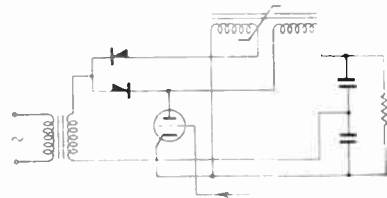

Fig. 9. Transductor regulator with voltage-doubler rectifier.

necessary to include a winding of the transductor in each half-wave path. Each winding will require approximately half the number of turns used in the bridge circuit for the same output voltage. The windings must be designed for the higher r.m.s. current of the half-wave circuit.

3.11. Alternative Control Method

A control circuit arrangement which may be useful practically is shown in Fig. 10. Here a resistor Rb draws sufficient current from the reservoir to carry the core substantially to negative saturation when the valve current is zero. The control winding is connected so that

the valve current will magnetize the core in the direction of positive saturation, opposite to its effect in the Fig. 1 circuit, so that increasing valve current will *reduce* the flux change of the control period.

There are a number of features of this arrangement that may be advantageous. The sense of control is reversed, which may be convenient in some cases. The valve anode voltage swing is also inverted, so that it is higher during the control period than in the rest of the cycle and the mean current where a triode is used will be lower than in the Fig. 1 circuit. Further, there
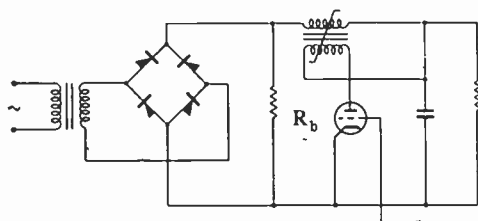
Fig. 10. Transductor regulator with alternative control circuit.

is less likelihood of the regulation range being restricted by insufficient control period anode potential. It may be also noted that in this arrangement the flow of current in Rb and in the rectifier reverse impedance is in the same direction, so that compensation for the latter can be effected by suitably proportioning Rb.

3.12. *Transductor Power Dissipation*

The losses within the transductor consist of the copper loss, $I_{rms}^2 R_t$, and the core loss. These are least when $S=0$, for then the r.m.s. current has its lower value and negligible flux changes occur in the core, and greatest when S has its maximum value. The copper loss presents no special problems, and the design of windings can follow usual practice in this respect. The core loss, however, is found to be three to four times that calculated from the known area of the *B-H* loop at the operating frequency. It is possible that the additional loss is caused by eddy currents in the saturation region, and further investigation is desirable.

In a transductor such as that used to obtain the results of Section 3.14, designed for a 1A,

250V supply, the maximum copper loss is about 10 watts and the core loss 2 to 3 watts.

3.13. *Response Time*

The time taken for the mean reservoir voltage to reach its new value following a step-change of control valve grid potential includes the delay, both in action of the transductor, and in changing the charge on the reservoir capacitor.

Since a constant datum exists for the control period flux excursion in every half-cycle (the remanent flux density; point (3) in Fig. 3), the inductive delay associated with some types of magnetic amplifier is absent in this case, as it is in related single-core self-saturated magnetic amplifiers. A change of control period flux excursion produces the corresponding change of conduction angle in the conduction period immediately following. The minimum transductor response time occurs when the step-change coincides with the start of a control period; the maximum when it coincides with the end of a control period, for then it cannot be effective until after the control period following. The minimum occupies about one half-cycle of the supply frequency and the maximum just under one cycle.

It would be expected that the time taken for the reservoir to reach a new value of mean voltage could be found fairly directly from the appropriate time constants: RC_R, the rectifier circuit resistance—reservoir capacitance time constant for increments, and R_lC_R, the load resistance–reservoir capacitance time constant for decrements. In practice, however, the observed response times have been less than might be determined in this way. With the supply frequency of 50 c/s, and measuring the time up to 90 per cent completion of the change, an increment occupies about one cycle (20 msec), and a decrement about $1\frac{1}{2}$ cycles (30msec), where $RC_R \simeq 7\cdot5$msec and $R_lC_R \simeq 125$msec. Whilst RC_R is less than 20msec, the duration of conduction is only about 3msec in each half-cycle. The explanation appears to lie in the variation in the starting time of full-conduction that results from the transductor action. In the case of a rise in output, for example, the reservoir charge builds up from an initially higher potential than in previous half-cycles because at the earlier instant of conduction less charge has been lost in the supply of load current.

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3.14. *Measured Characteristics*

The curves of Fig. 11 have been plotted from measurements taken on the basic circuit of Fig. 8. The grid potential of the valve was adjusted so as to hold the mean reservoir voltage at 250, whilst the a.c. input was varied, and for the two values of load current shown. The results are typical and give an indication of the capabilities of the regulator. It will be noted that the transformer secondary voltage is considerably less than would be necessary for a series-valve regulator.

The effect of the compensating winding, discussed in Section 3.8, has been to displace the curves to the right: the 1A curve by 2·2mA and the 75mA curve by 0·2mA approximately. The bias is hardly necessary for the particular rectifiers used in this case, but is provided to deal with rectifiers of greater reverse current which may be encountered.

Some details of the components used to obtain the curves are as follows:—

Transductor, storage, S_t: 0·54 volt-seconds
resistance R_t: 2 ohms
core: HCR alloy toroidal wound-tape, tape thickness 0·004 in.

Control valve: EF91 pentode.

Rectifier: high-grade selenium, 12 plates 4·5 cm square in each bridge arm.
R_r, (24 plates): 5·1 ohms
V_r, (24 plates): 18V.

Total value of R: 17 ohms.

Reservoir capacitance: 500μF.

Supply frequency: 50 c/s.

The maximum and minimum r.m.s. a.c. input voltages were calculated by means of eqns. (13) and (14) and are compared below with the measured values. The corresponding values of I_{rms} were calculated from eqn. (26).

I_0		1A	0·5A	0·25A	75mA
E_{ac2}	calculated	312	304	299	295
	measured	231	215	208	201
E_{ac1} (volts)	calculated	228	212	205	198
	measured	313	304	299	294
I_{rms1} (amp)	calculated	1·8	—	—	—
	measured	1·7	—	—	—
I_{rms2}	calculated	2·48	—	—	—
	measured	2·3	—	—	—
I_{p1} (amps)	calculated	4·1	—	—	—
	measured	3·85	—	—	—
I_{p2}	measured	6·5, 7·1	—	—	—

A slight discrepancy between the measured values of E_{ac} here and in Fig. 11 is due to the measurements being taken in the first case off-load, and in the second, on-load. The two values of I_{p2} refer to alternate half-waves; some asymmetry of the rectifier bridge is evident at the higher peak current.
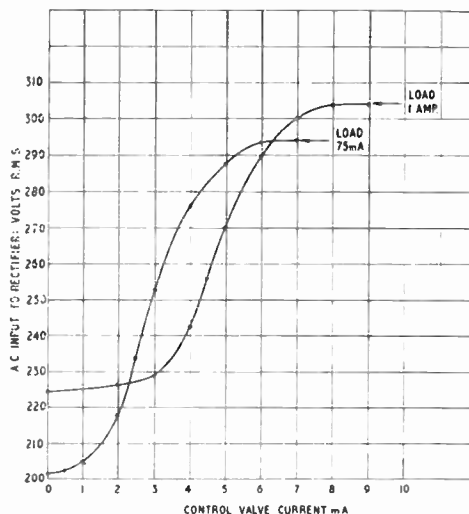

Fig. 11. Measured characteristics.

The limiting control period integral is just reached at $I_0 = 1$A and $E_{ae} = 312$ volts, the transductor being designed for the maximum useful storage under this condition.

The rectifier current, and transductor voltage waveforms are shown in Fig. 12(a) and (b) as photographed from a double-beam oscilloscope. The conditions are as for the measurements noted above. The two pairs of traces were taken at 1A output current, and at 231 and 313 a.c. volts input. (The kink in the transductor voltage trace for $E_{ac2} = 313$V was caused by an impulsive load on the local mains supply, and does not originate in the transductor regulator.) The scales are the same for the two sets of waveforms. The peak current values here are given above; the peak-to-peak voltage in Fig. 12(b) is 410V.

3.15 *Simple Transductor-Regulated H.T. Supply*

In a complete power supply to replace directly the conventional type the relatively slow response of the transductor necessitates the addition of a fast-acting valve regulator. Where a more

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limited performance is acceptable the valve regulator need not be included. An important function of the fast regulator, however, is to reduce the ripple voltage at the output terminals to a very small amount, so that in its absence an efficient smoothing filter is essential. As the filter will introduce a considerable delay into the regulating loop, suitable phase compensation must be applied to prevent oscillation and to obtain a satisfactory transient response.

The difficulty of achieving proper compensation proves to limit the usefulness of the simple regulator, and in an experimental unit using an L-C filter an acceptable transient response could not be obtained. However, by using a less efficient two-stage R-C filter, and including only the first stage in the regulating loop the resulting phase characteristic is more amenable to compensation, and this proves to be a practical arrangement.

As the series resistor of the second filter stage is outside the regulating loop the output impedance of the supply becomes substantially equal to the value of this resistor. This will be

higher than would usually be regarded as permissible, and relatively large variations of output voltage will occur with changes of load current. These variations can, however, be prevented from affecting critical sections of the load apparatus by feeding these through separate second filter stages from the regulated point.

The circuit diagram of a simple regulated supply designed on these lines is shown in Fig. 13. One half of the double-triode is used as an amplifier and the other as the control valve. The stabilized negative supply, providing bias for the valve grids, also serves as the reference voltage source, and is used externally for protective bias of certain valves in the load apparatus. R12 and C6 are the phase compensating components.

As the single amplifier stage drives the control valve grid positively for falling voltage at the regulated point, the alternative control scheme, described in Section 3.11, is used in order to operate the transductor in the correct sense. A voltage-doubler rectifier was required for the particular application, although the bridge type would generally be equally applicable.
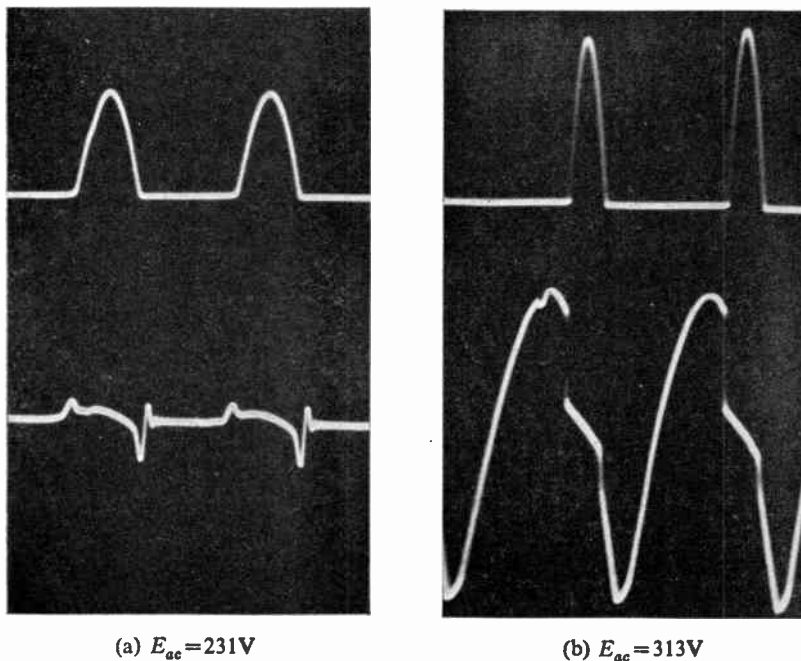

(a) $E_{ac} = 231$V   (b) $E_{ac} = 313$V

Fig. 12. Double-beam oscilloscope traces of rectifier current (upper traces) and transductor voltage (lower traces) $I_0 = 1$A.
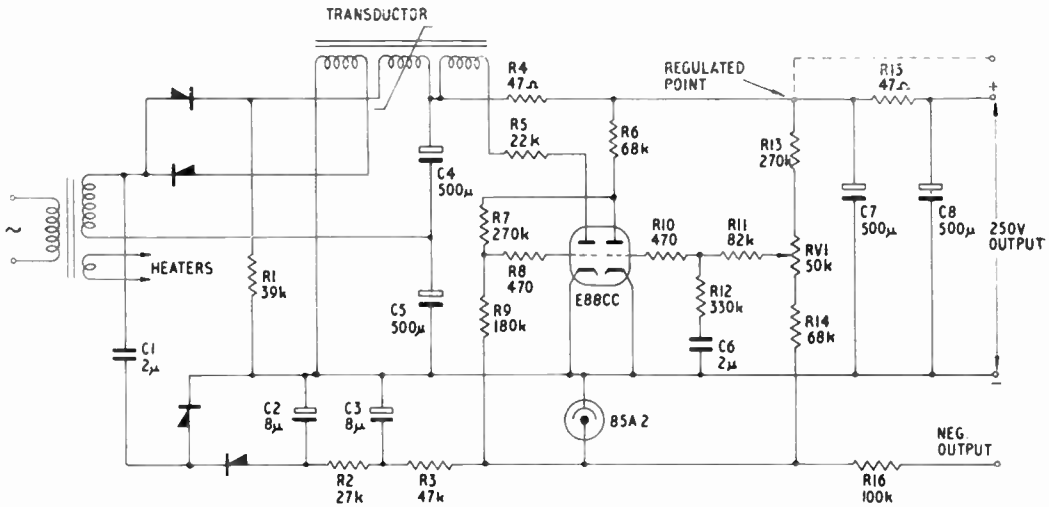
Fig. 13. Simple transductor-regulated power supply.

The power supply is rated at 250V, 400mA maximum output. Whilst the final output is not regulated for load current changes, by readjustment of the h.t. voltage control the output of 250V can be obtained for load currents from 200 to 470mA. The curve of Fig. 14 shows the regulation characteristic in respect of a.c. supply variations.
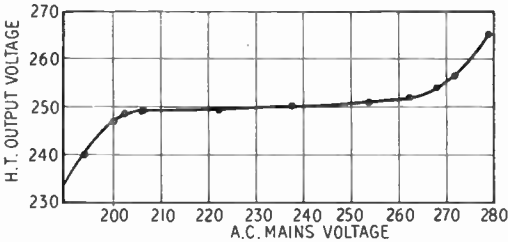

Fig. 14. Regulation characteristic of simple power supply.

4. The Fast Regulator

The principal external disturbances which the fast regulator has to deal with are step-changes in mains voltage and rapid variations in load current, consisting mainly of load current modulation (cyclic variations) but including some step-changes. In terms of frequency response the regulator must cover a range extending to well below the upper limit of the transductor system which is about 20 c/s with a 50 c/s mains supply,

in order to have a satisfactory overall transient performance. The upper limit will depend on the needs of the load apparatus; for most purposes it will be sufficient for the response to extend to about 10 kc/s, with a suitable capacitor shunting the output terminals to provide a low impedance at higher frequencies.

The basic circuit diagram of the fast regulator is shown in Fig. 15. V1 and V2 are the regulator valves, connected in series for d.c., and feeding into the power supply output terminals by means of a choke-capacitor coupling. The valve grids are driven in opposite polarity from the d.c. amplifier shared with the control valve. V1 and V2 are operated under Class B conditions to achieve high efficiency, and, in the absence of external disturbances, a small quiescent current, in the region of 4 per cent of the load current, flows in each valve. This current is sufficient to ensure that the ripple output is kept suitably low. The anode supply is taken directly from the reservoir capacitor, and under quiescent conditions the anode-to-cathode potential of each valve is approximately half that at the reservoir. The coupling capacitor is therefore normally charged to about half the output voltage.

All variations in voltage occurring at the output terminals are fed via the d.c. amplifier to both the control valve and V1 and V2, but since there are a.c. couplings in the fast regulator circuit, slow changes affect only the control valve,

and regulation for these is effected by the transductor. Rapid changes, whilst they reach the control valve, have little effect on the transductor.

In operation, if a rapid rise in output voltage tends to occur the grid of V2 is driven positively, and current is drawn from the output of the supply via the coupling capacitor. Since this current cannot immediately be supplied from the reservoir capacitor through the choke, the output voltage tends to fall as required to correct the error. The charge on the coupling capacitor is increased in the process, so that at the end of the transient flow of current in V2, a current flows in V1, to restore the charge to its quiescent value.

the case of alternating components of load current, the half-cycle swing above the mean value flows from the reservoir capacitor through V1 and the coupling capacitor to the load. An additional drain upon the reservoir is thus involved and should be allowed for in design. The transductor regulator loop acts to prevent the mean voltage falling as a consequence of the increased drain. The swing of current below the mean flows via the coupling capacitor and V2; no further drain on the reservoir is involved. The peak current in each valve is thus equal to the peak value of the modulation current, and the power dissipations in each valve are half the power drain from the reservoir which is
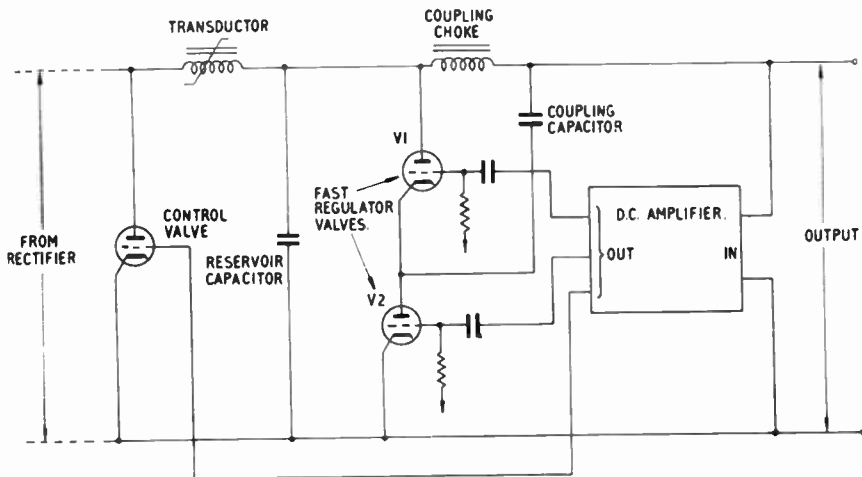

Fig. 15.   Basic fast regulator circuit.

If the output voltage tends to fall a similar action occurs; in this case V1 is driven positively and current is fed to the output via this valve and the coupling capacitor, so that the output voltage tends to rise, as required.

As the whole system is degenerative, residual deviations of the output voltage from the correct value must remain, as a consequence of the establishment of equilibrium; sufficient gain is provided so as to reduce the deviations to an acceptable level.

4.1. *Load Current Variations*

Since the output voltage is to remain constant despite rapid variations in load current, the varying components of current must flow in the regulator valves, rather than in the choke. In

$$P = \int ei.dt = \frac{E_0}{2\pi}\int_0^\pi i_m.dt \quad \dots\dots(28)$$

where P is in watts; E_0 is the mean reservoir voltage, as before, and i_m the instantaneous modulation current. Thus for sine wave modulation the dissipation in each valve is

$$P_a = \frac{E_0 I_m}{2\pi} \quad \dots\dots(29)$$

or for square wave modulation

$$P_a = \frac{E_0 I_m}{4} \quad \dots\dots(30)$$

where I_m is the peak value of i_m in each case.

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For step-changes of load current the appropriate valve must initially supply, or accept, the incremental current, until the transductor effects the necessary change in the choke current. The duration of valve current flow is thus equal to the overall transductor response time, noted in Section 3.13, and depends on the instant, relative to the instantaneous a.c. input voltage, at which the change occurs. The shape of the valve current pulse also depends on the relative timing.

4.2. *Mains Voltage Variations*

If a step-change of mains voltage occurs, then before the transductor can operate the reservoir voltage will begin to change. Since the output voltage is to remain constant the mean choke current must not change and the appropriate regulator valve must therefore supply a current equal to the transient current in the choke. When the transductor comes into operation and corrects the deviation of the mean reservoir voltage, the fast-regulator valve current returns to the quiescent value. In a 1A 250V power supply, where the choke inductance is 1 henry, the reservoir capacitance and coupling capacitance each $500\mu F$, the upper valve peak current resulting from a 5 per cent drop in mains voltage is about 80mA, varying slightly according to the relative timing of the a.c. wave and the step. The peak current in the lower valve for a 5 per cent upward step is about 75mA. These figures are quoted as a satisfactory general solution has not yet been found.

4.3. *Ripple Current*

As the ripple voltage at the output terminals is to be very small compared with that at the reservoir capacitor, substantially the full reservoir ripple voltage appears across the choke, and the current which flows as a result must be supplied by the regulator valves. Taking the reservoir discharge current as constant and equal to I_0 when the rectifiers are not conducting, the peak-to-peak ripple voltage is given by

$$E_r = I_0.t_r/C_R \qquad \ldots\ldots\ldots(31)$$

where t_r is the time remaining between period of rectifier full conduction and can be taken at $0.7T$; $T = 1/2f$. C_R is the reservoir capacitance. Assuming the ripple wave to be of sawtooth shape the peak-to-peak choke current is

$$I_{\iota}p = \frac{E_r T}{8L} \qquad \ldots\ldots\ldots(32)$$

where L is the choke inductance, the choke resistance is neglected as the intrinsic choke time-constant will generally be large compared with T. The choke current is parabolic in form, so that the positive peak value is $\frac{2}{3}I_{\iota}p$ and the negative $\frac{1}{3}I_{\iota}p$. It will in most cases be necessary to choose a quiescent current greater than would be sufficient to supply the choke current, as the mutual conductance of valves tends to be low at low anode currents, and the current may have to be chosen on a basis of minimum tolerable loop gain.

4.4. *Coupling Choke and Capacitor*

The resonant frequency of these components should be well below the upper-frequency limit of the transductor regulator, for as resonance is approached the demand of reactive current on the regulator valves becomes heavy. Also the stability of the fast regulator loop is aided by the overriding influence of the transductor, providing that the region of critical phase shift is within the transductor range.

The source of anode voltage for the regulator valves in the course of a transient is the charge on the coupling capacitor, and this should therefore be large enough to ensure adequate anode voltage during transients.

5. A Complete Power Supply Unit

Figure 16 is the circuit diagram of a stabilized power supply providing 1A output at 250V, embodying the transductor and fast regulators described. The circuit is substantially that of a production unit designed for use with a television camera, with some subsidiary features omitted. The unit was developed for production by Mr. W. T. Underhill.

The transductor used is of the type described in Section 3.14. One 6080 double-triode, an improved version of the 6AS7, widely used as a series-regulator valve, provides the two triode sections required for the fast regulator.

5.1. *Amplifier Details*

The d.c. amplifier includes V1A and B and V3A and B; V2 is the reference source stabilizer. The balanced form of the first stage is chosen
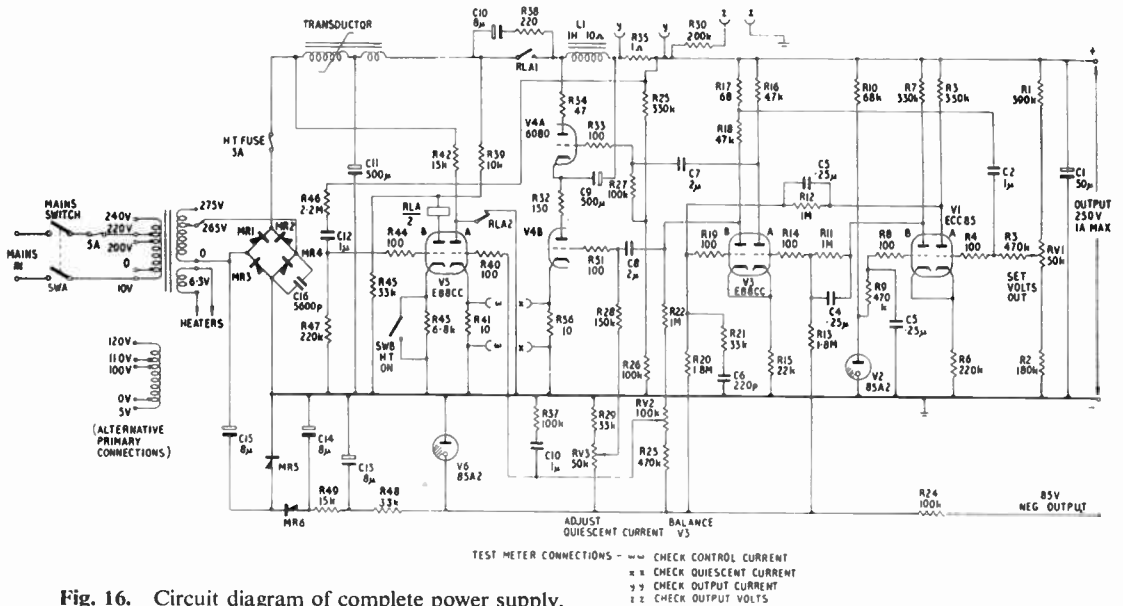
Fig. 16. Circuit diagram of complete power supply.

TEST METER CONNECTIONS – ᵚ ᵚ CHECK CONTROL CURRENT
      ᵡ ᵡ CHECK QUIESCENT CURRENT
      ᵞ ᵞ CHECK OUTPUT CURRENT
      ᶻ ᶻ CHECK OUTPUT VOLTS

in order to minimize the effect of heater-voltage variation on the output voltage stability, and it is convenient to extend the balanced form also to the second stage as a push-pull output is required to drive the fast regulator valves, V4A and B. The drive for the transductor control valve V5A is taken from the anode of V3B via R22 and RV2.

As the fast regulator valves require a much greater driving voltage than does the control valve, steps are taken in the amplifier to increase the a.c. gain. These include the shunting of the coupling resistors R11 and R12 with capacitors, and the provision of internal positive feedback. The feedback is obtained by connecting C2 at the grid of V1A to the junction of R17 and R18. Ordinarily this capacitor would be connected to the h.t. line with the object of eliminating the loss, for a.c. components, of the input potentiometer, R1, RV1 and R2; as shown it also serves for positive feedback coupling.

The components R21 and C6 are for high-frequency phase correction; the need for this arises from the phase-shift introduced by the stray capacitance associated with each anode load which, without correction, would in the present case cause oscillation at a frequency in the region of 50 kc/s. R37 and C10 in the grid circuit of the control valve are for compensation

of the amplifier loop in respect of the delay of the transductor regulator.

5.2. Controls

The output voltage control, RV1, is similar to that of a conventional supply. RV2 needs some comment, however. Any change in the d.c. potential at the grid of V1A will cause the anode current of one half of V3 to rise and the other to fall. The permissible range of current here is somewhat limited by the need of a large anode voltage swing for driving V4A and B. Whilst the current variation is normally small, the control was included as a precaution against unfavourable production component variations. It now seems likely that it could be eliminated. The control, being within the feedback loop, produces, by a very small change of output voltage, the desired differential variation of V3A and B anode currents.

RV3 is used for initial setting up of the quiescent current in V4A and B; the valves being in series for d.c. and the upper valve grid-to-earth potential being fixed by R25 and R26, it suffices to adjust the bias on the lower valve. The h.t. switch in the cathode circuit of V5B controls the operation of RLA, and the shunt resistor, R43, serves to limit the rise of cathode potential when the switch is off.

5.3. *Time-delay Relay Circuit*

This performs the usual function of delaying the application of h.t. until the valves in the power supply and comprising the load have warmed-up. The delay is provided by the heating time of V5B, and the action of the relay is made more decisive by the connection of C12 and R46 to the h.t. line on the output side of the contact RLA/1. An incidental but useful feature of this latter arrangement is that a momentary short circuit of the supply output terminals, in depressing the grid potential of the valve via C12 and R46, causes the relay to be de-energized and RLA/1 opens, removing the h.t. The relay re-operates after a delay of a few seconds, and in the case of a persistent fault, the h.t. fuse ultimately blows.

The second contact, RLA/2, closed when the relay is not energized, is used to connect the resistor R42 to earth. This resistor draws a large control current through the transductor so that the unloaded reservoir voltage during the warming-up period is reduced. The advantage of this arrangement is, of course, that the reservoir capacitor is not subjected to a high surge, and the voltage initially applied via RLA/1, when it closes, is less. The resistor R42 also serves to limit the anode dissipation of V5A under overload conditions.

The resistors R32 and R34 are to limit the peak currents in V4A and B occurring when the contact RLA/1 closes, or when the output is short-circuited. They are an unfortunate necessity, for they also tend to limit the peak current available for regulating purposes.

5.4. *Negative Bias Supply*

To avoid the need of a separate winding on the mains transformer for the bias supply, the voltage-doubler bias rectifiers MR5 and MR6 are fed from the same winding as supplies the load. As the connection of this rectifier system unbalances the bridge rectifier, the capacitor C16 is used to restore the balance, the need for which is mentioned in Section 3.8.

5.5. *Performance*

Output voltage:	250V
Maximum mean output current:	1·12A
Mains voltage range:	±7·5%
Load current range, mean:	±125mA
Load current range, mean (mains voltage range ± 3%):	100mA to 1A
Load current modulation, including step-changes: Total excursion:	250mA
Output voltage stability (excluding variations in reference voltage stabiliser):	0·05%
Output impedance, d.c.:	0·1 ohm
25 c/s to 2kc/s:	0·05 ohm
10kc/s:	0·3 ohm
Input V-A:	620

5.6. *Comparison with Conventional Series-valve Supply*

	Series-valve	Transductor
Rectifier valves:	6	(metal)
Regulator valves:	5	1
Small valves:	3	5
Internally dissipated power:	430W	150W
Input V-A:	980	620
Chassis height in standard 19 in. rack:	15¾in.	8¾in.

It is appreciated that the series-valve supply does not represent the best of its kind; it is nevertheless representative of units at present widely used.

6. Other Regulator Combinations

The regulators described here may be of use in combination with other types of regulator. An interesting possibility is the addition of a transductor system to a series-valve regulator. With the supply stabilized for slow changes by the transductor, the burden on the series valves would be considerably reduced, and their number or capacity could be reduced for a given output current, with the addition of a shunt resistor to carry part of the load current.

7. Conclusions

The combination of transductor and class-B fast regulators may offer advantages over previous techniques in stabilized power supply design, particularly in respect of efficiency and size, and where large output currents are

required. A high order of reliability should be expected, for the transductor is a relatively robust component, and the demand made upon the fast-regulator valves will generally not be heavy under average conditions. On the other hand, design is somewhat difficult, and some features need special attention, as in the choice of rectifiers, for example.

8. Acknowledgments

The author acknowledges the large contribution made by his colleague, Mr. W. T. Underhill, particularly in the experimental work and in many discussions. The work was initiated by Mr. E. Davies, to whom thanks are due for his guidance. The author is indebted to Marconi's Wireless Telegraph Company Ltd., for permission to publish this paper.

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10. Appendix : List of Symbols

A	Effective core area, cm^2	N	Number of turns
B	Flux density, gauss	P	Power, watts
B_{SAT}	Saturation flux density	P_a	Anode dissipation, watts
C_R	Reservoir capacitance, farads	R	Total rectifier circuit resistance
e	Instantaneous e.m.f.	R_r	Rectifier resistance
e_I	Effective inst. rectifier circuit e.m.f.	R_s	Source resistance
e_c	Instantaneous control period e.m.f.	R_t	Transductor winding resistance
E_a	Rectifier circuit average voltage.	R_l	Load resistance
E_{ac}	R.M.S. input voltage to rectifier	S	Storage, or voltage-time integral, volt-seconds
E_0	Mean reservoir voltage		
E_p	Peak input voltage to rectifier	S_e	Effective storage
E_r	Peak-to-peak ripple voltage	S_{max}	Maximum usable storage.
f	Mains frequency, c/s	S_t	Transductor storage
H	Magnetizing force, oersteds	t	Time, non-specific
H_c	Coercive force, oersteds	t_r	Interval between periods of rectifier full-conduction
i	Instantaneous current		
i_m	Instantaneous load modulation current	T	Time of one half-cycle, $1/2f$
I_m	Maximum or peak value of above	V_r	Rectifier apparent internal e.m.f.
I_0	Mean load current	α	Angle, $\cos \alpha = E_0/E_p$
I_{rms}	R.M.S. rectifier current	β	Saturation angle
I_p	Peak rectifier current	θ	Angle, non-specific
I_c	Mean control current (control period)	φ	Flux, maxwells
I_{LP}	Peak-to-peak choke current		
l	Effective core length, cm.		

The suffices $_1$ and $_2$ denote minimum and maximum values respectively.

APPLICANTS FOR ELECTION AND TRANSFER

As a result of its January meeting the Membership Committee recommended to the Council the following elections and transfers.

In accordance with a resolution of Council, and in the absence of any objections, the election and 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 Associate Member

APPERLEY, Eric Charles, B.Sc. *Gloucester.*
BURR, Albert Edward. *Boreham Wood, Herts.*
DAVIES, John Jenkyn. *Westcliff-on-Sea, Essex.*
GIBSON, Ernest. *Purley, Surrey.*
KHAN, Capt. Tasdique Amin, B.Sc., Pak.E.M.E. *Karachi.*
KING, Desmond. *Chelmsford, Essex.*
TOWNSEND, Lt.-Col. Edward Walter, R.Sigs. *London, S.E.21.*

Transfer from Associate to Associate Member

BENEDETTI, Edmund Joseph, M.Sc., B.A.(Hons.). *Bombay.*
WILSON, George Raymond Briscoe. *Whitehaven, Cumberland.*
WRIGHT, Frank William. *Kirkuk, Iraq.*

Transfer from Graduate to Associate Member

BARNARD, Clifford Albert. *Haslingfield, Cambs.*
BECKER, Simcha. *Givataim, Israel.*
BOICE, Capt. Cyril John. R.E.M.E. *Monk Moors, Cumberland.*
CHAUDHURI, Capt. Suhas Chandra, B.Sc.(Hons.), Ind.Sigs.. *Ranchi.*
DRUMMOND, Gilbert Park. *Zanzibar, East Africa.*
ELLIS, Brian Norman. *Stockton-on-Tees, Co. Durham.*
KUMAR, Satyendranath, M.Sc. *Bangalore, India.*
MELCIO, Joseph Albert. *Croydon.*
POTTS, Kenneth Harry. *Norwich, Norfolk.*
RODDY, Dennis. *Cambridge.*
ROSS, Clive William. *Barrhead, Renfrewshire.*
STERRY, Peter John. *Basingstoke, Hants.*
WADVA, Kuldip Singh, B.Sc. *London, N.8.*

Transfer from Student to Associate Member

LANGHOLZ, Hillel. *Ramat Gan, Israel.*
NOWICKI, Jerzy Ryszard. *Totton, Hants.*
STANLEY, Anthony Russell. *Cheam, Surrey.*

Transfer from Student to Associate

BENNETT, Peter Allen. *Sheffield, 6.*
SALISBURY, Henry Nance. *Sandown, I.O.W.*
WALLER, Dennis John. *Redditch, Worcs.*

Direct Election to Associate

BATES, Dennis Noel Frederick. *Birmingham.*
BIRD, Harold Richard Hughes. *Bridgnorth, Salop.*
BOWMAN, Alfred. *Canvey Island, Essex.*
LARSEN, John Wilhelm. *London, W.8.*
MORGAN, Wg.-Cdr. John Anthony Felix. R.A.F. *Maidenhead.*
WEDLOCK, Eric Herbert. *Kano, Nigeria.*
WILLIAMS, Clifford. *Cheltenham, Glos.*

Direct Election to Graduate

ARNOLD, John Spencer. *Liverpool.*
BENTHIN, Louis. *Brookmans Park, Herts.*
*BROADBERRY, Noel Edward. *Southampton.*
BURT, Dennis William. *St. Bees, Cumberland.*
CHURCHWARD, Reginald John. *Plymouth, Devon.*
GILL, John Oliver. *Chelmsford, Essex.*
HOLD, Bernard Henry. *Warrington, Lancashire.*
HUNT, Desmond Gerald. *Birmingham.*
JAMIESON, Clifford Park. *Liverpool.*
LANE, Colin. *St. Albans, Herts.*
McCUBBING, Eric Corrie. *Carlisle, Cumberland.*
OWENS, Victor Thomas. *Hatfield, Herts.*
REED, Alec Raymond. *Chelmsford, Essex.*
SELLERS, Walter Alexander George. *Harrow Weald, Middlesex.*
SHARPE, Robert Brittain. *Twickenham, Middlesex.*

Transfer from Associate to Graduate

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CRAPPER, David Hugh. *Portsmouth, Hants.*
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HARTWELL, Edward Harman. *Crawley, Sussex.*
NICHOLAS, Russell John. *Southampton.*
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The following 69 students were registered at the November and December meetings of the Committee. The names of a further 65 students registered at the December and January meetings will be published later.

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RICHARDSON, Geoffrey Donald. *Lae, New Guinea.*
ROIZIS, Mordecai. *Tel Aviv, Israel.*
ROLF, John Leslie. *Southwick, Sussex.*

SAVAGE, Kevin James. *London, N.1.*
SMITH, Derek Edward. *Luton, Beds.*
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BOCKING, Colin Stanley. *Colchester.*
BREARLEY, Harry. *Pudsey, Yorks.*
BREARLEY, Malcolm. *Pudsey, Yorks.*
CALLENDER, Peter C. *Fareham, Hants.*
CLARKE, Eric. *Guildford, Surrey.*
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GIBSON, Alfred. *Belfast.*
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GOPAL, Subramaniam, B.Sc. *Bombay.*
GRABER, Siegmund S. *Givataim, Israel.*
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* Reinstatement.

A Timing-Pulse Generator †

by

C. S. FOWLER, ASSOCIATE MEMBER‡

Summary : The equipment described generates timing pulses for general laboratory use. By reference to the MSF standard frequency transmissions, the pulses serve as time signals which do not differ from Universal Time (UT 2) by more than ± 0.05 millisec.

1. Introduction

In studies of radio wave propagation it is frequently necessary to superimpose timing signals on the recorded data obtained at the receiving station. This requirement is particularly important in the reception of signals transmitted from earth satellites. Ideally, the records obtained should include timing signals of such a nature that any particular feature of interest can be related directly to Universal Time.

Universal Time or, more correctly, Universal Time 2 is used in scientific and astronomical work. It is based, as is G.M.T., on the corrected rotational period of the earth but is measured from midnight to midnight, whereas G.M.T. is taken from midday to midday. The difference between the uncorrected rotational period UT 0 and UT 2 is between $\pm$ 20-30 milliseconds a year depending on the location of the observatory.

In the United Kingdom, suitable time signals for this purpose are provided by the MSF transmissions from Rugby on 2·5, 5, and 10 Mc/s.[1] Step corrections are made to these signals when the departure from Universal Time reaches a specified value;[2] announcements of such corrections are made monthly. Due to the interrupted nature of the transmissions and the low signal-to-noise ratio usually obtained at the receiver, these signals cannot be used directly as continuous time markers. They can, however, be compared against locally-generated pulses, and any drift of the latter, relative to the

MSF signals, can then be corrected as required. The equipment described below generates timing pulses, for use in this application, having time intervals of 0·1, 1·0, 10 and 60 seconds with pulse widths of 0·5, 1·25, 3 and 6 milliseconds respectively.

2. General Description

A block diagram of the apparatus is given in Fig. 1. A stable 100-kc/s oscillator is used as the local frequency standard, having a long term stability of a few parts in 10^8.

The equipment incorporates a series of counter units of commercial design, with the final unit modified to divide by six to give pulses every sixty seconds. The output of the second decade counter is fed to a tuned 1-kc/s amplifier to provide the phase-shifting magslip with a sine-wave input. The output of the magslip is converted in the pulse shaping unit to pulses of the required amplitude and shape to trigger the next counter circuit. A mechanical counter coupled to the magslip drive enables a record to be kept of the phase shift required to keep the locally-generated pulses in step with the reference transmissions.

The time relationship of the MSF second pulses and the local pulses is displayed on a 100-c/s c.r.t. time-base, the commencement of which is delayed with respect to the output pulses of the 100-c/s decade. Since the delay between the commencement of the time-base trace and the locally-generated pulses is fixed, the reference pulses move along the trace when the phase shifter is operated.

The outputs from the last four counter circuits are used to trigger monostable multivibrator circuits which in turn produce square waves of the required duration. These are combined
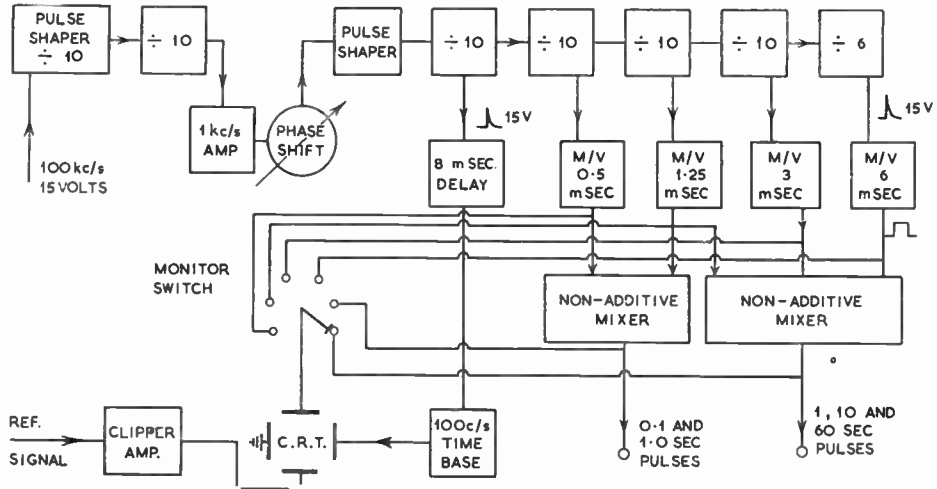
Fig. 1. Block diagram of timing-pulse generator.

in two non-additive mixer circuits. The first provides pulses with time intervals of 0·1 and 1 second; the second mixer produces pulses with time intervals of 1, 10 and 60 seconds.

Both these combined outputs are of one volt negative polarity. The decrease in amplitude of the one minute pulses is 10 per cent. when feeding a load of 100Ω in parallel with 0·1 μF. This represents the maximum loading due to the cable network used to distribute the timing pulses. Since the signals from MSF vary in amplitude, and are often only twice the noise level, provision is made to limit the positive half cycles between two adjustable levels, both of which are above noise. These pulses are then used to drive a tuned 1-kc/s amplifier the output of which is displayed on the c.r.t.

The Q factors of the 1-kc/s tuned circuits used in the reference signal chain are low and the delay of the final displayed pulse compared to the MSF signal is approximately 0·05 milliseconds. If greater accuracy than 0·05 milliseconds is required particular attention must be paid to circuit delays in the reference signal chain, and to changes due to propagation effects. It will be appreciated that the circuit arrangements are flexible and any desired combination of pulses of any desired width can be obtained.

Width modulation of the pulses was chosen as the most suitable method for differentiating between pulses for the present application. Additive-type mixer circuits could also be used

and the pulses identified by differences in amplitude: in this design the multivibrator circuits would not necessarily be required.

In use the pulses are recorded on tape or used to operate a marker pen on a chart recorder. The date, hour and minute of commencement is either recorded verbally on the tape or written on the chart against the appropriate minute pulse.

3. Conclusion

The technique described provides, in an instrument of relatively simple design, timing signals for general laboratory use. The equipment has been in continuous operation for 18 months and has also proved useful in checking the long-term drift of the 100-kc/s crystal oscillator used to drive the pulse generator.

4. Acknowledgments

5. References

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Reduction of Television Bandwidth by Frequency-Interlace†

by

E. A. HOWSON, B.SC., PH.D. and D. A. BELL, M.A., PH.D.‡

A paper read on 1st July 1959 during the Institution's Convention at Cambridge.

Summary : A method analogous to the N.T.S.C. colour television system is used to obtain a bandwidth reduction of a black-and-white video signal by a factor of approximately 2 : 1. The normal signal is split into two frequency bands, nominally zero to 1·5 Mc/s and 1·5 to 3·0 Mc/s. The latter is used to amplitude-modulate a sub-carrier, whose frequency is an odd multiple of half the line scanning rate. The lower sideband of the modulator output is selected and combined with the original zero-to-1·5 Mc/s band, so that the spectra of the two signals interleave. The combined signal may now be sent over a channel of 1·5 Mc/s nominal bandwidth. At the receiving end of the channel the composite signal is applied to a synchronous demodulator, fed also with sub-carrier of the same frequency as at the transmitter. The lower sideband of this demodulator is taken and combined with the received signal, to yield a "normal" video signal extending from zero to approximately 3 Mc/s, together with an "interleaved" signal. The interleaved signal is such as to give an interference pattern on the display which in a stationary picture should optically cancel after four successive frame scans. However, the pattern is built up in such a way as to give rise to a "crawling" motion which is very noticeable at close viewing distances. Photographs of typical pictures obtained with an experimental apparatus are given, showing various interference effects produced.

1. Introduction

The bandwidth required for the faithful transmission of a television signal generated by conventional scanning processes is somewhat extravagant, and since the inception of television broadcasting methods have been sought of reducing this bandwidth while maintaining adequate picture quality.

The first major reduction in bandwidth was achieved by two-fold interlacing of lines, which is possible because the minimum rate of frame§ scanning is governed by visibility of flicker and not by the requirements of information transmission. The extension of the interlace principle, whether to fourfold line interlace or to dot interlace systems, gives rise to "crawling" effects at the edges in the picture. Various other devices or bandwidth economy have been reviewed by one of the authors.[1] But the only system which has proved to be acceptable in

picture quality and practicable in terms of apparatus required at the receiver is the N.T.S.C. colour system using frequency-interlacing in order to accommodate colour information within the bandwidth of a conventional black-and-white picture signal. The present paper reports an investigation into the application of this same principle to the reduction of bandwidth in a black-and-white picture signal.

2. Systems involving Redistribution of the Signal Spectrum

2.1. *The Nature of the Spectrum of a Line-interlaced Signal*

A number of theoretical and practical investigations have been made [2,3] of the energy spectrum of a television signal. In general, the waveform obtained on scanning a blank scene can be analysed into a series of harmonics of the line scanning frequency, whose amplitudes

§ "Frame" in the British usage, as distinct from U.S.A., refers to the area which is generally covered by the scanning raster, not to a completely scanned picture.

vary inversely as their order. When a normal stationary picture is scanned, the brightness signal is repetitive at the picture scanning frequency, and may be thought of as modulating the amplitude of the "blank raster" waveform. Accordingly each line harmonic has associated with it a pair of sidebands, each consisting of a series of harmonics of the picture frequency. The general trend of the spectrum is shown in Fig. 1, resulting from measurements made on typical 405-line video signals.
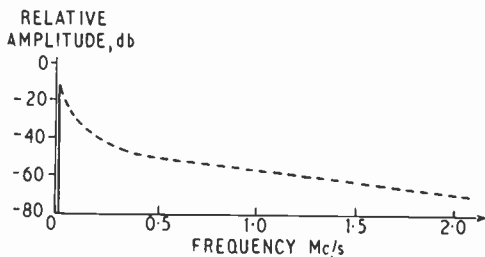

Fig. 1. Variation in amplitudes of line harmonics.

If it were required to transmit a 405-line picture at a rate of fifty pictures per second, the line scanning frequency would be 20·25 kc/s. Accordingly, the energy spectrum would consist of a series of line harmonics at a spacing of 20·25 kc/s, each having a pair of sidebands consisting of a set of picture harmonics at 50 c/s intervals.

However, suppose 2 : 1 line interlace is used and the frame rate is maintained at 50 c/s; the line frequency is now 10·125 kc/s, which is not an integral multiple of 50 c/s. Although the overall picture repetition frequency is 25 c/s, successive frames differ only slightly, and so the picture frequency is still preponderantly 50 c/s. Accordingly each line harmonic still has its sidebands of 50 c/s harmonics, but in the region midway between any two line harmonics, their respective sidebands frequency-interlace, the spectral lines occurring at 25 c/s intervals. A full development of the 25 c/s harmonics would occur with a picture having alternate black and white lines, which on scanning yields a 25 c/s picture repetition frequency.

As the line scanning rate has been halved the bandwidth required to transmit a picture of given vertical detail has been halved also.

A minor disadvantage is that two frame scans are required for the complete transmission of a picture; also under certain conditions of picture movement the display appears to lose interlace.

2.2. The "Frequency-interlacing" of Additional Signals

Many proposals have been made to utilize the "spaces" between the spectral lines of a video signal for the transmission of additional signals.[4] The frequency of such a signal is arranged to fall midway between two adjacent spectral lines; this frequency does not occur in a stationary picture, but does give rise to a spurious pattern which is fully developed after two frame scans in a line-interlace system. However, during the two succeeding scans, the phase of the pattern is reversed, so that after four scans complete cancellation occurs.

Unfortunately if the amplitude of this frequency-interlaced signal is increased beyond a certain level, the optical cancellation is incomplete, an effect due in part to the non-linearity of the cathode-ray tube characteristic. However the main reason is that in successive frame scans, the spurious pattern is displaced vertically by one scanning-line spacing, and so a crawling motion is apparent. The brightness level at which the crawling motion is noticeable varies with the frequency of the interleaved signal, high frequencies being less obvious.

If it is required to transmit a signal of appreciable bandwidth by a frequency-interlace technique, not only must its individual frequency components interlace with those of the main signal, but also high-level interleaved components should be arranged to occur in the high-frequency region of the main band.

A second television signal is thus suitable for frequency-interlacing. It may be used to amplitude - modulate a frequency - interlaced "sub-carrier" situated near the high frequency limit of the main video band. Minimum interference is caused to the main signal if the sub-carrier is placed midway between two adjacent line harmonics, by arranging its frequency to be an odd multiple of half the line scanning rate. It should be noticed that this also causes it to interlace with the various frame harmonics.

2.3. *The N.T.S.C. Colour Television System*

A typical example of a frequency-interlace system transmitting additional video signals in the band occupied by a normal signal, is the N.T.S.C. colour system.[5]

The "main" signal in this case is the brightness or "luminance" signal, and the additional signal comprises two "chrominance" signals which are sent on one sub-carrier by a simple polyphase modulation process. However, to avoid cross-talk between the chrominance
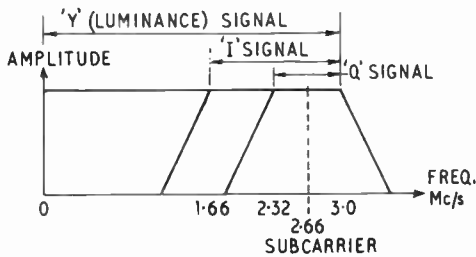
Fig. 2. Utilization of video bandwidth by "British N.T.S.C." system. Video bandwidth of "I" signal = 1·0 Mc/s, of "Q" signal = 0·34 Mc/s.

signals, both sidebands of the sub-carrier have to be transmitted, but advantage is taken of the eye's lack of perception of fine detail of certain colours to curtail the bandwidths of the chrominance signals. Double sideband transmission is then only necessary in the region where the two chrominance signals overlap, and the remainder of the upper sideband of the wide-band chrominance signal may be attenuated. Figure 2 illustrates the utilization of video bandwidth by the British modification of the N.T.S.C. system, and shows how the subcarrier is sited as high as possible in the frequency band.

The mutual interference between luminance and chrominance signals is considerably reduced by the restricted chrominance bandwidth. High-level luminance interference may be filtered out of the chrominance signals, and high-level chrominance interference occurs only over a relatively narrow bandwidth of the luminance signal.

A very important feature of the N.T.S.C. system is that it is "compatible" in that reception of a "colour" signal by a "black-and-white" receiver will yield a correct black and

white display, at the expense of a certain amount of interference. Compatibility could still be achieved by the transmission of the colour information in a band separate from that of the luminance, but the N.T.S.C. system yields an adequate performance with bandwidth economy.

2.4. *Bandwidth Reduction of "Black-and-White" Signals*

It has been suggested that a modification of the above technique might be used to obtain a reduction in bandwidth of a conventional black-and-white video signal.

The sub-carrier is set near the upper limit of the normal video band. The video signal is divided into two equal half-bands, as illustrated in Fig. 3 (a), and the upper half-band is used to modulate the sub-carrier. The lower sideband of the modulator output is selected and combined with the original lower half-band; the resultant signal contains all the original signal information, but in half the bandwidth, as shown in Fig. 3 (b).
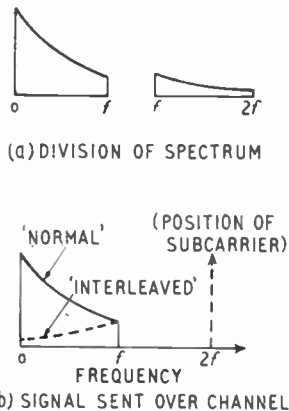
(a) DIVISION OF SPECTRUM

(b) SIGNAL SENT OVER CHANNEL

Fig. 3. Bandwidth reduction process at transmitter.

This frequency-interlaced composite signal may then be transmitted over a reduced bandwidth channel. However, instead of dividing the main video signal into two half-bands, and modulating the sub-carrier with the high-frequency half-band only, the entire main signal is used. The lower sideband in the modulator output contains the required interleaved signal in correct frequency relationship with the main signal. If the modulator output is added to the main signal and the result passed through a

low-pass filter cutting off at approximately half sub-carrier frequency, the filter output consists of the correct composite reduced-bandwidth signal. This system avoids the use of complementary low- and band-pass filters.

Figure 4 (a) illustrates the recovery of the high-frequency components of the main signal at the receiver. The composite signal is applied to a second modulator which is also supplied with sub-carrier. The lower sideband of the output of this modulator consists of the main "high" half-band, together with an interleaved signal corresponding to the original "low" half-band. When this is added to the received signal, the overall effect is to give a full band-width main signal and a frequency-inverted and interleaved interfering signal, as in Fig. 4 (b).
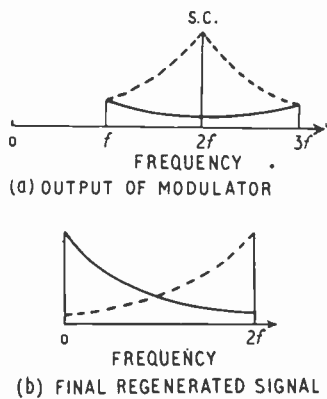
Fig. 4. Recovery of high frequency band at receiver.

The interference produced by the inter-leaved signal in this system is somewhat distracting, but can be reduced considerably by eliminating the sub-carrier itself by means of a balanced modulator. The large-amplitude side-bands of the sub-carrier, corresponding to low frequencies in the main signal, may be attenuated by means of suitable filters; some of the difficulties which arise in reducing the interference are considered in the next section.

It should be noted that the dot-interlace system mentioned in Section 1, although developed as a special case of time division multiplex, gives rise to an energy spectrum essentially similar to that of this frequency-interlace system.[4]

3. Special Features of the Frequency-Interlace System

3.1. Synchronization

In a practical system there arises the problem of maintaining synchronism between transmitter and receiver. The correct sub-carrier phase at the receiver must be maintained accurately as otherwise the phases of the high-frequency components of the regenerated full-bandwidth signal will vary with respect to the low-frequency components. The expedient used in the N.T.S.C. colour system, of transmitting a "burst" of several cycles of sub-carrier after each line sync.-pulse period, which is used as a reference signal for the control of a local oscillator, cannot be adopted; for such a "burst" would be outside the pass-band of the transmission channel.

However, it should prove possible to send a burst of some sub-multiple of the sub-carrier, and arrange a locked oscillator and frequency multiplier at the receiver to generate the local sub-carrier.

A second problem arises due to possible interference between the interleaved signal and the sync.-pulses of the normal signal. When the modulated sub-carrier is applied to the transmission filter, a pronounced "ring" occurs in the output of the filter at times corresponding to the original sync.-pulse edges. This ring is superimposed upon the normal sync.-pulses, whose rise-time has been approximately doubled by the filter. As the major peak of the filter ringing is "dot-interlaced", the composite signal as received over the channel cannot be used for generating accurate sync.-pulses. If such a waveform were applied in the usual manner to the line time-base generator, the time-base would probably tend to synchronize on to the "ring"; and accordingly cancellation of interfering signals in the display would be seriously affected.

In order to overcome this difficulty, the normal video signal can be passed through a clipper, to remove the sync.-pulses, before application to the first modulator. At the receiver, the synchronizing information is extracted before the regeneration process takes place. Although the rise-time of the sync.-pulses is longer than usual, they are uncontaminated

by the interleaved signal, as that signal does not now exist during the pulse period.

3.2. Effect of the Filter upon the Normal Signal

The transmission filter also causes ringing in the normal channel at sharp transitions in the video signal. However, if the cut-off frequency of the filter is carefully controlled, and its phase characteristic is linearized over its passband, these disturbances may be eliminated in the regenerated full-bandwidth signal, the final overall frequency response being uniform. This condition also requires accurate control over the phase of the sub-carrier at the receiver, and also equal overall gains in the low-frequency and high-frequency main signal paths.

3.3. Possibility of Pre-emphasis

An interesting possibility is that of using different relative levels for the main and interleaved signals. The interleaved signal may be sent at a higher level than the main signal over the narrow band channel.

At the receiver, when the overall response of the main signal is adjusted to be uniform, the interleaved signal has two levels. The components in the interleaved signal corresponding to low-amplitude high-frequencies in the main signal are emphasized and large-amplitude components in the interleaved signal are attenuated.

This should tend to reduce interference due to the sub-carrier and its neighbouring large amplitude sidebands, at the expense of increased interference from what were originally high frequencies in the main signal. This last effect is particularly troublesome in test charts and similar pictures with areas of large-amplitude fine detail.

A theoretical analysis shows that the amplitude of the major peak of ringing due to the interleaved signal is relatively unaffected by various amounts of "pre-emphasis"; although the period for which the ringing has appreciable amplitude is reduced with increased pre-emphasis. Accordingly, a bright twinkling vertical edge would be given to the various objects in the scene.

3.4. Effect of Sub-carrier Interference

In the N.T.S.C. colour system the sub-carrier and its nearby sidebands are often attenuated by a narrow-band "notch" filter.[6] This also attenuates a band of luminance fine detail, but this effect is generally outweighed by the general improvement in the picture. Normally the unattenuated sub-carrier is partially rectified by the non-linear characteristic of the display tube, and so a general increase in luminance occurs, to the detriment of the colour saturation; by the use of a suitable notch filter this effect can be reduced, but a certain amount of ringing is introduced giving rise to desaturation and dot-crawling interference at transitions in the picture.

In the bandwidth reduction system, continuous sub-carrier is eliminated by the use of a balanced modulator but components at approximately the same frequency still occur during transients, as demonstrated in Fig. 11.

However, the sub-carrier frequency need not now be inside the main video band, as in the colour system, but can be placed outside with considerable advantage; filtering of the sub-carrier may then be accomplished without unduly disturbing the high-frequency components of the regenerated main signal. The degree of bandwidth reduction is then slightly less than 2 : 1.

The upper sideband of the output of the receiver modulator consists of both interleaved and non-interleaved frequency components; the latter may in certain circumstances combine additively with the main high-frequency recovered signal and produce undesired distortion. The interleaved signal would give a very fine grain interference pattern which would be relatively invisible. However, by the use of a low-pass filter the sub-carrier and its entire upper sideband may be completely eliminated with a general reduction in interference; the filter can be phase-corrected to improve its transient response.

4. Experimental Apparatus

An experimental bandwidth reduction system has been built to investigate the various principles outlined in the preceding sections, and a simplified block diagram is shown in

Fig. 5. The B.B.C. programme radiated from the nearby Sutton Coldfield transmitter is utilized as a source of video signals, a simple t.r.f. receiver having a somewhat restricted video bandwidth of 2·6 Mc/s being used.

A conventional synchronizing-pulse separator is fed with the video signal and provides line and frame pulses for use in various other parts of the equipment.

The sub-carrier frequency used in the system must be maintained at an odd multiple of half the line frequency. However, the line frequency

2·9615625 Mc/s; this frequency is inside the 3·0 Mc/s video band, but owing to the restricted bandwidth of the receiver, no difficulties have been experienced due to signal components in the range of 2·96–3·0 Mc/s.

The sync.-pulse clipper consists essentially of a biased diode arranged to conduct the video signal only above black level. A black-level clamp is incorporated so that variations in input do not affect the clipping level; the necessary clamp pulses are derived from the line synchronizing waveform.
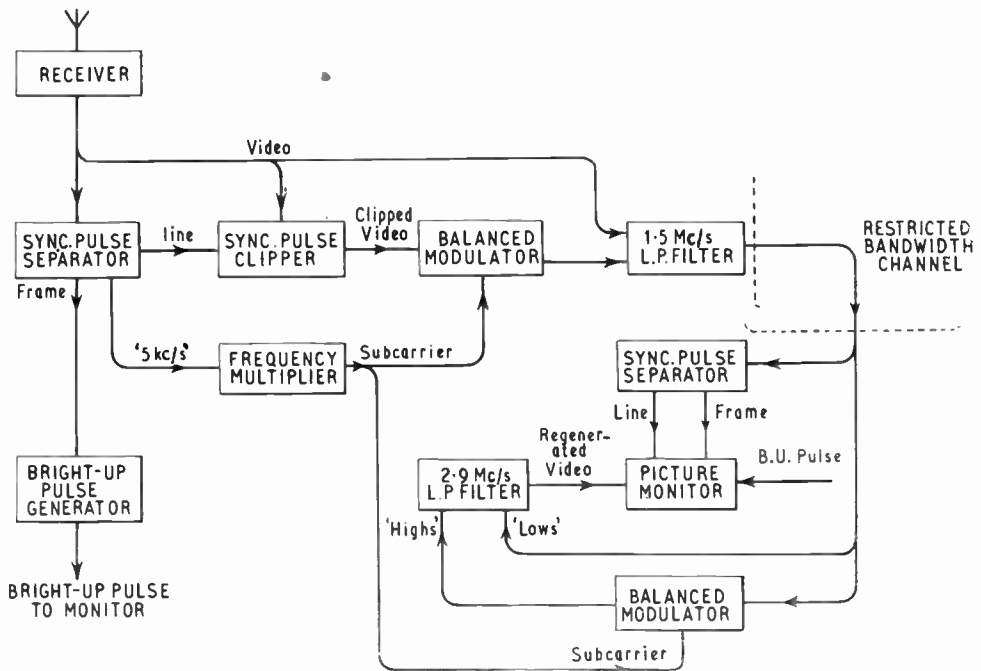
Fig. 5. Block diagram of experimental apparatus.

used by the B.B.C. is indirectly locked on to the 50 c/s mains via a divider chain, and so may vary slightly. Accordingly, the sub-carrier is generated by means of a frequency multiplier system from the received line sync.-pulses. A nominal 5062·5 c/s multivibrator is synchronized from the line sync.-waveform and this half-line-frequency signal is then applied to the multiplier, which consists of four simple class C stages giving an overall multiplication of 585. This is made up of $\times 13 \times 5 \times 3 \times 3$, individual stages of necessity giving an odd multiplier.

The output frequency is then nominally

The video signal undergoes a phase reversal in passing through the clipper unit, but this is unimportant as a correctly-phased output from the modulator can be obtained by reversing the phase of the subcarrier as it is a suppressed-carrier system.

A pair of pentodes, supplied with push-pull video on their control grids, and push-pull sub-carriers on their suppressors, form the basis of the balanced modulator. A double-sideband suppressed-carrier output is taken from their paralleled anodes. By suitable adjustment of control- and suppressor-grid bias the carrier

and video "leaks" in the output may be reduced to a sufficiently low level; normally the video leak is approximately – 30db relative to the wanted output sideband. It was pointed out in Section 3.1 that the sub-carrier could be represented within the frequency band, given suitable equipment in the receiver for re-synthesizing it from bursts of oscillation transmitted at a lower frequency. Since the primary object of the work is to evaluate the characteristics of the system, the sub-carrier in the experimental arrangement is supplied independently of the video signal from the transmitter to the receiver. A simple phase-shift network is included in the sub-carrier supply so that the relative phases of sub-carrier at transmitter and receiver may be adjusted.

The output of the modulator is combined with the main video signal in the input circuits of the transmission filter, which consists of a balanced lattice fed by cathode followers in push-pull. The frequency response of the filter is arranged to be – 6db relative to the response at low frequencies, at 1·48 Mc/s, i.e. half sub-carrier frequency. The phase characteristic of the filter is arranged to be linear up to a frequency approaching cut-off, where some non-linearity occurs.

At the receiving end of the system, the composite signal is applied to a second sync.-pulse separator which feeds the time-base circuits of the picture monitor. As mentioned previously, the sync.-pulses in the composite narrow-band signal are transmitted "uncontaminated" with interleaved signal, so as to avoid erratic synchronizing of the line time base, although their rise-time is lengthened.

A second balanced modulator, of similar design to that at the transmitter is used to recover the high-frequency components of the main signal.

A second linear-phase low-pass filter, of cut-off frequency 2·9 Mc/s, selects the lower sideband of the modulator output, to the exclusion of the upper sideband, and a narrow band, of width approximately 60 kc/s, of the highest frequencies in the lower sideband. This has little effect on the main video signal but reduces the crawling-dot interference very considerably

except after sharp transitions, where ringing occurs for about 2 microsec.

The low-frequency band of the main signal is also passed through the 2·9 Mc/s filter so as to equalise the delays of both the low-and high-frequency bands.

It should be noticed that the general behaviour of the bandwidth-reduction system may be readily simulated. With the sub-carrier at the transmitter arranged as described above, the modulator output is added to the main signal, and both are then applied to a low-pass filter cutting off just below sub-carrier frequency. The resultant signal corresponds to that obtained after the regeneration process in the normal system, and may be used to investigate various interference effects except those due to pre-emphasis. Figure 6 illustrates this arrangement.
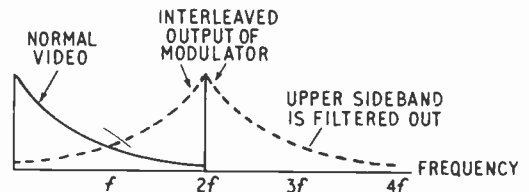

Fig. 6. Method of simulating behaviour of system without pre-emphasis.

The picture monitor consists of part of a commercial television receiver, with a 9-in. studio monitor cathode-ray tube having a metal-backed luminescent screen. In order to record various interference effects, a modified oscillograph camera is arranged to view the cathode-ray tube screen. For photography, the tube is normally blacked out, and is brightened up to a pre-set brilliance by the application of a voltage pulse to its grid from the bright-up pulse generator.

The bright-up pulse is of 1, 2, or 4 frames duration, and its timing is controlled by a three-stage binary counter set to count 2, 3, or 5 frame pulses after the pressing of an "initiate" button.

Any crawling-dot interference will be fully developed in a two-frame photograph, but should cancel more or less completely in four frames, although picture movement will affect this to some extent.
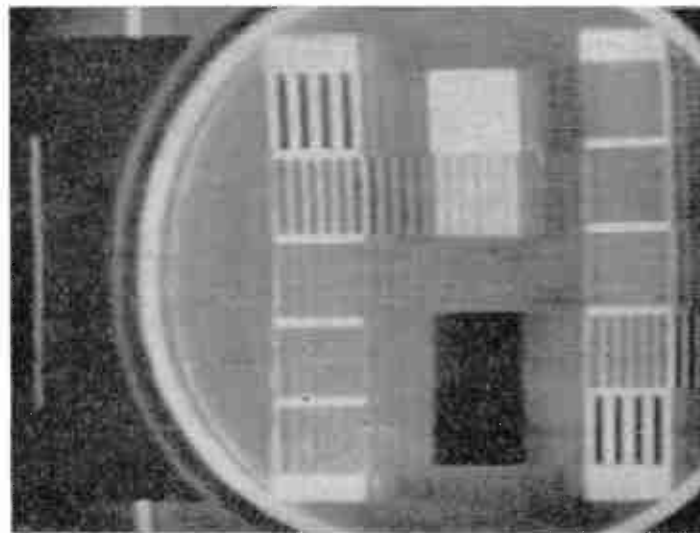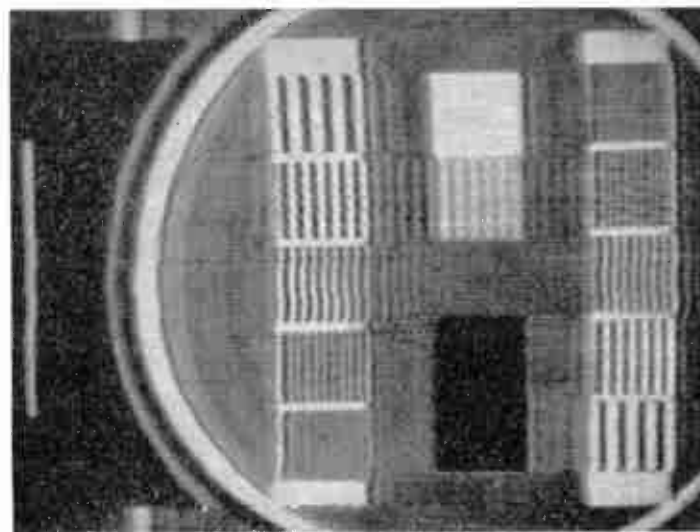
Fig. 7. Original signal, full bandwidth.

Fig. 8. Original signal, 1·5 Mc/s bandwidth.
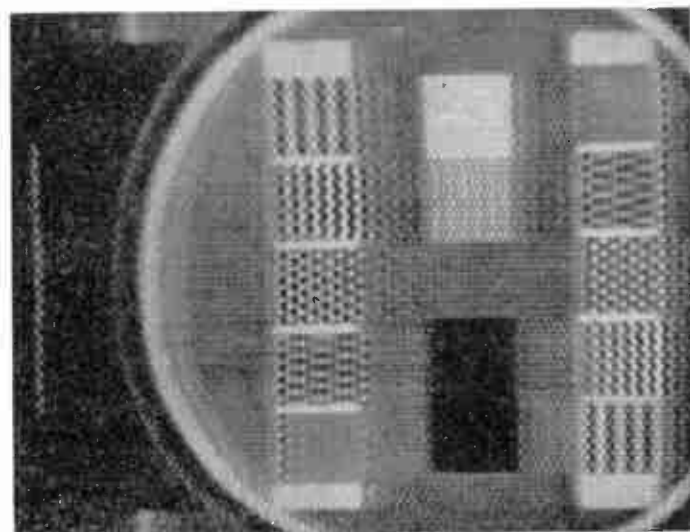

Fig. 9. Regenerated signal, after two frames.

Fig. 10. Regenerated signal, after four frames.

5. Results of Experimental Work

The accompanying photographs illustrate the overall performance of the system with a stationary test-pattern image.

Figure 7 is a photograph of the centre portion of Test Card "C", with no bandwidth reduction in operation, to serve as a basis for comparison. The video response of the system is seen to extend to at least 2·5 Mc/s, although the response falls off rapidly above this frequency. A very faint "ghost" image may also be discerned.

Figure 8 shows the effect of passing the normal signal, corresponding to Fig. 7, through the 1·5 Mc/s filter, but with no "inter-leaved" signal present. There is some ringing after the 1·5 Mc/s bars in the image, which occur at about the cut-off frequency of the filter; the general sharpness of the image is also reduced.

The effect of the complete bandwidth reduction and regeneration process is shown in Figs. 9 and 10. Figure 9 is a photograph of two frames duration and illustrates the type of interference experienced. The ringing of the 2·9 Mc/s filter is noticeable after transitions in the image, and some ghosting is apparent; this latter is caused by imperfections in the various low-pass filters. Considerable interference is evident in the region of the vertical "frequency bars", due to the large-amplitude interleaved signal.

The 2·9 Mc/s filter does not provide sufficient attenuation in its stop range, and hence a general low-amplitude dot pattern may be seen over the entire picture, instead of only at transitions as in the ideal case. However, the transient response of the phase-corrected filter is a considerable improvement over that of a simple "notch" filter, which was found to give a much greater peak amplitude of ringing.

Figure 10 shows a four-frame exposure of the same picture, and should be compared with Fig. 7 which shows the original signal. The high frequency components have been fully recovered; however, a certain amount of disturbance is seen after the 1·5 Mc/s bars. This is due to the lower and upper half-band responses not being exactly complementary; the

rate of cut-off of the 1·5 Mc/s filter is quite rapid, the response falling by 20 db in about 50 kc/s, and difficulty has been experienced in obtaining a uniform overall response.

Complete optical cancellation of interfering frequency components is not possible if the peak excursion of the combined main and inter-leaved signals extends below black level; the vertical white line on the extreme left of Fig. 10 shows considerable blurring due to the average light output of the tube being increased by the presence of the interleaved signal.

The reader may have noticed that we have in the end the characteristics of a dot-interlace system in that four frames must be averaged together in order to eliminate the dot pattern, and there are two ways of looking at the mechanism of this. In a frequency analysis, one says that the half-line-frequency used for the sub-carrier includes an element of 12·5 c/s, so that ideally all the components of the spectrum associated with the sub-carrier will be situated at 12·5 c/s from neighbouring components of the basic video spectrum. This corresponds to modulation of an otherwise stationary picture at a rate of 12·5 c/s, i.e. requiring four frames for a complete cycle. Alternatively, in performing an analysis in the time domain, one notes first that if sub-carrier leaks through, it will produce a visible dot pattern. If the sub-carrier were randomly phased relative to the other picture components, this would be a 3 Mc/s dot pattern; but comparison of Fig. 9 with Fig. 7 shows that the scanning lines are paired producing a 202½-line structure, as far as dots are concerned. The spacing of dots parallel to the line direction, i.e. the length occupied by a light dot plus a dark dot is approximately equal to the two-line width since it contributes two picture elements. In a balanced-modulator with suppressed carrier and no d.c. transmission, the transition from black to white will be represented by a reversal of phase of carrier as sketched in Fig. 11. The carrier is suppressed in the sense that if one averages over a long time the Fourier component at carrier frequency vanishes. But in the television application one does not average the waveform directly, because parts of it which occur consecutively in time are exhibited at

different points in space. It follows that if one has a space of constant intensity wide enough to contain many cycles of sub-carrier frequency, one looks at the right-hand half only of Fig. 11 and one says that there is sub-carrier present. This can be eliminated by including a "notch filter", but only at the cost of introducing a
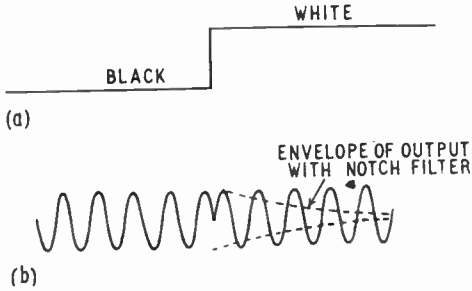

Fig. 11. Illustrating operation of balanced modulation.
(a) Input to modulation.
(b) Suppressed-carrier output.

disturbance or "ring" at the transition point, of which the periodicity is inversely proportional to notch sharpness (proportional to notch width). Thus a very narrow notch produces a low-frequency ring which extends over a considerable length in the picture.

It should be pointed out that the photographs represent an abnormally close view of the screen, and at normal viewing distances the crawling interference is relatively inconspicuous, except in large areas of fine detail, such as scenes involving patterned fabrics.

6. Conclusions

A simple method has been described of obtaining a 2 : 1 reduction in bandwidth of a television signal by a frequency-interlace technique.

The economy in bandwidth is achieved at the expense of a certain amount of "crawling-dot" interference, which is rendered relatively unobtrusive, at normal viewing distances, by

filtering. However, some low-frequency interference is caused by large areas of fine detail, such as might occur in pictures involving patterned fabrics. Normal filtering techniques cannot be employed to reduce this type of interference.

It may be worth-while noting that on filtering a typical normal picture at $1 \cdot 5$ Mc/s, as distinct from a special test card, a quite acceptable picture may be obtained; this effect has also been pointed out elsewhere.[7] The question whether one prefers the $1 \cdot 5$ Mc/s picture with its reduced definition or the frequency-interlaced picture with its dot structure is a matter for subjective judgment.

7. Acknowledgments

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Reflection Coefficient Curves of Compensated Discontinuities on Coaxial Lines and the Determination of the Optimum Dimensions†

by

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Summary: Discontinuities on coaxial lines are caused either by irregular cross-section or variation o. dielectric constant. Equivalent circuits of different types of discontinuity are given. The node-shift technique employing an adjustable short circuit for determining the reflection coefficient is described. Test results are discussed at length and a series of curves for various configurations given.

1. Introduction

Coaxial lines cannot always be homogeneous: in instruments, at the ends of cables, and in connectors, discontinuities of some degree are unavoidable. Dielectric supports for inner conductors, right-angle transitions and tapers are examples of discontinuities of defined configuration very commonly encountered in practice. In cables, dielectric supports are often used to support the inner conductor, such a cable representing in itself a cascade of a great number of identical inhomogeneous sections. Discontinuities in transmission lines cause reflections. In the configurations mentioned above, these reflections are noticeable at frequencies above 20 Mc/s. In systems with large line cross-sections, e.g. transmitter systems, compensation is necessary for frequencies of 100 Mc/s and higher. With microwaves, where the physical dimensions of a discontinuity may be significant fractions of the wavelength used, it is difficult to compensate the line components so that reflection is low within a wide frequency band. It is important to be able to determine the geometry of components providing optimum compensation and to know the reflection coefficient as a function of frequency.

2. Conventional Equivalent Circuits for Discontinuities

Discontinuities in coaxial lines can arise from irregular cross-sections (see Fig. 1) or variation

of dielectric constant. Compared with the homogeneous line, the discontinuity has too high or low a capacitance or inductance per unit length, and may be represented by a lumped capacitance or inductance in the equivalent circuit. The discontinuity may be defined as that
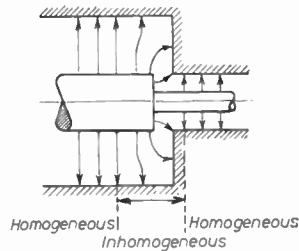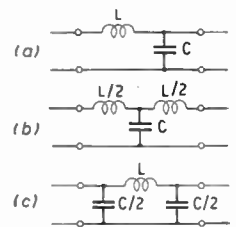

Fig. 1.
Field configuration at change in section of coaxial lines.

Homogeneous | Homogeneous
Inhomogeneous

Fig. 2.
Equivalent circuits for discontinuities.

part of the line in which irregular distribution of the electro-magnetic field occurs. Dividing planes may be drawn through the points where irregular distribution begins and ends and the line element between may be shown as a four-terminal network of lumped capacitances or inductances in L, Π or T filter sections. (Fig. 2.) As long as the dimensions of the discontinuities are smaller than 0·1 wavelength the approximation by a simple *LC* combination is permissible,

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C representing the capacitance and L the inductance of the irregular line section [1, 2]. The section could more accurately be represented by sub-division of the reactances and re-combination in a number of different four-terminal networks in cascade. There is, however, great difficulty in determining the values for the equivalent circuits and in calculating the frequency response, so that only the simple equivalent circuit is commonly used. However, in measurement, diverging curves show repeatedly that the approximation by the simple LC section is inadequate. Certain discontinuities, e.g. those due to a change in cross-section, behave as pronounced capacitances at low frequencies and hence appear in the equivalent circuit as a shunt capacitance. The magnitude and frequency response of this capacitance at high frequencies have been calculated by Whinnery and others[3].

3. Possibility of Compensation for the Reflection Coefficient due to a Discontinuity

Experience has shown that the reflection coefficient of a discontinuity can be much reduced by changing the configuration of the discontinuity itself or of its immediate vicinity. When, for example, the line capacitance is too high at the discontinuity, then the inductance may be increased to match the characteristic impedance of the equivalent four-terminal network to that of the homogeneous line. This increase in inductance can be achieved by increasing the diameter of the outer conductor or by decreasing that of the inner conductor. Reverse measures may be used to obtain a desired increase in capacitance. The discontinuity is not eliminated but compensated. To compensate a discontinuity due to a change in cross-section the L/C ratio is varied by displacement of the transitions of the inner and outer conductors relative to each other. As long as the dimensions of the discontinuity are small compared to the wavelength the approximation by the simple LC section is quite good. The same holds for the conformity of measured values with calculated values, e.g. by Griemsmann[4] for a sudden change in characteristic impedance, by Meinke[1] and Kaden[5] for the dielectric support, and by Meinke[2] for a change in cross-section. Considerable differences between measured and calculated values are obtained at relatively high frequencies, and are explained by the fact that the lumped constants no longer hold good and the distributed reactances in the irregular line section result in a complex frequency response. Until now, there has been no publication of general and systematic measurements of the behaviour of compensated discontinuities of the basic forms, providing a basis for calculated results, especially at very high frequencies. The need has been made urgent by recent technical developments imposing stringent requirements of freedom from reflection in components. Experience has shown that, for example, in the range about 4000 Mc/s, where coaxial lines are still used, specifications call for very low reflection coefficients while the dimensions of typical discontinuities far exceed 0.1 wavelength. The same applies to television systems at lower frequencies between 200 and 500 Mc/s because the high powers employed require large cross-sections of line.

4. Advantages in Determination of Optimum Compensation by Measured Curves

A practical approach to the problem of compensation is made by plotting families of curves. This method has the advantage over calculation in that it is not based on approximation but upon the components as they appear in practice. Optimum compensation (as considered in this paper) is obtained when a reflection coefficient has a minimum value over a maximum frequency range.

A series of measurements determines the reflection coefficient as a function of frequency. The curves drawn from these measurements are then used to assess the quality and suitability of the components for different purposes.

One example is in the choice of rectangular-step or tapered transitions of diameters in coaxial lines. From the mechanical point of view a right-angle transition is always simpler than a taper. The relative merits have been debated with some opinion in favour of the taper. Measurements illustrated by curves discussed later show that both forms yield the same result when properly dimensioned: the dimensions usually advocated for the taper are not optimum, being based on approximation.

When dielectric supports are used, grooves in the outer conductor should be avoided for mechanical reasons. The difference between supports compensated at the inner conductor and those compensated at the inner and outer conductors is also made clear by the curves. Other important problems are clarified by discussion of the measurements.

5. Accuracy Requirements

A high standard of absolute accuracy is necessary in plotting curves which may be used as a basis of design. Present-day components are required to have reflection coefficients of less than 0·5 per cent., which is considered practicable; the accuracy of the families of curves shown here must therefore be better. The difficulty in these fundamental measurements lies not only in the manufacture of suitable means, which must have a considerably better accuracy than those commonly used, but also in the identification and elimination of new measuring errors, which may occur with refined measuring methods. This calls for considerable experience and scrupulous care in the test set-up to ensure satisfactory results.

6. Test Methods and Test Set-up

The most accurate method of determining small reflection coefficients of four-terminal networks is the node-shift method described by Weissfloch[7]. With modern means an accuracy of 0·15 to 0·2 per cent. is obtainable. The quantities necessary for calculation of the reflection are obtained by physical measurement of length and of the ratio of diameters of coaxial lines and can therefore be determined with good accuracy. A brief description of the node-shift method is as follows: The discontinuity, represented by a four-terminal network with defined contact planes E_1, E_2, is inserted between a short-circuited coaxial line of variable length and a slotted line providing indication of standing waves (Fig. 3). The short-circuited length of line is varied in small lengths (l_1) until a complete half-wave length is traversed. Simultaneously the shift of the voltage node is observed on the slotted line, in corresponding increments of length (l_2). The lengths $l_1 + l_2$ are plotted as a function of l_1 to obtain the node-shift curve, from which is derived the error Δl,
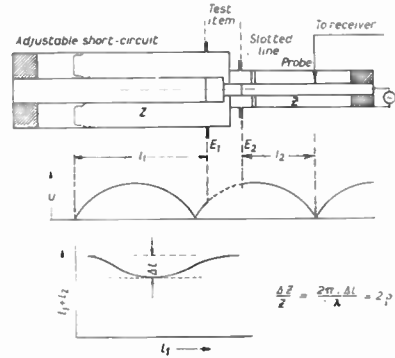

Fig. 3. Measurement of the reflection coefficient ρ by the node-shift technique with a slotted line and a movable short circuit.

due to the discontinuity. If the characteristic impedances of the slotted line and of the adjustable short-circuit are equal (= Z), then a simple relation is obtained for small reflections of the four-terminal network under test:

$$\frac{\Delta Z}{Z} = \frac{2\pi \Delta l}{\lambda} = 2|\rho|$$

$|\rho|$ = reflection coefficient

The advantage of this method is that the variation in characteristic impedance is determined from the measured length Δl. In general, length can be measured with an accuracy of 0·1 mm which yields good accuracy. To check the test set-up itself, the measurement may be repeated with an adjustable short-circuit connected directly to the slotted line. The inherent error of the test set-up is given by the remaining Δl and therefore easy to determine. The adjustable short-circuit can be made without a dielectric support and is then a homogeneous line which can be regarded as a reactance standard of calculable characteristic impedance. It serves to check the slotted line and to provide for correction where the slot and/or the unavoidable end-supports cause appreciable error. The complete test set-up requires, in addition to the slotted line and the adjustable short-circuit, a signal generator which is unaffected by load-reaction, and a sensitive indicating receiver.

6.1. The Adjustable Short-Circuits

To be a characteristic-impedance standard, the adjustable short-circuit itself must not show any discontinuity, i.e. it must be straight, of uniform cross-section and without supports for

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the inner conductor. The difficulty in fulfilling this requirement lies in that the inner conductor will sag since it is too long to support itself. In this condition the adjustable short-circuit would not be usable. A satisfactory solution is obtained by using the inner conductor of the item under test to support that of the adjustable short-circuit. To this end, a centring coupling for the inner conductors must be provided, which operates when the adjustable short-circuit is connected to the item under test. The construction of such an adjustable short-circuit has been described in an earlier article[9].

Adjustable short-circuits of five different cross sections were made, so that enough points could be obtained by measurement for the curves of transitions. The diameters of the outer conductors were 21 mm, 38 mm, 58 mm, 78 mm and 115 mm. For the characteristic impedances of 50, 60 and 75 ohms, the same outer conductors and driving mechanisms were used with different inner conductors and their associated plungers. The useful length of each adjustable short-circuit was 500 mm, catering for measurements down to 300 Mc/s. The diameters of the inner and outer conductors were measured at three points, giving an average deviation in characteristic impedance $\Delta Z/Z$ of 0·2 per cent. for three instruments and a value still better for the other items.

6.2. Slotted Lines

For the slotted lines two diameters were found especially advantageous namely, 21 mm and 58 mm. For measurement on transitions the 21-mm line was used. This gives, in combination with the adjustable short-circuit described above, the diameter ratios D/D_1: 21 : 38, 21 : 58, 21 : 78 and 21 : 115. A further slotted line with 58 mm diameter was provided for measurement of dielectric supports.

6.2.1. Choice of cross-section and test frequencies

The choice of the two cross-sections of 21 mm and 58 mm was not arbitrary, but made after consideration of dimensions and the reflection coefficient of components as a function of frequency. As Konig has shown[8], in cavity resonators the ratio between resonant wavelength λ and the volume G of the cavity is linear: $\lambda \propto G$. A linear increase or decrease by the factor k

of the volume of a resonator whose resonant frequency is λ_0 gives the new resonant frequency $\lambda_k = k \lambda_0$.

In waveguides the cut-off wavelength λ_g is proportional to the width b for rectangular cross-section or the diameter d for circular cross-section. By analogy, the same relations must hold for discontinuities. The reflection coefficient $|\rho|$ is hence a function of the geometrical dimensions $G_{x, y, z}$ of the discontinuity and of the product of the frequency f by a constant k:

$$|\rho| = F(f.k, G_{x, y, z})$$

Discontinuities which differ only by a linear variation by the factor k of the three dimensions x, y, z have the same reflection-coefficient curve with, however, a frequency scale varying linearly by the factor k. Hence the reflection coefficient of discontinuities of similar geometry can be determined by one measurement made on a model. By choosing the dimensions of the model it is possible to shift the frequency so that the reflection coefficients may be more accurately determined.

The proportionality of the reflection coefficient to the geometry of discontinuities in coaxial lines follows from the fact that the zone of the disturbed field in the direction of the longitudinal axis is proportional to the diameter, while distributed reactance C', L' and the characteristic impedance remain unchanged. The greater the dimensions of the distorted field, the greater is the absolute value of the resulting reflection coefficient. In transmission lines of small cross-section the dimensions of the distorted field are small, so that appreciable reflection occurs only at relatively high frequencies. Experiments have shown that the measuring range of reflection coefficients for supports for the 21-mm section falls in the frequency range about 1500 to 3000 Mc/s where the accuracy obtainable is limited by mechanical tolerances. For this reason the diameter of 58 mm, which is roughly three times as much, was chosen for the measurement of supports, so that the reflection curves fall in the frequency range of 500 to 1500 Mc/s. For the same reason the adjustable short-circuits of larger diameter than 21 mm were made so that the reflection coefficients of right-angle transitions could also be brought into a lower frequency

range, i.e. 300 to 2000 Mc/s. A shift of the measuring range towards lower frequencies has advantages in the receiver and signal generator: triode signal generators usable at the lower frequencies are simpler to operate than the klystron generators necessary for higher frequencies. In addition, contact instability of detachable connections increases with frequency. Flexible cables often show defects in density and characteristic impedance at frequencies above 1500 Mc/s. Cable connections are unavoidable between the signal generator and slotted line, and especially between the movable probe of the slotted line and the receiver. Briefly, the present state of the art provides for satisfactory measurement of very low reflection coefficients only within the frequency range of about 300 to 2000 Mc/s.

6.2.2. Construction of the slotted lines

The construction of the 21-mm slotted line used in the measurements was described in an earlier article[9]; an aperiodic probe is used however instead of the tuned probe and is connected directly to the measuring receiver.

The mechanical connection of the adjustable short-circuit, slotted line and transition under test is shown in Fig. 4. The adjustable short-circuit is extended at its open end by the transition. The outer conductors are connected by means of detachable flange couplings. The contact of the inner conductors at the separating
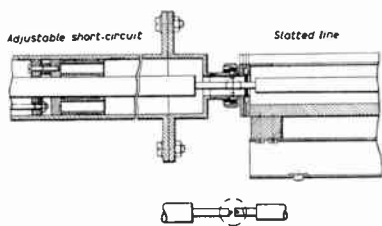

Fig. 4. Sectional view of the test setup for a change in section.

plane of the slotted line is ensured by axial pressure of the inner conductors against each other. The whole system is supported and centred by the support of the slotted line. Thus, the adjustable short-circuit proper can be perfectly homogeneous and the transition up to the

contact plane of the slotted line can be free from additional discontinuities, so that only the reflection occurring in the transition enters into the measurement.

The slotted line employed for measurements of supports had a uniform diameter of 58 mm. Its inner conductor was rigid enough to dispense with support since a slight eccentricity has a negligible effect on the characteristic impedance.
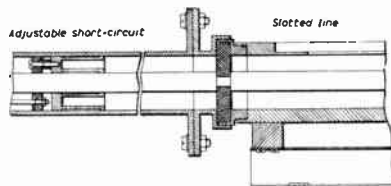

Fig. 5. Sectional view of the test setup for dielectric supports.

This was confirmed by a check of the overall accuracy. The probe was of the same design as that of the 21-mm line. The mechanical connection between slotted line, item under test, and adjustable short-circuit is shown in Fig. 5. Threaded sleeves are provided for the connection of the outer conductors. The contact of the inner conductors is made by pressure of the spring-loaded inner conductor of the adjustable short-circuit against the inner conductor of the slotted line holding the support. The inner conductors of the adjustable short-circuit and of the slotted line are thus centred and supported by the support under investigation.

6.3. Signal Generators and Test Receivers

Three signal generators were used (Type SDR, 300 to 1000 Mc/s, Type SCR, 950 to 1840 Mc/s, Type SBR, 1700 to 2700 Mc/s, together with the UHF Test Receiver Type USVD, 275 to 4600 Mc/s.) The frequency stability of these signal generators is better than 5 parts in 10^5 over 15 minutes, ten times better than the value necessary for the minimum measurable reflection coefficient of 0·1 per cent. Reaction of the load—which varied continually during measurement—on the test frequency was negligible, even with the relatively high output voltage of 0·5V. The ease of operation and stability of these instruments were invaluable in carrying out the large scale measuring programme.

6.4. *Check of Accuracy*

As mentioned above, the relative accuracy of the test set-up can be found by a node-shift measurement with an adjustable short-circuit of equal cross-section connected directly to the slotted line. The length Δl obtained is a measure of the uniformity of the slotted line and adjustable short-circuit. The following maximum values of Δl were obtained:

Adjustable short-circuit and slotted line of 21 mm diameter:

f (Mc/s)	500	1000	2000
Δl (mm)	0·2	0·1	0·1

$$\frac{\Delta Z}{Z} = \frac{2\pi\,\Delta l}{\lambda} \qquad 0·21\% \qquad 0·21\% \qquad 0·42\%$$

Adjustable short-circuit and slotted line of 58 mm diameter:

f (Mc/s)	300	500	1000	1500
Δl (mm)	0·15	0·1	0·05	0·05

$$\frac{\Delta Z}{Z} = \frac{2\pi\,\Delta l}{\lambda} \quad 0·094\% \ 0·105\% \ 0·105\% \ 0·157\%$$

The Δl values quoted are maximum values from a great number of measurements which were repeated frequently. The average value is lower than the values quoted. The tables show that relative accuracy is better for the 58-mm section than for the 21-mm section. This is due to the fact that, for obvious reasons, the accuracy in diameter is better for the larger cross-section. The remaining reflection of the support and the transition in the 21-mm slotted line contribute to the measuring error, while the 58-mm slotted line, having uniform cross-section and no support. does not show these errors. For the adjustable short-circuits the deviation of the characteristic impedance from the nominal value was less than 0·2 per cent.

7. General Discussion of Test Results

The measured reflection coefficient curves for compensated discontinuities differ greatly in each case. a few examples being shown here in Figs. 7. 10. 11, 12, 14 and 17. These differences are not due to measuring errors, the relative accuracy of the test set-up having been determined between 0·1 to 0·2 per cent., while the discrepancies of the curves are much greater. The irregularities found in the reflection coefficient curves are hence in conformity with physical laws and characteristic of compensated discontinuities, as shown by the great number of measurements. It is remarkable that slight variation in the configuration of a discontinuity can cause considerable variation of the reflection coefficient curve. Three characteristic shapes of curve recur frequently and are shown in Fig. 6. In curve 1 the reflection coefficient increases continuously with frequency; the curve is similar to a parabola, but part may also be more or less straight. Curve 2 is slightly S-shaped, while curve 3 after an initial rise falls
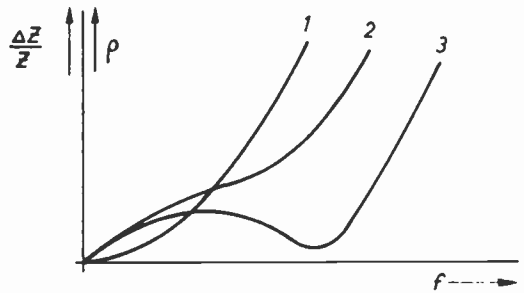

Fig. 6. Characteristic reflection-coefficient curves of compensated discontinuities.

again, turns round and finally rises steeply. All other reflection curves of compensated discontinuities are combinations of these three basic forms, approaching one or the other type. When the compensation is made in the usual manner, e.g. by changing the displacement a in a diameter transition or by changing the diameter of the inner conductor for a support, then the reflection curves grow flatter with decreasing reflection until an optimum is reached; then reflection increases again. The optimum is clearly distinguishable; apart from rare exceptions a curve of shape 1 is obtained. However, before and after the optimum, curve types 2 or 3 may appear without obvious reason. Hence, in the vicinity of the optimum, the reflection coefficient curve is determined by the geometrical configuration of the discontinuity. To show the character of the discontinuity three curves in the immediate vicinity of the optimum are shown in most of the figures.

At high frequencies compensation according to curve 3 has a considerably lower reflection coefficient than that of curve 1, i.e. its bandwidth is much wider. In this sense the compensation of type 3 may be regarded as optimum.
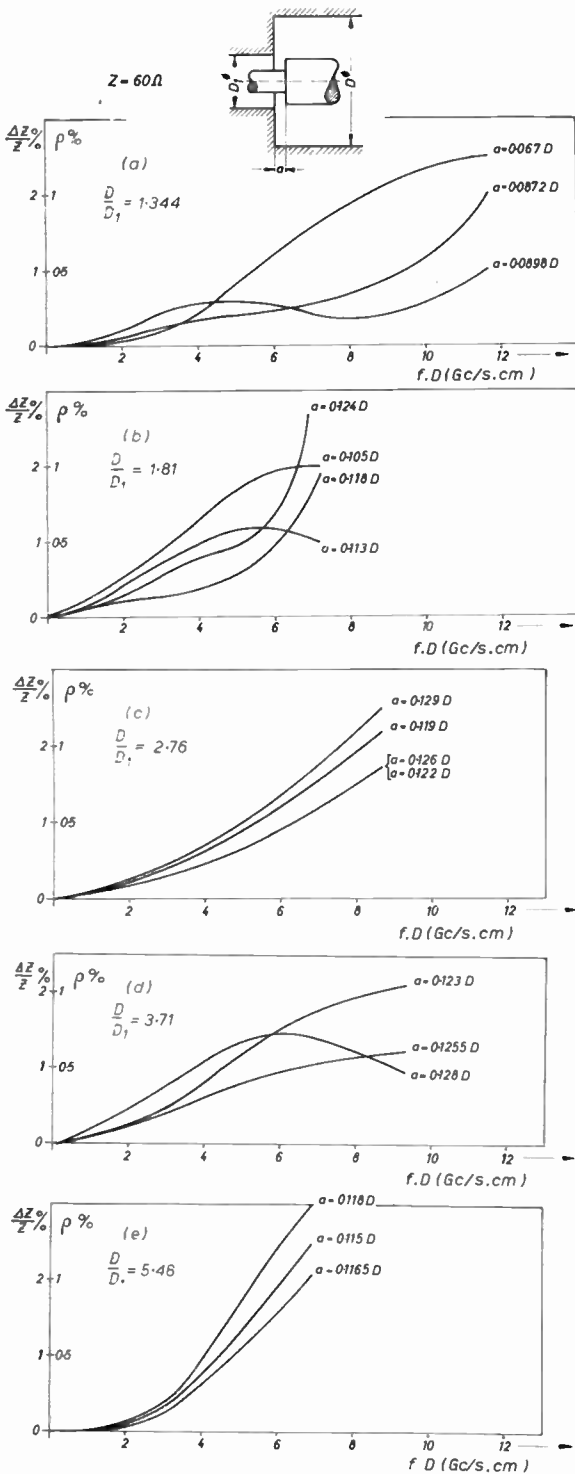
Fig. 7. Reflection-coefficient curves of the right-angle transition for various diameter ratios D/D_1.

In this consideration, this definition is not accepted since measurements show that the curve 3 is not obtained with all discontinuities and that at lower frequencies it has a higher reflection coefficient than curve 1. At lower frequencies compensation according to curve 1 always gives the minimum reflection coefficient, and in the author's experience, can nearly always be realized. Thus it can be chosen generally to represent optimum compensation. The curves of optimum dimensions determined by the measurements always refer to compensation according to curve 1.

Compensation according to curve 3 is of high technical interest because of its broader useful frequency band (broadband compensation). It will therefore often be the aim to be able to control the compensation of any discontinuity in such a way that the curve of reflection coefficient versus frequency takes a predetermined shape. The frequent recurrence of certain shapes of curves suggests that it may be possible to control the process, for instance by introduction of a second variable. This may lead to modified or new configurations of the components commonly used, having reflection coefficients still lower than those of the components dealt with in the following, in which adjustment was made with one variable only.

7.1. Normalized Scales in Curves of Measurements

The programme of measurements of transitions and dielectric supports was carried out choosing in each case the line diameter affording the highest measurement accuracy, i.e. 58 mm for supports and 38 to 115 mm for transitions. To provide a common basis for comparison of the curves of different discontinuities they must be referred to a common line diameter. It is convenient to choose a reference diameter of $D = 1$ cm. All the curves shown have been normalized and have for abscissae the product $f \times D$ in Gc/s and centimetres, other dimensions given being referred to the line diameter $D = 1$ cm. The reflection coefficient of similar discontinuities, e.g. transitions of equal diameter ratio D'/D_1' or supports of equal ratio B'/D' of breadth B' to diameter D', can be read immediately by dividing the scale of the abscissae by the model factor D' (cm).

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7.2. *The Right-angle Transition*

Reflection coefficient curves for transitions of various diameter ratios D/D_1 are shown in Fig. 7 as a first example. The parameter of the curves is the displacement a between the planes of cross-section change of the inner and outer conductors. The abscissa scale $f \times D$ is standardized for the larger diameter; the figures inscribed refer to the diameter of 1 cm.

As already discussed three characteristic curve shapes, 1, 2, and 3, are found in the reflection curves. The tendency towards broadband compensation (curve shape 3) is predominant with small and medium diameter ratios, while with larger ratios, e.g $D/D_1 = 5.48$, the curve shape 1 is always obtained. The explanation is that a transition presents two discontinuities separated in space, one situated in each corner of the step between the smaller and larger diameters. With a relatively low diameter ratio the discontinuities are near to each other, the effects adding together to increase the reflection coefficient although the absolute dimensions of the discontinuity are still small. With increasing frequency it may be that one discontinuity is still capacitive while the other one is inductive so that their effects compensate to some degree. This is the case of broadband compensation according to curve 3.

With the ratio D/D_1 increasing, the line diameter D_1 becomes smaller. The disturbance at the small diameter decreases proportionally down to values which are smaller in comparison with that of the large diameter. This is the case, for instance, of the transition with $D/D_1 = 5.48$ where only the lumped disturbance at the
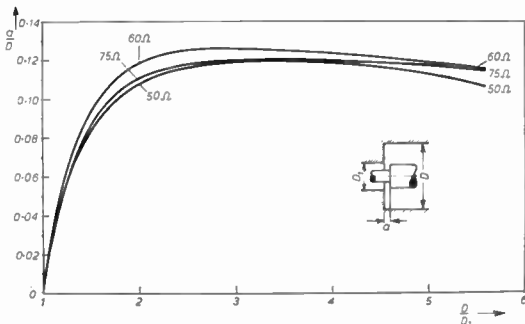

Fig. 8. Optimum dimensions of the right-angle transition, a/D being a function of the diameter ratio D/D_1.
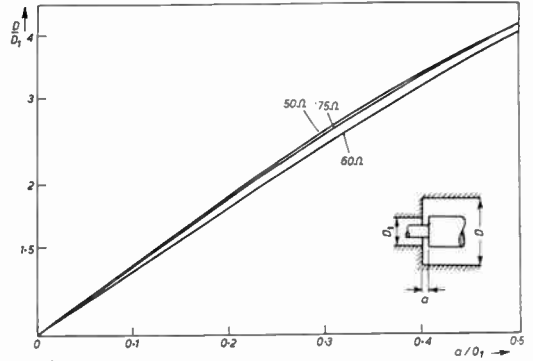

Fig. 9. Optimum dimensions of the right-angle transition as shown in Fig. 8. but with the displacement a related to the diameter D_1.

large diameter determines the reflection curve and results in curve shape 1. It is notable that for transitions the reflection coefficient is practically constant for all values of D/D_1 between 1.5 and 5.5 (see Fig. 8). With ratios D/D_1 lower than 1.5 a noticeable decrease is found. This phenomenon is due to the addition of disturbances, which has its full effect with small diameters, while with large diameters it contributes only slightly to the overall reflection coefficient. About $D/D_1 = 1.8$ a maximum of reflection is found. Transitions with lower and higher ratio are slightly better but the differences are not great.

Figure 8 shows the curves for optimum dimensions of the right-angle transition of 50, 60 and 75 ohms characteristic impedance, a/D being a function of the diameter ratio D/D_1. The same curves are shown in Fig. 9 with D/D_1 plotted to a logarithmic scale and a/D_1 to a linear scale. For ratios D/D_1 lower than three these curves are practically straight. The following simple formula can be deduced therefrom for the axial displacement a:

$$a = \frac{D_1}{k} \log_e \left(\frac{D}{D_1} \right); \qquad \frac{D}{D_1} < 3$$

k is a constant dependent on the characteristic impedance

$$k_{50\,\Omega} = 3.09$$
$$k_{60\,\Omega} = 2.90$$
$$k_{75\,\Omega} = 3.04$$

a is a maximum for the characteristic impedance of 60 ohms, decreasing at 50 and 75 ohms. This

is due to the field intensity at the inner conductor being at minimum with a characteristic impedance of about 60 ohms. For a certain change in capacitance—to compensate the discontinuity in the corners of the transition—a greater displacement of the conductors is necessary in low field intensity than in high field intensity.

Inversely, the optimum correction expressed by the length *a* decreases with increasing field strength. Therefore the 60-ohms transition requires the greatest displacement *a*, and is unique in that the reflection coefficient has a minimum and the displacement *a* a maximum value.

7.3. *The Taper*

The reflection curves for tapers of 60-ohms characteristic impedance and slopes of 12 deg and 16 deg at the inner conductor are given in Fig. 10. The scale is the same as for the right-angle transition, so that the curves can easily be compared. The parameter is the displacement *a* between the apices of the tapers of the inner and outer conductors. If the apices coincide, then $a = 0$: axial displacements in both directions are marked by $+$ and $-$.

The curves show that optimum compensation is obtained if the apices coincide, i.e. $a = 0$. Only in certain cases and for particular frequencies can lower reflection be obtained by displacement of the apices. (See Fig. 10.)

It can be assumed that for slopes higher than 16 deg displacement of the apices would in fact give a better result, since the taper approaches and finally passes over into the right-angle transition. Such tapers, however, are of no technical interest, since they may as well be replaced by the right-angle transition. The curves also show that slight displacement of the apices causes a noticeable variation of the reflection coefficient. Therefore the tapers must be aligned with high accuracy to ensure low reflection.

Comparison of the reflection coefficients of the taper with those of equivalent right-angle transitions, i.e. for the same diameter ratio D/D_1, shows that the two types of transition differ very little from each other. Because of the scattering of the plotted curves the average value must be taken for the comparison. Slight
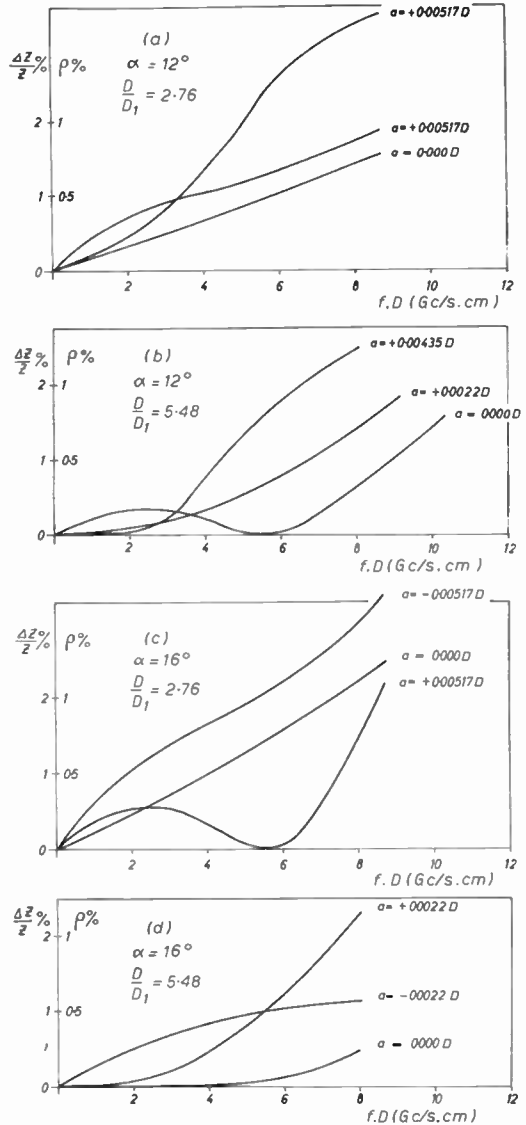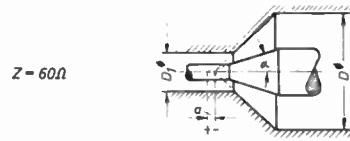

Fig. 10. Reflection-coefficient curves of the taper for various diameter ratios and slopes of the taper.

superiority of the taper can be found only for the high diameter ratio $D/D_1 = 5\cdot48$.

7.4. *The Supports*

Examples of reflection curves for dielectric supports are given in Figs. 11 and 12. The
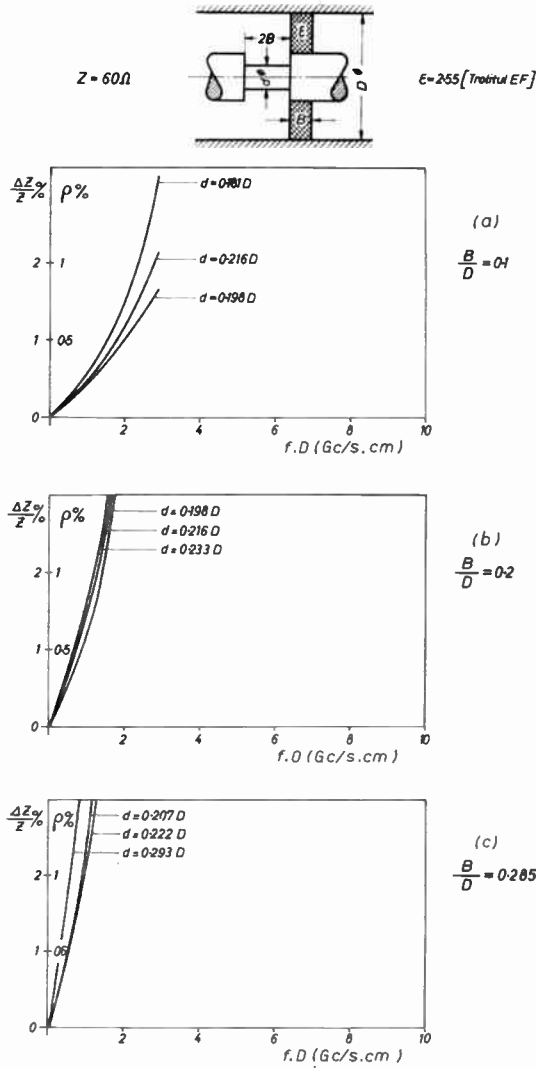
Fig. 11. Reflection-coefficient curves of the dielectric support compensated for at one side of the support.

for both types of support the overall breadth of compensation is $2B$, the support with compensation at both sides[2] has a lower d/D ratio, as shown in Fig. 13 which gives optimum dimensions. This is due to the inhomogeneous field at the re-entrant and projecting edges of the inner conductor, which is enlarged by the division of the groove into two parts.

The support with compensation at one side is inferior in comparison to the supports described in the following. Division of the groove into two parts gives an improvement by
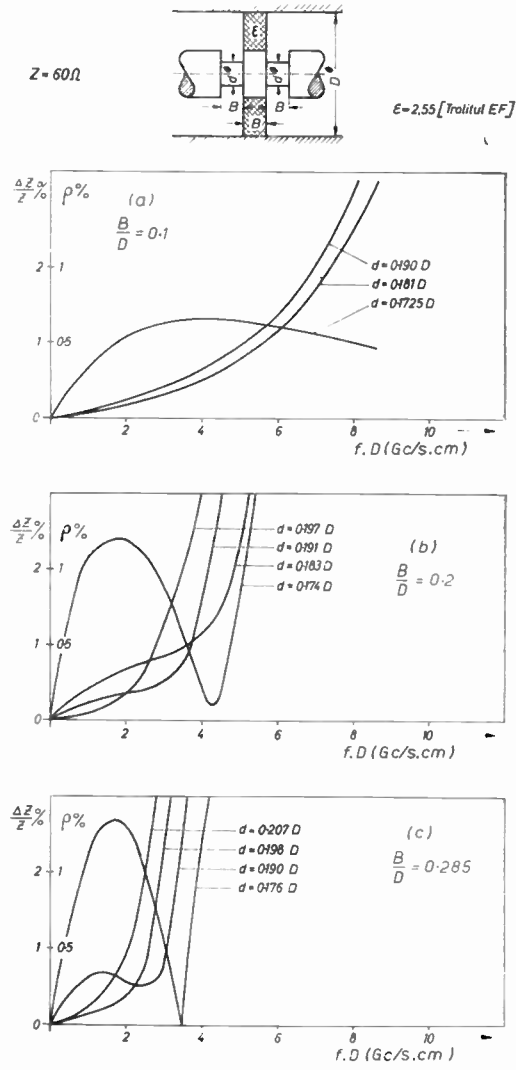
Fig. 12. Reflection-coefficient curves of the dielectric support compensated for at both sides of the support.

abscissa scale is again $f \cdot D$. The breadth of a single compensating groove is $2B$; alternatively, two grooves B, one on each side of the support may be used. The dimensions are designed to secure optimum compensation without unduly weakening the inner conductor, a necessary practical requirement. As is shown by the measurements, the reflection coefficient cannot be affected appreciably by variation of the axial length of the groove without making the core diameter d extremely small, when the core would become so weak that application would be limited for mechanical reasons. Although
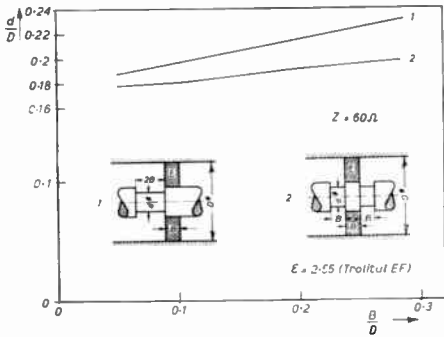
Fig. 13. Optimum dimensions of the dielectric supports compensated for at one side and at both sides of the support.

a factor about 3. This support is still unsuitable for practical application because of its high reflection coefficient. Since these supports are in practice of minor importance, measurements were made only for the characteristic impedance of 60 ohms and dielectric constant $\varepsilon = 2.55$ (trolitul).

Of greatest technical importance is the case where the groove in the inner conductor is immediately under the support and filled by the support itself. Reflection coefficient curves for $\varepsilon = 2.55$, $Z = 60$ ohms and various thicknesses of support are given in Fig. 14. The geometrical extension of the discontinuity is here reduced to about one-third of the preceding case and the number of re-entrant and projecting edges disturbing the field is only four instead of eight.

One might suppose that the reflection coefficient also would decrease by the factor three. Comparison of the curves, however, shows that the decrease is only by 1·3 to 1·5. This indicates the extent to which detail structure, which
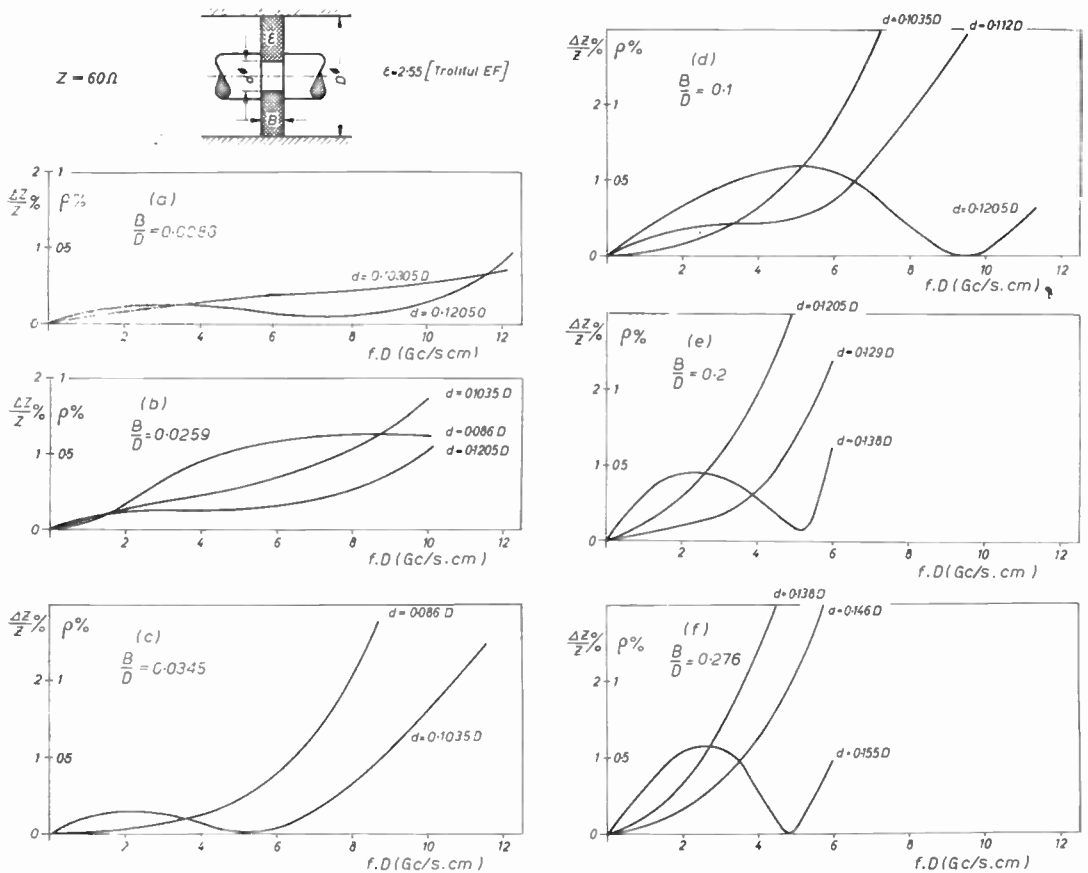

Fig. 14. Reflection-coefficient curves of the dielectric support with grooved inner conductor, for various thicknesses.
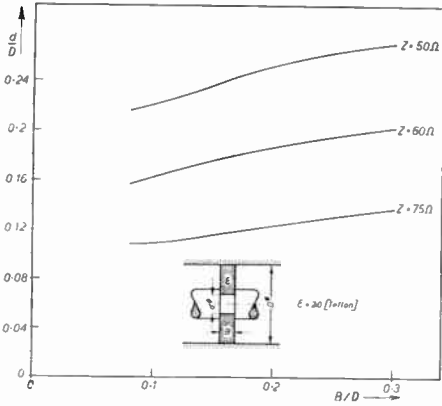
Fig. 15. Optimum dimensions of the dielectric supports for a grooved inner conductor, ε being 2·0.

Fig. 18. Optimum dimensions of the dielectric supports for grooved inner and outer conductors, ε being 2·0.

is not taken into account by the compensation here described, contributes to the final result. Curves for optimum dimensions are given in Figs. 15 and 16. "Detail-structure" is defined as that arrangement and number of reactances in a four-terminal network which would exactly be equivalent to the discontinuity.

An obvious consideration is to make a groove not only in the inner, but also in the outer conductors, as shown in Fig. 17. This has the advantage that the inner conductor need not be weakened so much, since compensation is also achieved at the outer conductor. For the curves plotted the groove in the outer conductor was constant in all cases at $D_1/D = 1·475$ and dimensioned so that for a dielectric constant $ε = 2·55$ and $Z = 60$ ohms, the groove ratios in the inner and outer conductors were roughly equal. The field disturbance, originally com-

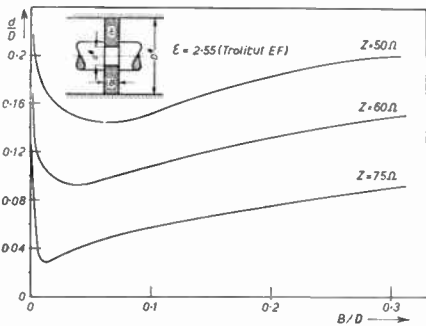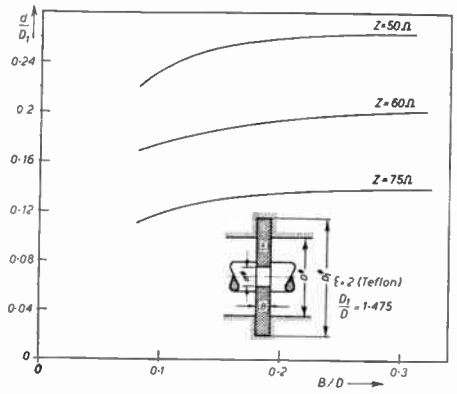pensated at the inner conductor only, is now divided and about equal at both conductors. The grooves are less deep so that the change in field-strength is less pronounced. However, the measured reflection curves for $ε = 2·55$ and 50, 60 and 75 ohms do not show a lower reflection coefficient than that of a support having a groove only in the inner conductor, rather there is some indication that the reflection coefficient of the compensated inner conductor is better. The same holds for $ε = 2·0$. Curves for optimum dimensions are given in Figs. 18 and 19.

Figure 20 shows the four types of compensation in the order of their relative reflection coefficients.

7.5. The Ratio D/d and Z/Z₀ for Supports

Let us now consider the characteristic impedance Z of the supporting disc, which is

Fig. 16. Optimum dimensions of the dielectric supports for a grooved inner conductor, ε being 2·55.
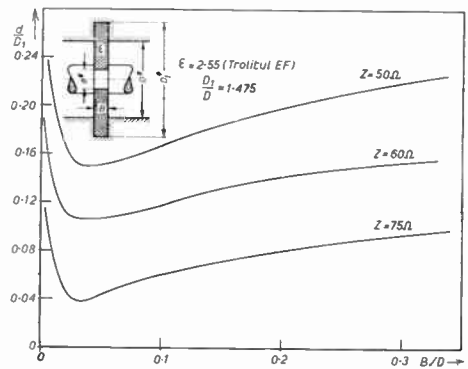

Fig. 19. Optimum dimensions for dielectric supports for grooved inner and outer conductors, ε being 2·55.

determined by the dielectric constant and the diameter ratio D/d or D_1/d, respectively, and relate it to the characteristic impedance Z_0 of the homogeneous line with air as dielectric, where $\varepsilon = 1$. It is known that Z must always be greater than Z_0 in order that the excessive self-inductance can compensate the additional capacitance produced by the inhomogeneous field. For the supports used in practice, where the ratio B/D is between $0 \cdot 1$ and $0 \cdot 3$, the curves for optimum dimensions, shown in Fig. 15, are practically straight and of slight slope for any type of compensation.

$\rho_{relative}$

| 3·9 - 4·5 | 1·3 - 1·5 | 0·8 - 1·0 | 0·8 - 1·0 |

Fig. 20. The four types of compensation in order of relative reflection coefficients.

The quantity d/D increases with increasing B/D. This means that thin supports must enter more deeply into the conductors than thick ones. It is convenient in the simple equivalent circuit to divide the four-terminal network into three series-connected four-terminal networks (Fig. 21). To this end one assumes a four-terminal network of the characteristic impedance Z_0 equal to that of the homogeneous line,
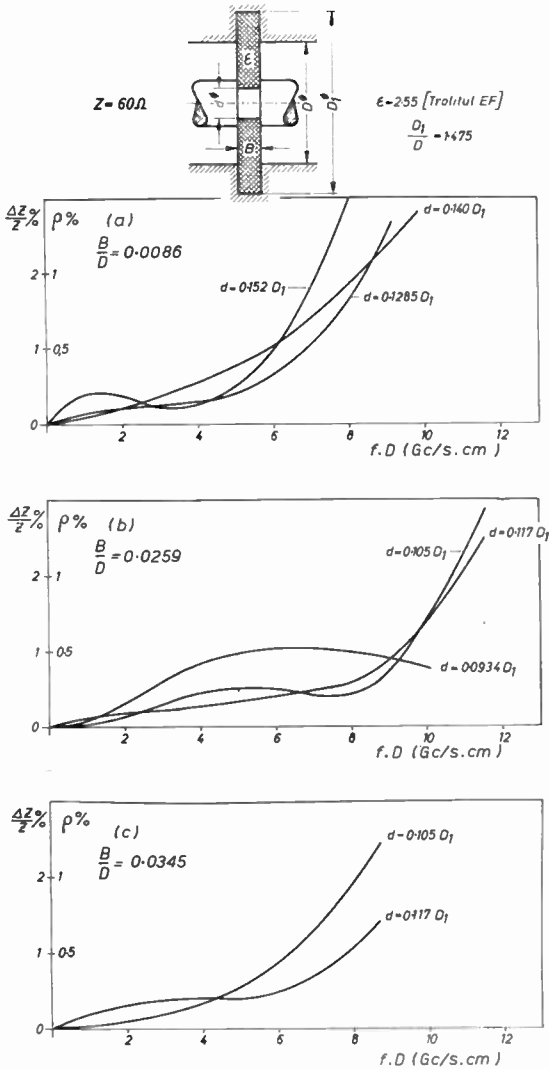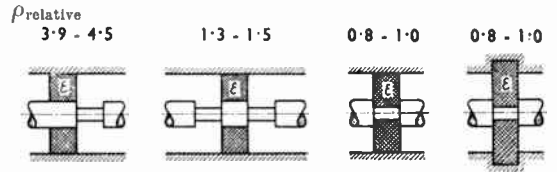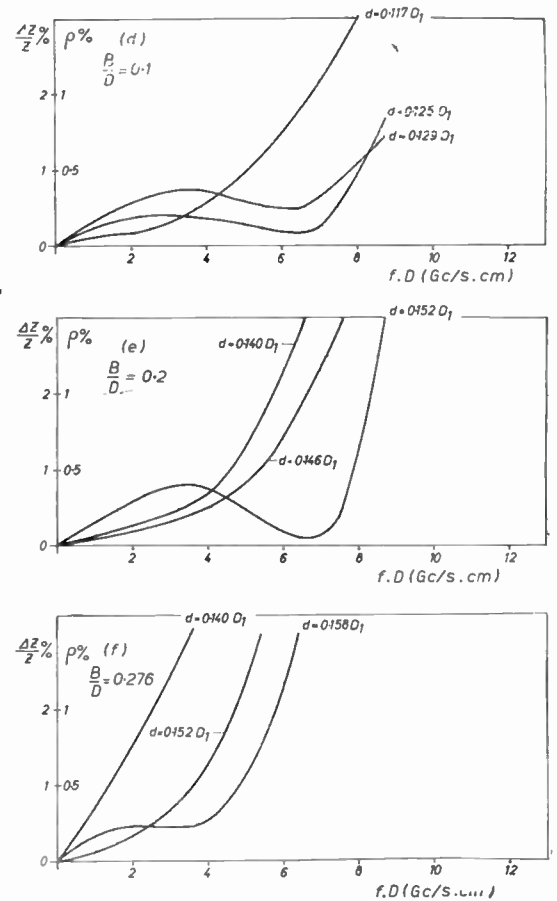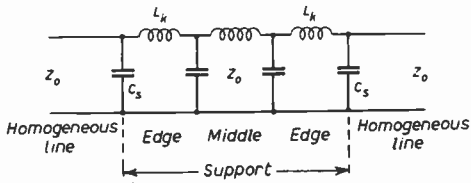

Fig. 17. Reflection-coefficient curves of the dielectric supports for grooved inner and outer conductors, for various thicknesses.

Fig. 21. Equivalent circuit of a compensated dielectric support.

in between the disturbing capacitances C_s, which are compensated by L_k representing the additional self-inductance due to the groove. Since the curves for optimum compensation are practically straight and of slight slope, it can be concluded that the absolute value of the disturbing capacitance C_s itself is practically constant for B/D between 0·1 and 0·3. Hence for thin supports the groove must be deeper to give a high enough L_k, because B, being small, does not give much self-inductance. This equivalent circuit does not take account of the detail structure which is one determining factor of the reflection coefficient curve and can therefore only be used for general consideration.

In general it can be said that a discontinuity for which C_s is great and must be compensated by a large L_k will have a low cut-off frequency. C_s cannot be read directly from the curves measured, but its magnitude is shown through L_k by the known dimensions of the groove. Generally, supports for small grooves must be better than those for large grooves: the dimensions of the groove can be taken as a measure of the quality and compensation of a support. This does not give a rigorous distinction between individual supports with respect to their reflection coefficient, since detail structure also plays a part, as measurements have shown. However, the groove, being accurately defined, is a visible quantity and convenient for comparison of supports of various design. The grooves are less deep if both the inner and outer conductors are grooved (Figs. 18 and 19), but there is only slight difference from the support entering the inner conductor only (Figs. 15 and 16), especially for 60 and 75 ohms and $\varepsilon = 2·55$. This is borne out in the small differences between reflection coefficient curves of the two types.

To compare supports of different characteristic impedance, the characteristic impedance Z must be determined from the dielectric constant and the diameter ratio D/d or D_1/d,

i.e. $Z = \dfrac{60}{\sqrt{\varepsilon}} \log_e \left(\dfrac{D}{d}\right)$ or $Z = \dfrac{60}{\sqrt{\varepsilon}} \log_e \left(\dfrac{D_1}{d}\right)$

depending on the type of support. The characteristic impedance Z is here a purely theoretical quantity, present in the precise centre of the support, where the field is homogeneous within a short axial length. Supports of similar construction but differing characteristic impedance can be compared with each other by relating Z to Z_0, the characteristic impedance of the homogeneous line at both sides of the support. The nearer Z/Z_0 is to unity, the lower is the original disturbance to be compensated by the excess self-inductance of the line section of characteristic impedance Z. In Fig. 22 Z/Z_0 is plotted against B/D with the characteristic impedance as parameter. It is shown that for supports of similar construction the curves for 50 and 60 ohms largely coincide, the 50 and 60-ohms supports being about equal in performance, while the 75-ohms support shows a less favourable ratio Z/Z_0.

The support for grooved inner and outer conductors has always a better ratio Z/Z_0 than that for a groove only in the inner conductor, since its original field disturbance is lower.

While for diameter transitions the 60-ohms line was shown to afford optimum compensation, this does not apply to supports. In many cases the 50-ohms support has a better Z/Z_0 ratio. The reason is that for the diameter-transition, minimum field strength at the inner conductor is obtained at 60 ohms, while the support, due to its dielectric, is far from the point of minimum field strength. To obtain minimum field strength at the inner conductor the diameter ratio should be D/d or $D_1/d = 2·72$. In practice, optimum dimensions for supports according to Fig. 16 with $B/D = 0·2$ yield the values $D/d = 5·5$ for $Z = 50$ ohms and $D/d = 7·7$ for $Z = 60$ ohms.

Hence for both characteristic impedances the diameter ratios are far from the value for minimum field strength although in the same direction. The 50-ohms line, being nearer to the optimum, has a better Z/Z_0 ratio than the 60-ohms line. This is shown in Fig. 22 especially for broad supports of $B/D = 0·3$,

where the 50-ohms curve runs below the corresponding 60-ohms curve. With thin supports of $B/D = 0.1$ the distorted field in the air preponderates in effect over the distorted field in the dielectric, and conditions change in favour
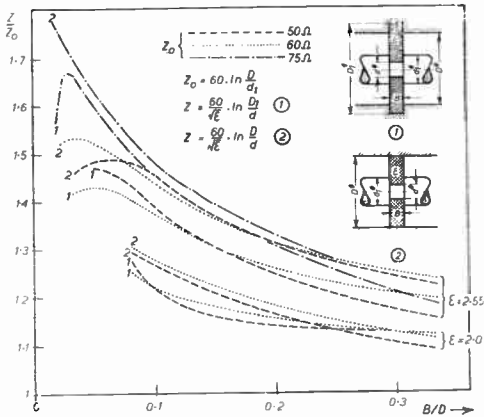

Fig. 22. The calculated value of the characteristic impedance Z in the dielectric support, related to the characteristic impedance of the smooth line Z_0, for various thicknesses B/D.

of the 60-ohms support, the curves intersecting at about $B/D = 0.2$ for a dielectric constant of 2·55 and at a much lower ratio for a dielectric constant of 2·0.

7.6. *Very Thin Supports of $B/D < 0.1$*

It has been shown that for supports in the range B/D between 0·1 and 0·3 the groove ratio D/d becomes worse with decreasing breadth of support, since the disturbing capacitance to be compensated (C_s, Fig. 21) remains approximately constant. When now the support breadth is further reduced, the curves make a turn and rise again (Figs. 16 and 19). Therefrom the conclusion can be drawn that with very thin supports the disturbing capacitance C_s decreases. Apparently the two side walls including the dielectric are so near each other that the lines of force entering the dielectric from both sides are straightened by the repulsive force of the field and so the field becomes more homogeneous (Fig. 23). With very thin supports a uniform electrical field must be assumed within the dielectric and in the air space between the conductors. At the low characteristic impedance of 50 ohms the turn-

ing point of the curve lies at relatively thick supports, e.g. in Fig. 16 at $B/D = 0.06$, while for 75 ohms the turning point is at $B/D = 0.01$. This fact agrees with the theory of the electric field, for, in the 50-ohms line, where the spacing between the conductors is small, the electrical lines of force can distort only slightly, while in the greater space of the 75-ohms line the deflection is correspondingly greater. It is possible that this effect reduces reflection in supports that are composed of many very thin layers. The same effect would result from grooves consisting of a number of sections. The metal remaining between the sections would then act as a field director.

7.7. *Influence of Detail Structure on the Reflection Coefficient Curve*

The magnitude of the reflection coefficient depends principally on the geometry of the discontinuity. Thin supports in general have a lower reflection coefficient than thick ones. Another criterion of quality is the groove ratio D/d or the ratio Z/Z_0 of characteristic impedance in the support referred to the homogeneous line. Accordance between these criteria and the measured reflection coefficients is, however, not dependable; it happens frequently that supports behave in a way different from that expected. Broad supports often have a lower
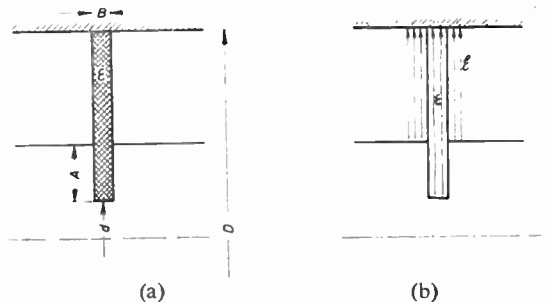

Fig. 23. (a) Very thin dielectric supports, (b) the largely homogeneous field associated with (a).

reflection coefficient than narrower ones; the supports for grooved inner and outer conductors may not show a better reflection coefficient than supports for a grooved inner conductor only, although they have definitely better D/d and Z/Z_0 ratios. The breadth B/D and ratio of characteristic impedance Z/Z_0 do not permit

unambiguous conclusions to be drawn as to the reflection coefficient, especially in the case where no previous experience applies. The unknown factor, of considerable influence on the reflection coefficient of compensated discontinuities, is the detail structure of the electrical field, i.e. the precise arrangement of the individual four-terminal networks of which the discontinuity is composed and their frequency-dependent reactances. The precise form taken will depend on the geometry, dimensions and the frequency. It may, for instance, happen that in a broad support four-terminal networks with opposite reactance curves follow each other, accidentally effecting very good compensation of the overall reflection, while in a thin support the four-terminal networks happen to be arranged so that their reflections add.

It has been pointed out in the foregoing that —for a compensated discontinuity of the types discussed—compensation can only be made on an average basis, where detail structure is more or less neglected. The reflection coefficient of the compensated discontinuities described is in fact the result of the remaining apparently random, variations of detail structure. Thus divergence in reflection coefficient between individual cases is relatively great even for similar designs. The measurements show that detail structure can influence the reflection coefficient by a factor about 2. Cases where this value is exceeded are statistically possible, but seldom occur. Reliable information on the behaviour of the reflection coefficient of a compensated discontinuity can at present be obtained only from a carefully measured curve.

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News from the Sections . . .

South Midlands Section

The Section held its third meeting of the session on 27th November at the North Gloucestershire Technical College when a paper —"A Vidicon Camera Channel"—was read by Mr. B. S. Pover (Associate Member) of the B.B.C. Engineering Training Dept., Evesham.

Mr. Pover began by describing the principle of operation and construction of the vidicon tube. Dark current, gamma characteristics and the frequency response of the channel, with details of the aperture distortion and cable correction circuits, were then considered in turn. The last section of the paper was devoted to the derivation and application of the pulses required to drive the channel to produce the standard television waveform.

Mr. Pover followed the presentation of his paper by giving a demonstration of the camera channel. During the subsequent discussion the author answered questions on target print-through, lag, colour response, size and sensitivity. Comparison with other tubes and noise were other points raised. G. W. M.

(*Editor's note.*—Members who attended the Institution meeting in London on 27th January will have seen in operation the vidicon camera channel which Mr. Power described. The paper which was presented on that occasion dealt with the training of operators and technical assistants concerned with television studio broadcasting, and the camera channel was used in conjunction with a number of monitor receivers to show the techniques of studio lighting.)

North Western Section

At the meeting held in Manchester on 10th December, Mr. P. Huggins (Associate Member) gave a most interesting lecture on "Learning Machines." He described against a background of simple machines which he demonstrated, the basic functions of machines that learn.

Commencing with a circuit which emulated the behaviour of a Pavlovian dog, Mr. Huggins illustrated how a machine can apparently "learn by association." He then took his argument further, by exploring the possibility of the machine taking advantage of what it has learnt. A simple games-playing machine was then demonstrated and its design as a learning machine described. Bar graphs were shown which gave the results of experiments where this machine played in turn against a child of eleven-plus grade, a dice and a machine programmed to win.

From these experimental games Mr. Huggins deduced, firstly, that learning machines provide a cleverness not specifically detailed by the designer. Secondly, this cleverness is abstracted from external sources (opponents) and is stimulated by the richness of such experiences. Machines can abstract this cleverness from a random input. Thirdly, self-organizing machines can be misdirected by allowing them to attain their objective on false premises.

Mr. Huggins then went on to describe practical applications of learning machines and their techniques in industry. He described the Solartron Key Board Instructor and the Westinghouse Electric Corporation's "Automex." He also explained how learning machines may be of use in such fields as pattern recognition, error correction and automatic inspection. He concluded with a brief description of some work he hopes to attempt in which a learning machine is trained by a process operator and then takes over the task from him. F. J. G. P.

Merseyside Section

The new session opened on 2nd November with a lecture by Mr. A. Bedford, B.Sc., on "The Electronics in the Autopilot of the Firestreak Missile." Mr. Bedford gave a detailed description of the circuitry associated with the missile and indicated how the control functions were performed.

These circuits were mainly of the high gain d.c. amplifier variety using thermionic valves and magnetic amplifiers. Analogue computer techniques were used extensively in the control functions, the main operations being conducted using servomechanisms. The vibration and temperature range which the apparatus had to stand was very extensive and needed careful mechanical design, while an efficient cooling

system having a very compact layout had to be designed. In spite of all these difficulties it was claimed that high reliability had been achieved.

On 10th November an enthusiastic audience enjoyed a lecture on "The Use of Transistors in Communication and Control Equipment" given by Mr. E. Wolfendale, B.Sc.(Eng.). His very informative paper was well supplemented by various interesting pieces of equipment.

Mr. Wolfendale opened his paper by discussing the factors affecting the stability of h.f. transistor amplifiers, referring to the development of stabilizing circuits using the correct amount of feedback for unilateralization. The usefulness of the diffused alloy transistors for u.h.f. circuits was explained and a very convincing demonstration of a transistorized u.h.f. receiver was performed. Mixer circuit techniques for transistors were then discussed and details given of i.f. amplifier techniques.

Mr. Wolfendale concluded his paper by discussing logic circuitry in computers with a comparison of resistor transistor, transistor, ferrite core and diode logic. A ferrite core matrix store with its associated transistor circuits was shown.   G. B.

West Midlands Section

On 8th December the Section held its first meeting in Birmingham for several years at the request of the growing number of members in the area who find this to be a more convenient centre than Wolverhampton, where meetings are usually held. A paper was presented by Mr. P. J. Guy, a broadcasting engineer, on "The Development of H.F. Tape Recording."

Mr. Guy's paper opened with an explanation of the need for recording in modern television techniques. This covered widespread aspects of programme production including programme timing, editing, economy in studio equipment and personnel, together with the special cases of certain outside features which would be extremely difficult to present without the aid of high-quality tele-recording.

Following this introduction, Mr. Guy reviewed the various complex optical and mechanical techniques involved in photographing a flying spot scanning a picture. Much of the earlier work in this field was on the development of systems compatible with standard ciné equipment, an outstanding problem being the need for fast frame pull-down in order to obtain the maximum ratio of scanning time to film handling time. Several examples of the ingenious systems employed were described.

Next, Mr. Guy went on to discuss modern magnetic tape recording systems and their advantage in providing high picture definition. The relationship between signal bandwidth and scanning speed was developed from first principles, then consideration was given to the mechanical and electrical methods by which this high information reading rate was obtained. This involved high accuracy pulse-synchronized servo systems of extremely sophisticated design and an interesting pneumatic system for forming the tape to a precisely controlled radius during reading. In conclusion, the arrangement of picture and control information tracks on the tape was considered, and a sample of tape which had been specially etched to outline the track layout was provided for inspection. F. D.

North Eastern Section

At the fourth meeting of the session on 13th January a paper on "Electronic Business Machines" was presented by Mr. J. Allan and Mr. D. Keating.

It was contended in their paper that the conventional computer has serious limitations for many business problems. Such problems demand, not so much extreme rapidity of computation, as fast and economical reference to, and maintenance of, very large files of information. In some cases it is necessary not only to store large quantities of semi-permanent data but also to be able to refer at random to any portion of such data without consideration of record sequence. There is therefore considerable need in the busines world for special purpose machines partaking of many of the characteristics of the conventional computer, but making available unusual resources of storage and of access thereto.

Some of the devices which have been developed for storage both by the authors' company and other organizations were then discussed. Those present found that the paper provided a useful introduction to the proceedings of the Symposium on "Large Capacity Storage Systems" which were published in the January 1960 *Journal*.

The Combined Television – Radio Receiver and its Problems †

by

R. S. HILDERSLEY, ASSOCIATE‡

A paper read on 2nd July 1959 during the Institution's Convention in Cambridge.

Summary : The large number of frequency allocations in and around Band II compared with Bands I and III causes a serious selectivity problem in domestic combined receivers for television and f.m. sound broadcasts. The sound bandwidth of a television receiver is usually of the order of 500-1000 kc/s and since the frequency allocations of commercial radio transmitters are liable to be within 500 kc/s of the B.B.C. transmitters, interference can occur. Interference can also occur between the various B.B.C. regional transmitters whose frequency separation is usually 400 kc/s, but may be as little as 200 kc/s. The problem and a number of different solutions are discussed in detail. The circuit details of a commercially available combined receiver are then described. The sound i.f. circuits incorporate a double superheterodyne system, and the frequency of its r.f. oscillator is stabilized in Band II by means of an inexpensive a.f.c. system. Methods are described for measuring accurately the selectivity of a Band II receiver.

1. Introduction

Since the inauguration of a sound broadcasting service in the v.h.f. Band II, a new type of domestic receiver has been evolved. This is basically a television receiver which can be switched to receive Band II sound transmissions. When receiving a f.m. signal the television scanning circuits and cathode-ray tube are inoperative. The r.f. unit is provided with tuned circuits to cover Band II and the sound i.f. circuits are followed by a suitable frequency-modulation detector. The audio-frequency circuits and power supply are common to both sections of the receiver. The problems in such receivers will be discussed and the circuit arrangements for a commercially available receiver that represents one solution described in detail.

2. Statement of Main Problems

2.1. *Selectivity*

The main difficulty in a combined receiver is to achieve sufficiently good selectivity.

Interference from powerful mobile transmitters close to the band, or from adjacent channels within the band, impose stringent requirements on a receiver. Transmitter frequency allocations in Band II differ radically from those in the television Bands I and III, in that for most areas of the United Kingdom the frequency separation on Band II between geographically-adjacent sound transmitters is usually 400 kc/s. Sound receivers are designed to cater for these conditions and, with i.f. circuits operating at the standardized frequency of 10·7 Mc/s, have little difficulty in achieving satisfactory performance.

In the case of television receivers, however, the sound and vision i.f. circuits operate at the standardized frequencies of 38·15 Mc/s and 34·65 Mc/s respectively. Pass-band responses for the i.f. circuits of a typical television receiver are given in Fig. 1, from which it will be seen that the sound i.f. bandwidth is approximately 700 kc/s for 6 db attenuation. This degree of attenuation is adequate for rejecting both the accompanying and adjacent vision signals, and it is undesirable greatly to reduce the sound bandwidth of a television receiver for several reasons. The relatively wide bandwidth permits some latitude in setting
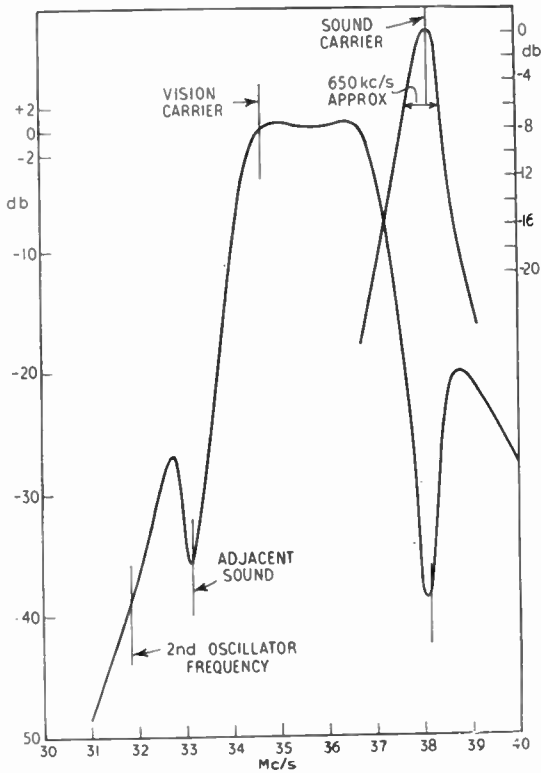
Fig. 1. Passband response for typical television receiver.

wavebands. Although each region provides a channel spacing of 2·2 Mc/s, geographically adjacent regions are separated by only 400 kc/s or 600 kc/s. If an unselective receiver were sited at the fringe of two areas broadcasting the same programmes, distortion and difficulty in tuning would occur. Were the programmes different, however, the interference might be such that the unwanted programme would be heard in the background. Radio telephone systems operated by various organizations such as taxi services can also cause severe interference. Although operating at relatively low power, the local field strength of their base transmitters can be comparable or even greatly exceed that of the B.B.C. transmitters. This applies particularly in B.B.C. fringe areas. Since the frequency separation may be only approximately 600 kc/s and also because of very different modulation material, severe " break-through " interference can occur. The point is illustrated graphically in Fig. 3, where the selectivity curves of two typical receivers of the type under discussion are compared. In this graph the selectivity has been measured by the two signal method, a technique which will be described later.

the frequency of the local oscillator which can thus be adjusted for optimum picture quality. It also reduces the requirement on oscillator frequency stability (i.e. thermal drift). Another important aspect is the cost of the tuned circuits necessary to decrease the bandwidth.

While there is no reason to reduce the sound bandwidth of a television receiver, trouble arises when the conventional sound i.f. circuits are adopted for f.m. sound reception in combined receivers. In that event a receiver cannot provide s u ffi c i e n t discrimination against adjacent signals spaced by 400 kc/s, and reference to Fig. 2 will show when such a requirement will arise. The diagram shows the present B.B.C. transmitter frequency allocations in Band II, and also allocations for the adjacent

2.2. Aerials

In many cases combined receivers are installed by using the existing television aerials for Band II reception. Although the resulting signal pick-up may be sufficient, the arrangement is poor. Not only are the aerials tuned
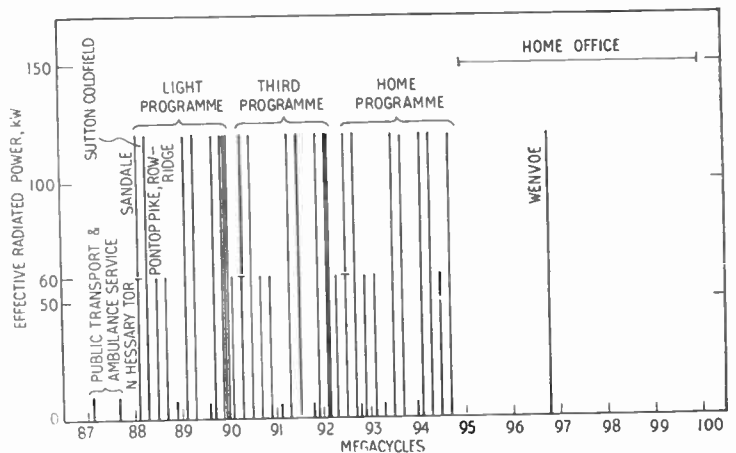

Fig. 2. Frequency allocations in Band II.

to the wrong frequency, but they are also usually vertically polarized whereas the Band II transmissions are horizontally polarized.
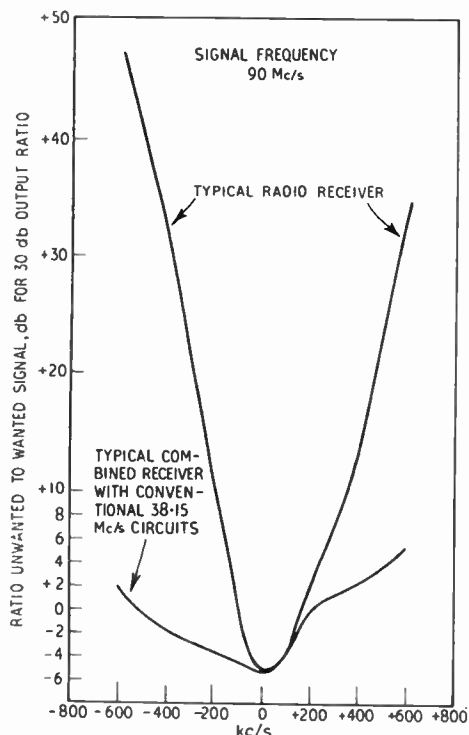

Fig. 3. Band II selectivity. Typical radio receiver compared with typical combined receiver with 38·15 Mc/s sound i.f. circuits.

Such an aerial is relatively inefficient on Band II, and cannot be expected to have good directional properties. The signal input to the receiver may be below the level at which good a.m. rejection is available, and severe distortion resulting from multi-path reception may arise. It is this susceptibility to multi-path reception that reveals another disadvantage of the combined receiver when compared with a conventional Band II sound receiver.

2.3. *A.M. Rejection*

Ratio detectors operating at 38·15 Mc/s are usually found to be lacking in both a.m. rejection and sensitivity when compared

with 10·7 Mc/s detectors. Optimum performance depends upon the balance of such small values of capacitance and inductance at 38·15 Mc/s that production specimens tend to vary considerably from a given standard. a.m. rejection suffering in particular.

3. Proposals to Increase Selectivity

In the broadest terms, the aim is to produce a combined receiver with selectivity performance comparable with that of a typical conventional Band II sound receiver. Such a receiver normally has five i.f. tuned circuits operating at 10.7 Mc/s. Three main methods are now described for increasing the selectivity of a combined receiver.

3.1. *Method 1 : Increasing the number of tuned circuits at 38·15 Mc/s*

The block diagram of Fig. 4 shows the circuit arrangement of a typical television receiver having one common i.f. stage followed by one sound and one vision amplifier. It is seen from this diagram that at most only four tuned circuits operate at 38·15 Mc/s. As all the tuned circuits preceding the sound i.f. valve are required to respond both to sound and vision signals, they do not greatly assist in the quest for selectivity. It is estimated that approximately nine tuned circuits at 38·15 Mc/s would be needed before the performance could meet the requirements. An additional six circuits would therefore be necessary in this case. Such a receiver would be expensive and difficult to produce. The ratio detector would continue to operate at 38·15 Mc/s with all its attendant difficulties. It seems that a better approach to the problem is to reduce the operating frequency of the f.m. sound i.f. circuits.
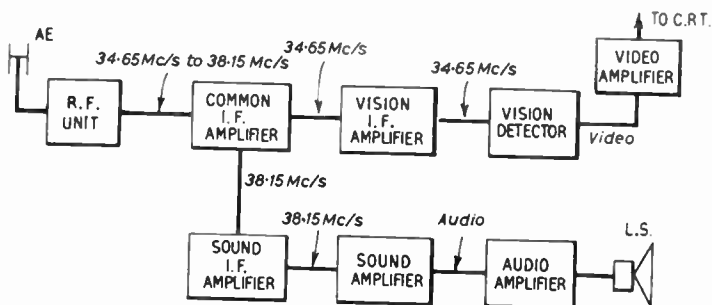

Fig. 4. Block diagram of typical television receiver.

3.2. Method 2: An i.f. amplifier responding to two frequencies

It is possible to design an i.f. amplifier that will respond to two frequencies. For a combined receiver a logical choice of frequencies is 38·15 Mc/s and 10·7 Mc/s. Referring to the block circuit diagram of Fig. 5 it will be seen that an additional pair of tuned circuits can be incorporated between the frequency changer and common i.f. stage. This new transformer will have optimum coupled circuits and a comparatively narrow passband at 10·7 Mc/s, whilst the bandwidth between 34 Mc/s and 38·15 Mc/s remains as for a normal television receiver. It is not necessary to switch the circuits. Between the common i.f. and sound i.f. amplifier valves are two transformers; one resonates at 10·7 Mc/s, the other at 38·15 Mc/s.

The anode of the sound i.f. amplifier feeds into two tuned circuits. One is the normal 38·15 Mc/s sound a.m. detector, and the other is a ratio detector operating at 10·7 Mc/s. These two demodulator systems are entirely separate, and a simple change-over switch is provided to select the audio-frequency output of one or the other.

There are three main disadvantages to this proposal. (1) Cost—The additional 10·7 Mc/s circuits are employed only when the receiver is operated on Band II. Two 10·7 Mc/s i.f. transformers, the ratio detector with its diodes and associated components, are completely inoperative on television. (2) Performance— Although adequate i.f. selectivity should readily be achieved on Band II, the circuit would probably lose overall gain at 38·15 Mc/s as compared with a simple amplifier. The interposition of a pair of 10·7 Mc/s tuned circuits into a single interstage coupling has been found to reduce the gain of the existing 38·15 Mc/s circuit by about 2db. As care has to be taken in the design of 38·15 Mc/s sound i.f. amplifiers to eliminate a tendency towards instability, the additional 10·7 Mc/s circuits may be expected
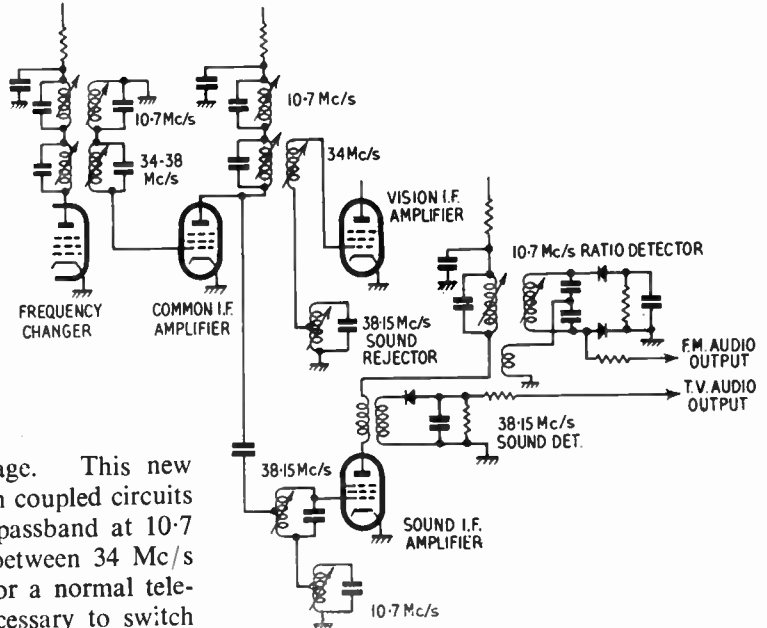

Fig. 5. Circuit diagram of i.f. amplifier responding to two frequencies.

to aggravate this problem. Maximum possible stage gain with stability at 38·15 Mc/s would almost certainly be reduced by more than 2db per stage. (3) Because of the broadly tuned r.f. circuits, it may be difficult to secure adequate image rejection. To prevent the oscillator 2nd harmonic from falling in Band III, it is required that the oscillator shall operate below signal frequency in a Band II receiver with 10·7 Mc/s i.f. circuits. Consequently the image frequency for, for example a 97 Mc/s carrier is 75·6 Mc/s which is within the band allocated to public utility radio services. Although a combined receiver of this type will represent an improvement over the conventional one with 38·15 Mc/s i.f. circuits, it will have become liable to second-channel interference. The problem is particularly acute when a single set of broadly tuned coils is provided to cover the entire Band II, with an oscillator having a continuously variable frequency.

3.3. Method 3: The double superheterodyne

Another method of reducing the operating frequency of the f.m.i.f. system is to employ a double superheterodyne circuit.

The method offers the choice of operating the double superhet only for Band II reception, or

both for this and television. In the case of the first alternative, if a suitable frequency converter valve were employed in an early position of the i.f. amplifier system, it would be arranged to operate as a conventional amplifier for television reception. For Band II reception, however, it would be switched to operate as a frequency changer. An i.f. amplifier resonating at two frequencies would follow, and it could be of the type described in Section 3.3. Because the first i.f. is 38·15 Mc/s the image rejection should be satisfactory. The cost would be even higher than that of method 2 owing to the second frequency changer valve and the additional switch contacts required. A double superheterodyne circuit is potentially an economical one if it can be made to operate continuously both for television sound and Band II reception. The number of components that are then unused in one or the other condition can be reduced to a minimum.

The chief difficulty likely to be encountered results from the new oscillator circuit, some care being necessary in selecting the best second i.f. Figure 6(a) is a circuit diagram showing an arrangement for a receiver of this type. It will be seen that there are basically three tuned circuits operating at this new intermediate frequency, the ratio detector operating as though it were a single tuned circuit. As the overall selectivity due to these tuned circuits must be comparable with that of an amplifier having five tuned circuits at 10·7 Mc/s, it is advantageous to select the lowest practicable second intermediate frequency. Another reason why 10·7 Mc/s is an unsuitable choice of second i.f. for a combined receiver with a continuously-operating double superhet sound i.f. circuit is that the second oscillator must operate either at 48·85 Mc/s or 27·45 Mc/s, the first i.f. being 38·15 Mc/s. If the higher frequency were chosen it could possibly interfere with channel 2 sound carrier at 48·25 Mc/s; also its 4th harmonic lies close to channel 9 vision carrier. The lower second oscillator frequency also presents difficulties because the second harmonic occurs at 55 Mc/s, which is within channel 3 vision passband. Although these difficulties could possibly be overcome, experience with this type of receiver shows that a second i.f. of about 6 Mc/s is a satisfactory choice.

4. Circuit Details of a Combined Receiver with Double Superhet Sound I.F. System

4.1. *General Introduction*

The receiver to be described is of the double superheterodyne type. The r.f. unit incorporates switched channel selection on Bands I, II, and III, and the oscillator frequency is stabilized on Band II by means of an a.f.c. system. The circuit diagram of Fig. 6(b) shows the essential details of the receiver.

4.2. *The R.F. Unit*

For Band II reception the r.f. tuned circuits of this unit are pre-tuned to the centre of the required frequency range. Figure 2 shows that this range extends from 88 to approximately 98 Mc/s, although most of the B.B.C. transmitters are located between 88 and 95 Mc/s. It will also be seen that this band is divided into groups for the three sound programmes, i.e., Light, Third and Home, in ascending order of frequency. The r.f. circuits of the tuner are consequently arranged to have a bandwidth appropriate to these frequency groupings (Figure 7). The oscillator coil inductance can be varied by means of a small brass slug screwed into the threaded coil former. This is exactly similar to the method used for tuning the oscillator circuit on the television channels.

4.3. *Variable Reactance and A.F.C. System*

It is well known that the capacitance of a junction diode can be varied according to the magnitude of an applied d.c. bias potential.[†] This effect occurs under reverse bias conditions. Negative bias is applied to the anode, sufficient completely to cut off all current. Increasing the bias potential results in a reduced capacitance between the diode anode and cathode. Such a junction diode may be used as part of the tuning capacitance of a v.h.f. oscillator circuit whose frequency is then dependent upon the magnitude of the diode bias voltage.

Since no bias current flows, a simple a.f.c. circuit can be developed in which the d.c. output from the discriminator is used to vary the fixed bias potential applied to the diode. The

† G. G. Johnstone, " Junction diode a.f.c. circuit," *Wireless World*, **62**, pp. 354-355, August 1956.

circuit suffers from two disadvantages, however—firstly, high cost; a junction diode tends at present to be too expensive a component to incorporate in a quantity produced receiver. Secondly, the frequency stability of the circuit depends upon a stabilized bias supply. The a.f.c. system proposed by the author does not suffer from either of these disadvantages, and it makes use of a point-contact germanium diode as the variable reactance device. The application of this diode into an oscillator circuit is basically similar to the one using a
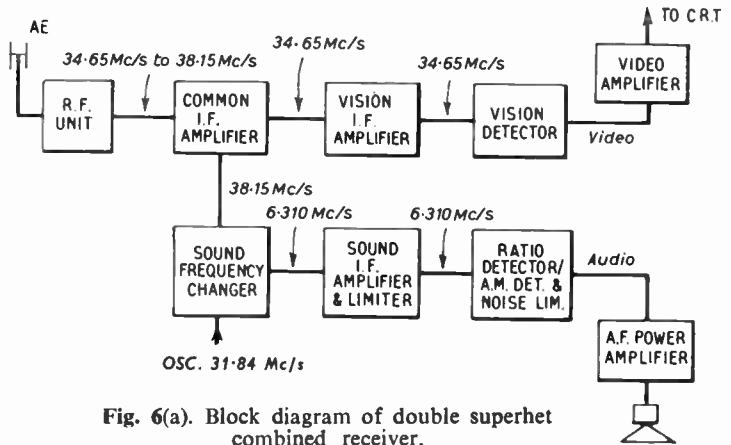

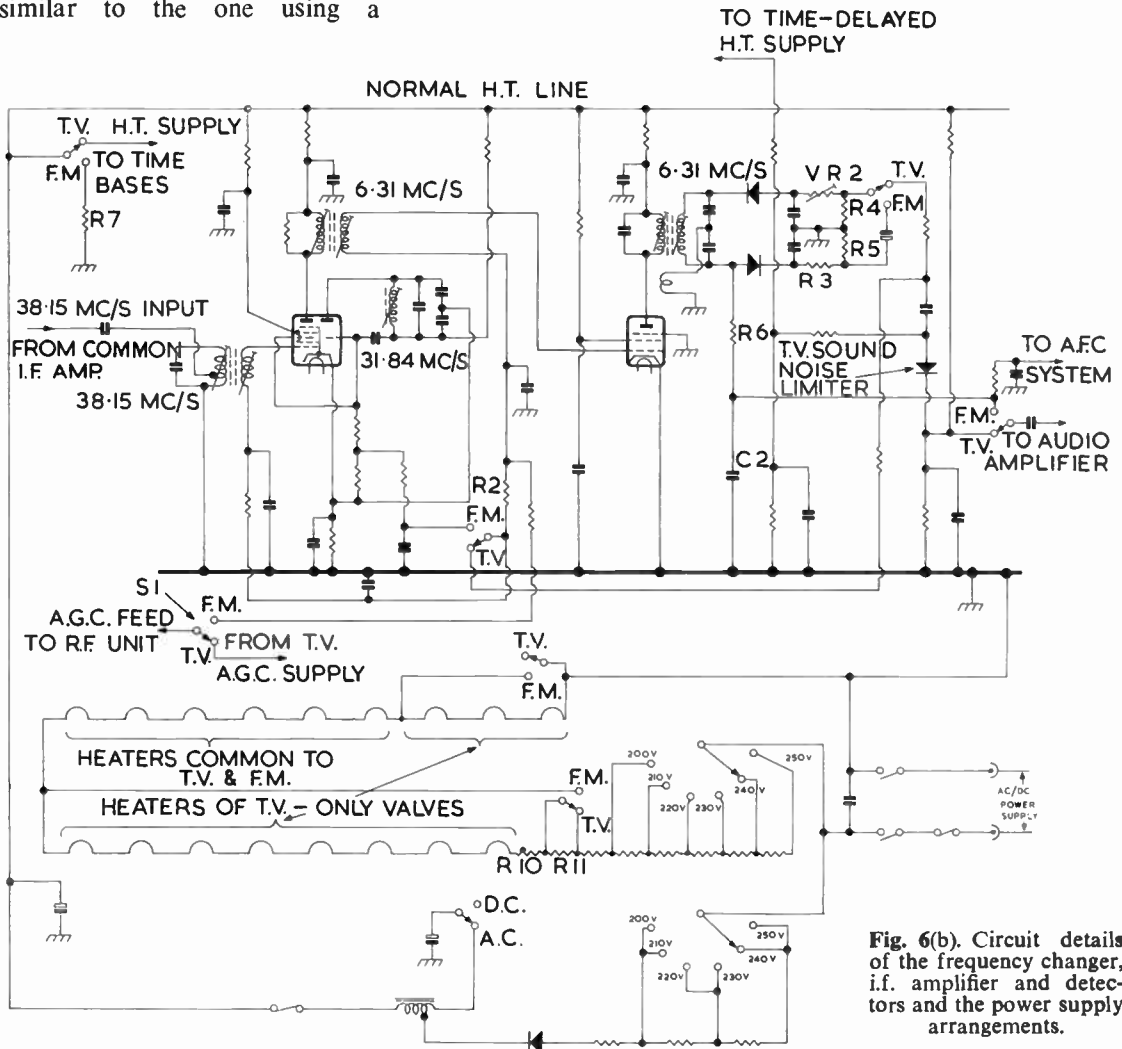Fig. 6(a). Block diagram of double superhet combined receiver.

Fig. 6(b). Circuit details of the frequency changer, i.f. amplifier and detectors and the power supply arrangements.

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junction diode, although the operat-
ing characteristics are very different
as the point-contact diode is
operated in the conducting con-
dition. Variation of an applied
bias potential results in a change in
both rectification current and oscilla-
tor frequency. The frequency is
inversely proportional to diode
current, an increase of current
resulting in a reduction of frequency.
Reference to Fig. 8 shows this
dependence of frequency upon current. Once
the diode is biased beyond the cut-off point
there is negligible further change in oscillator
frequency. Although achieving the same result,
the point-contact diode operates in a different
manner from the junction diode, the frequency
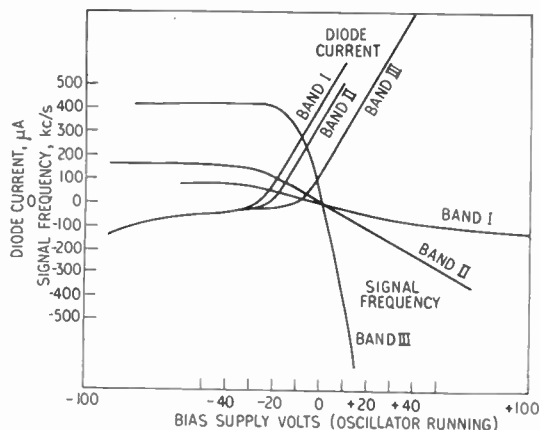characteristic being sensibly linear.

Fig. 7. Bandwidth curves for Band II coils.

of the maximum current that can flow through
the crystal. However, since the resistor is con-
nected directly across part of the oscillator
tuned circuit, the lower limit for the series
resistor is set by the requirements of oscillator
drive voltage.

As this a.f.c. circuit is essentially a
balanced bridge network, it is inherently inde-
pendent of supply voltage fluctuations. In
practice a loop gain figure of the order of
10 times is achievable, reducing frequency
drift errors by this amount on Band II. At
the present stage of development the a.f.c. sys-
tem is rendered inoperative on Bands I and
III. Although experiments indicate that it
could function on these bands, it is considered
not to be so necessary.

4.4. The Second Frequency Changer and I.F. System

In Section 3.4 reference was made to the
choice of operating frequencies for the second
oscillator and i.f. circuits. In order to reduce
interference resulting from harmonics of this
oscillator, a frequency of 31·84 Mc/s has been
selected. Since the first sound i.f. is 38·15

Fig. 8. Frequency vs. current curves for point-
contact germanium diode.

Because rectification current flows through
the circuit, it is not possible to connect the
variable reactance directly to the discriminator
circuit. The d.c. potential output from the
discriminator is consequently fed to a low-gain
d.c. amplifier, with the variable reaction crystal
coupled to its anode. The basic circuit is
given in Fig. 9. The potentiometer RV1 is
used to set the quiescent crystal current, this
being approximately 200 microamperes. The
resistor in series with the crystal diode
controls the loop gain of the system. Reducing
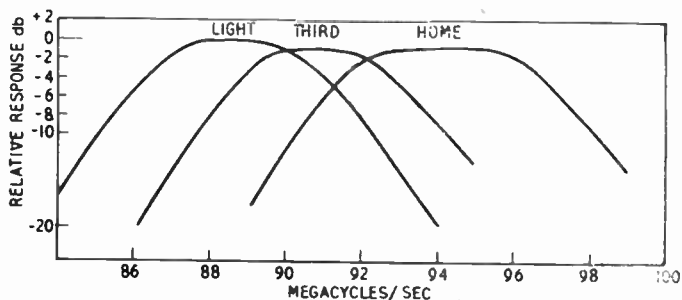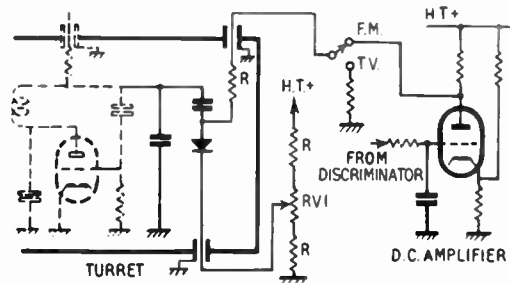the value of the resistor results in an increase

Fig. 9. Basic circuit of a.f.c. system.

Mc/s, the resulting second i.f. becomes 6·310 Mc/s. Harmonics of these two frequencies, which the author does not regard as the only possible choice, are given in Table I. Radiation of these harmonics from the second frequency changer stage must be reduced to the lowest practicable level. This may be achieved by suitable attention to the chassis layout, and the reduction to a minimum of all component lead lengths, particularly cathode and screen

Table I

Harmonics of Sound Oscillator and Second I.F.

Fundamental		31·84 Mc/s	6·310 Mc/s
2nd Harmonic		63·68	12·620
3rd	,,	95·52	18·930
4th	,,	127·36	25·240
5th	,,	159·20	31·550
6th	,,	191·04	36·860
7th	,,	222·88	43·170
8th	,,	254·72	49·480

decoupling capacitors. In addition to these harmonics, the oscillator fundamental of 31·84 Mc/s must be considered. Since the vision i.f. carrier is 34·65 Mc/s the difference frequency is 2·81 Mc/s. Reference to Fig. 1 will show the relative response of a typical vision i.f. system at this frequency. Although the response at 31·84 Mc/s compared with 34·65 Mc/s is of the order of − 40 db. it is evident that a 2·81 Mc/s beat frequency pattern could occur if the second oscillator system were allowed to couple sufficiently into the vision i.f. circuits. This is a problem that governs the choice of second frequency changer valve to be used. Although a triode pentode yields a high conversion conductance, the 38·15 Mc/s signal grid input circuit with such a valve also contains a large component of oscillator frequency. This then enters directly into the vision i.f. circuits via the relatively low impedance coupling of the sound input circuit. By comparison, the triode-heptode type of valve results in an appreciably lower stage gain. It is, however, inherently superior because of low internal capacitances between the signal and oscillator grids. Not only is the sound oscillator to vision coupling reduced to a

practicably low level, but it is also possible to apply an a.g.c. voltage to the input grid if required, the oscillator frequency being unaffected by a.g.c. bias.

4.5. *Second Oscillator Frequency Stability*

As stated previously, it is necessary to consider up to the 6th harmonic of the second oscillator when designing this type of receiver. Study of the permissible frequency latitude adjacent to the 6th harmonic will indicate the degree of frequency stability that is required from the circuit at fundamental frequency. It is found that this latitude does not exceed ± 100 kc/s approximately, i.e., 15 kc/s for the fundamental at 31·84 Mc/s. A Colpitts oscillator is used. Coil and tuning capacitors C
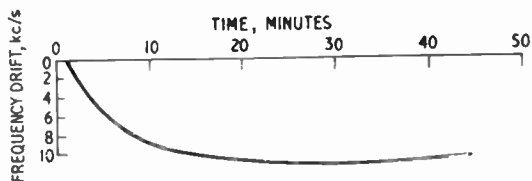
Fig. 10. Thermal frequency drift curve for second oscillator.

and C1 having inherently stable thermal characteristics are selected, and any residual compensation that may be necessary is provided by selecting a suitable temperature coefficient of capacitance for capacitor C1. The curve of Fig. 10 shows the degree of stability achieved in a typical receiver.

4.6. *6·31 Mc/s Sound I.F. amplifier*

Apart from the lower operating frequency, the sound i.f. circuits do not differ appreciably from those of a conventional Band II sound receiver. There are effectively three 6·31 Mc/s tuned circuits, i.e., the band-pass transformer coupling between the frequency changer and i.f. amplifier and the ratio detector which behaves as a single tuned circuit. Since these circuits operate both for television sound and f.m. reception, however, the overall bandwidth must satisfy both requirements. A compromise solution results in overcoupling of the band-pass transformer, the overall amplitude responses of the three circuits being as shown in Fig. 11. This "flat-topped" response allows some latitude in setting the frequency

of the first oscillator, and also renders frequency drift less noticeable when receiving television sound. At the same time the system has sufficient linearity in the f.m. condition to result in a low percentage of harmonic distortion. The i.f. amplifier is of conventional type. When switched to receive f.m. signals,
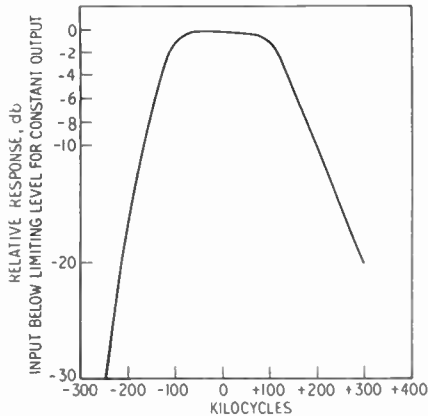

Fig. 11. Amplitude response of experimental 6·310 Mc/s i.f. amplifier.

it operates with a low value of grid bias; limiting occurs when the input signal exceeds the value of this bias. The voltage developed across R2 as a result of grid current is then fed via S1 to the r.f. amplifier grid. A measure of a.g.c. is thus obtained for signals whose amplitude is sufficient to operate the limiter stage.

4.7. The Ratio Detector

In describing the ratio detector system for this receiver no attempt will be made to discuss details of the many parameters involved, which follow well established principles. It will suffice, therefore, to say that by operating the ratio detector at the comparatively low frequency of 6·31 Mc/s, the following advantages have been secured as compared with a circuit operating at 38·15 Mc/s:

(1) The performance of the circuit is dependable, production specimens following closely the design prototypes.

(2) It is relatively easy to achieve good balance over a long period of time. This stability is essential for the a.f.c. system to operate satisfactorily.

(3) The performance of germanium diodes compared with thermionic diodes is better at 6·31 Mc/s than at 38·15 Mc/s. The full benefits of the crystals can thus be realized, i.e., small size, low noise level, and efficient screening.

(4) Finally, the low frequency ratio detector has, in practice, yielded a more satisfactory performance particularly in the matter of a.m. rejection.

In order to operate the a.f.c. circuit, it is necessary to extract from the ratio detector a d.c. output from the discriminator. The voltage is zero at centre frequency with the polarities reaching opposite maxima either side of resonance. This requirement demands that a balanced ratio detector circuit be used. Because of the a.f.c., it is very important that the frequency at which maximum a.m. rejection occurs should coincide with that for maximum audio output and zero d.c. potential from the discriminator. The resistors VR2 and R3 in the circuit diagram of Fig. 6 enable this to be achieved. In order to use this detector circuit for a.m. television sound reception the circuit has to be rearranged to a certain extent. Firstly, it is necessary to disconnect the stabilizing capacitor from the diode load circuit. Voltages due to amplitude modulation then appear across the resistors R4 and R5. No such voltages appear at the junction of R6 and C2, across which the audio output due to f.m. appears, because of the balanced circuit. The outputs from the two diodes are equal, and being in opposite phase, cancel out at centre frequency. It is required, therefore, to insert a switch contact to select the audio frequency outputs due to one or the other modulation system. The noise limiter diode which is necessary for a.m. reception is accordingly placed in series with the appropriate circuit and bypassed for f.m.

4.8. Power Supply Switching

When switching the combined receiver from television to the radio reception condition the following operations are carried out, in addition to switching the ratio detector and r.f. tuning circuits.

(1) The heater of the cathode-ray tube is switched off.

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(2) The line and frame scanning systems are rendered inoperative.

(3) Because of (2) the h.t. voltage tends to increase; voltage compensation arrangements therefore are introduced in order to prevent valves and components from being over-run.

(4) Switch contacts are incorporated in the a.g.c. system to select the appropriate source of control bias.

In addition to the c.r.t. heater, the heaters of the line oscillator, output and economy diodes are switched, also the frame oscillator and output valves. The h.t. feed to these circuits is broken when the receiver is switched to the f.m. condition. Because this receiver is designed to operate with series connected valve heaters, a resistor R12 must be inserted into the heater chain when these valves are switched off. Its value equals the hot resistance of the heaters that it substitutes. In this way the current through the circuit remains the same whether the receiver is switched to television or Band II. However, after the receiver has been operating for some time in the Band II condition, an excessive surge of current would occur after switching to television due to the low resistance of the heaters when they are cold. This surge is prevented by the thermistor R10 in series with the switched valves.

Since the time-base circuits of a television receiver account for more than 60 per cent. of the total h.t. consumption, a considerable rise in h.t. voltage occurs when they are switched off. This is prevented by connecting a ballast resistor across the h.t. supply. The power that this resistor must dissipate is comparable to the power consumption of the time base circuits; typically it is about 20 watts. Alternatively, it is possible to limit the rise in h.t. voltage with reduced load by arranging a resistor to be switched in series with the main reservoir capacitor. The power dissipated in this resistor may be only $\frac{1}{4}$ of the dissipation in a ballast resistor of cost, but also the loading on the power rectifier is reduced.

The vision circuits of this receiver incorporate an a.g.c. system which is applied to the i.f. and r.f. amplifiers. However, when operating as a Band II sound receiver, this system is inoperative. In order to enable the receiver to handle strong signals without cross modulation on Band II, an a.g.c. system will still be required. The r.f. amplifier gain is consequently controlled by a bias voltage derived from the grid of the limiter valve. A switch is necessary in the feed circuit to this valve in order to select the appropriate control bias source.

4.9. *A Simpler Power Circuit*

Before concluding this section devoted to power supply switching, reference must be made to the alternative approaches that are possible.

It is possible to render the time-base valves inoperative simply by removing their h.t. supply, the heater circuits being left running as for television reception. The valve characteristics could be impaired by long periods of operation in this condition, a " cathode poisoning " effect taking place. This may be prevented by allowing a reduced current to flow through a resistor in the h.t. feed circuit. The current must be below the level at which the time-base oscillators can function. This system is less costly than that described in Section 4.8, since no additional heater dropping resistors are required. Fewer contacts are required on the TV/FM change-over switch, and because some residual h.t. current is drawn by the time-base valves the h.t. voltage may not rise unduly. An h.t. ballast resistor may not, therefore, be required.

5. F.M. Receiver Bandwidth and Selectivity Measurements

It is possible to derive some information about the bandwidth of an f.m. receiver by applying techniques similar to those normally used for a.m. receivers. This information can be valuable for comparative purposes in the design of a receiver, although reliable results can be secured only by taking suitable precautions.

If the i.f. amplifier of a Band II receiver contains an amplitude limiting stage, this will modify the bandwidth measurement in two ways.

Firstly, as the input signal amplitude increases beyond the level at which limiting

commences, the input-output relationship ceases to be linear. This tends apparently to flatten the peak of the bandwidth curve unless the measurement is made at an input below the limiting level. The measurement must, for the same reason, be made at a constant input. If a constant output measurement were to be attempted, as for an a.m. receiver without a.g.c., limiting would be liable to occur at some point down the side of the response curve.

Secondly, depending upon the design of the transformer and limiter circuit, the passband response of an i.f. transformer will be modified by the limiter action. In the case of a grid limiter, the transformer secondary tuned circuit will be damped by grid current. This loading will also affect the coupling coefficient between primary and secondary windings, and the response changes are liable to be more serious when high impedance circuits are employed.

In spite of these two difficulties, simple and satisfactory bandwidth measurements can be made by using an unmodulated r.f. source, and measuring the d.c. voltage developed across the ratio detector stabilizing capacitor. In order to investigate the relative amplitudes of the central and side responses due to the demodulation circuit, a frequency modulated carrier must be used. Again, the measurement should be made at a constant input level, but instead of measuring the d.c. voltage across the stabilizing capacitor, the a.f. output should be measured in the normal manner.

Whilst these simple bandwidth measurements are of value during the detail design stage of a receiver, a more elaborate system is essential if practical reception conditions are to be simulated in the laboratory.

The resultant of two adjacent modulated carriers is an extremely complex wave form.†
This arises particularly when the two carriers are of comparable amplitude. Because of this fact, and also the complications occurring within the receiver itself, it is necessary to

adopt the two-signal technique of selectivity measurement.

Two r.f. carriers are simultaneously applied to a receiver. Initially these carriers are of identical amplitude and modulated to the same depth; 22·5 kc/s deviation is a convenient level. The receiver audio gain control is adjusted to give a convenient output power
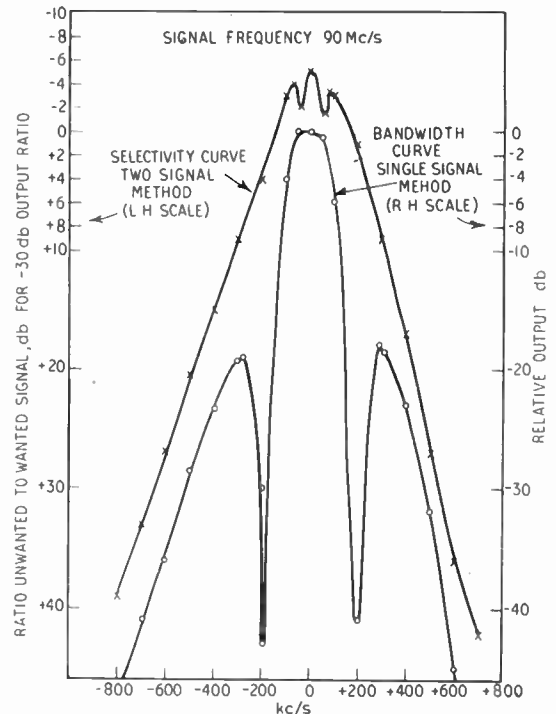

Fig. 12. Overall selectivity of double superhet combined receiver. Comparison of two methods of measurement, conventional, and two-signal method. Input 300 μV at 90 Mc/s.

reference level with only the wanted carrier operating. The modulation of the wanted carrier is then switched off and the signal level of the interfering (unwanted) carrier adjusted until, at any given input frequency, it causes an output to appear in the receiver 30 db below the reference level. The selectivity curve is plotted as a voltage ratio between the two carriers to give this 30 db ratio in output levels. Although the technique provides important information, it is difficult to apply in practice. Several precautions are necessary if incorrect conclusions are to be avoided.

† Murlan S. Corrington, "Frequency Modulation." pp. 323-357, Vol. 1 (R.C.A. Princeton, N.J. 1948).
  G. D. Browne, " A.f.c. in Band II f.m. receivers." *Mullard Tech. Commun.*, **4**, No. 35, pp. 152-157, November 1958.

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(1) The presence of audible beats when the carriers are within approximately $\pm$ 50 kc/s of each other may yield unreliable output readings unless the output meter contains a filter tuned to modulation frequency.

(2) Because of the presence of side responses by the ratio detector, the measurement must take these into account.

(3) The signal generators must possess a high standard of frequency stability.

Examples of response curves of a given receiver taken by the single signal method and by the two signal method are shown in Fig. 12.

6. Conclusions

When this project was begun, the design of a combined receiver having a double superhet sound i.f. system appeared most unlikely to achieve a successful conclusion. The double superheterodyne principle has been used before in conventional radio receivers, but even these have suffered from difficulties due to harmonics of one or the other oscillator. It seemed unreasonably audacious to attempt to use the technique for a television receiver when, theoretically, beat-frequency patterns could occur on every television programme channel. In the event, however, the problems have appeared to be capable of solution. Combined receivers of this type perform satisfactorily as television receivers, whilst the Band II performance has been found to compare well with that of conventional Band II sound receivers. Not only is the adjacent channel selectivity of a comparable order, but the usable sensitivity may be higher. This is because the performance of the r.f. unit fitted to a television receiver tends to be superior to the relatively simple and inexpensive circuits used in the majority of Band II radio receivers. The low noise factor of a good television r.f. unit is especially valuable in a combined receiver in this respect.

7. Acknowledgments

Radio Engineering Overseas . . .

The following abstracts are taken from European and Commonwealth journals received in the Library of the Institution. Members who wish to borrow any of these journals should apply to the Librarian, stating full bibliographical details, i.e. title, author, journal and date, of the paper required. 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.

PARAMETRIC AMPLIFICATION

In attempts to design parametric amplifiers for radio-waves, methods of "pumping" usually rely on the use of voltage-dependent capacitances (e.g. germanium diodes) or current-dependent inductances. The theory of these "autoparametric" systems is less amenable to analysis than that of "heteroparametric" systems, in which the oscillation to be amplified does not react on the pumping. A simple heteroparametric system has been described by Dutch engineers which uses a magnetostrictive resonator as pump, the pump frequency being 22 kc/s. With this system they have been able to verify various theoretical predictions. A piezo-electric pump for a pumping frequency of 3 Mc/s is also described.

"Experiments in the field of parametric amplification." B. Bollee and G. de Vries. *Philips Technical Review*, **21**, pp. 47-51, 1959/60. (No. 2). (In English.)

VACUUM MEASUREMENTS

A cold-cathode ionization gauge with axial magnetic field is described in a recent Canadian paper which is capable of measuring pressure in the range 10^{-3} to 10^{-12} mm Hg, and is primarily useful in the range above 5×10^{-10} mm Hg. The gauge is operated with an anode voltage of 6 kV and a magnetic field of 1000 gauss. The relationship between ion current and pressure is linear in the pressure range 10^{-4} to 5×10^{-10} mm Hg; at lower pressures the relationship $i_+ = cP^{1.7}$ obtains. In the linear region the ion current per unit pressure is given by i_+ (amperes) $\cong 9P$ (mm Hg) for nitrogen (i.e. a sensitivity about 45 times greater than obtainable with a thermionic triode gauge).

"The magnetron gauge: a cold-cathode vacuum gauge." P. A. Redhead. *Canadian Journal of Physics*, **37**, pp. 1260-1271, 1959.

VALVE OPERATING CHARACTERISTICS

The a.c. drive range of triodes and multi-grid valves should not be considered to be equal to the bias voltage difference between the operating point and the point where grid current starts. The complicated relationship between these quantities, also involving the impedances of the layer and intermediate layer on the cathode, can be illustrated approximately in a lucid form, which has been demonstrated in a recent German paper. The effect of the geometrical and design data on the drive range is determined by calculations. The rectifying effect can only be neglected in case of small a.c. voltages, while the drive range is increased in the case of larger a.c. voltages. Calculations and measurements are compared by using high-slope pentodes.

"The drive range of grid controlled valves." R. Wolfram. *Nachrichtentechnische Zeitschrift*, **12**, pp. 573-578, November 1959.

TRANSISTOR STABILIZATION

One of the most important parameters affecting temperature: to ensure optimum performance and to avoid thermal runaway some measure of the operation of transistor circuits is junction temperature stabilization must be included. An Australian paper has recently described some of the specific problems involved in the stabilization of class B amplifiers. In this case, normal methods of stabilization cannot be used to their fullest extent without excessive loss of signal power. The factors affecting the stability are discussed, with emphasis on those which are important and/or are peculiar to the class B circuit. Mention is also made of the methods of obtaining temperature stabilization, including non-linear compensation. Finally, experimental results are correlated with theory.

"Temperature stabilisation of transistors in class B amplifiers." K. L. Webber. *Proceedings of the Institution of Radio Engineers Australia*, **20**, pp. 726-733, December 1959.

AERIAL THEORY

A recent German paper investigates the influence on the effective area of a surface-type radiator of a deviation from the cophasal distribution. With the active radiator this deviation is proportional to the square of the ratio of the field strengths generated with non-cophasal and cophasal distributions; with the passive radiator (diverting reflector), however, it is proportional to this ratio itself. With the passive radiator the effective area can no longer be defined on the basis of the relationships holding for the active radiator, and it cannot therefore be thus calculated in the general case for an uneven surface, but from the transfer equation for the deflecting path.

"The calculation of the effective area of the non-cophasal distribution of a surface type radiator." G. F. Koch. *Archiv der Elektrischen Übertragung*, **13**, pp. 462-466, November 1959.

TRANSISTOR CIRCUIT TECHNIQUES

By the use of modified hybrid parameters the normal expressions for transistor amplifier characteristics can be extended to special cases where the circuit is modified by shunt, series, or common impedances in the circuit. Methods are given in a paper read at last year's Convention of the Australian I.R.E. for the calculation of such modified hybrid parameters, and of the approximate expressions which are applicable for the various transistor configurations. These expressions indicate methods of simplifying the design of transistor amplifiers, by reducing to within prescribed limits the effect of variations in the parameters. An extension of this method can be applied to two-stage amplifiers with single or multiple feedback loops.

"Transistor circuit design using modified hybrid parameters." R. E. Aitchison. *Proceedings of the Institution of Radio Engineers Australia,* **20**, pp. 673-679, November 1959.

TELEGRAPH DISTORTION EQUIPMENT

The economical and technical considerations have been outlined which led to an optimum design for a German f.m.-v.f. transistorized telegraphy system with a channel spacing of 120 c/s and a bandwidth of 80 c/s and an outlay comparable with a.m. systems. The problems of the choice of frequency deviation, transmission function and phase equalization and the circuit principle applied to this system are discussed. The improvements obtainable from keying speeds higher than 50 baud are illustrated by measured curves. Complete transistorization has led to a new design principle.

"A v.f.-telegraphy system with transistors for narrow band f.m." H. Heller. *Nachrichtentechnische Zeitschrift,* **12**, pp. 595-601, December 1959.

An electronic indicator for relative distortions of teleprinter signals has been developed by Siemens & Halske which gives an alarm when the distortion of these signals exceeds a preadjusted maximum value. The values can be selected in steps of 2 per cent. within the range ± 2 per cent. to ± 48 per cent. The measurement method is based on a comparison of the step onsets of the teleprinter signals with a time function adjusted according to one of the maximum values. This time function is derived from a crystal stabilized synchronizing pulse by a counting method with the aid of a binary counting chain. The measured values can be recorded in figures by means of an ancillary pulse counter.

"The teleprinter distortion indicator, an automatic test equipment." E. Schenk. *Nachrichtentechnische Zeitschrift,* **12**, pp. 609-612, December 1959.

STATISTICAL METHODS AND RELIABILITY

The increasing application of complicated electronic equipment emphasizes the question of reliability which is substantially determined by the life of components. The applicability of a simple mathematical law for the life of a component is investigated by means of a statistical model test in a German paper. Model tests and results from observations on valves lead to considerations according to which life data can be expressed by simple statistical aids.

"Mathematical and statistical methods of considering life data and their application to electron valves." A. Deixler and E. Rusch. *Nachrichtentechnische Zeitschrift,* **12**, pp. 613-618, December 1959.

V.H.F. MEASUREMENTS

For impedance measurements on balanced objects in the v.h.f. bands (80-300 Mc/s), a balanced, screened transmission line about 2 metres long has been designed in the Philips laboratory at Aachen. The characteristic impedance is approximately 105 ohms. The probe is insensitive to unsymmetrical waves, the detector diode (a miniature germanium diode OA95) being introduced directly into the r.f. field inside the line. The high-frequency supply voltage is modulated in amplitude at 1000 c/s. The total sensitivity is such that, at a high-frequency supply voltage of 5V and a standing-wave ratio of some hundreds, the voltage minima can still be accurately measured.

"A slotted lecher line for impedance measurements in the metric and decimetric wave bands." G. Schiefer. *Philips Technical Review,* **21**, pp. 88-91, 1959/60 (No. 3). (In English.)

CALCULATION OF AERIAL PATTERNS

In a paper from the National Research Council of Canada three antenna patterns are derived from the same linear aperture and may be described in terms of an angular spectrum of spatial frequencies ranging from zero to a common cutoff frequency. The band-passes according to the shape of the spectrum are rectangular, triangular, and cosinusoidal for the three patterns, and give resolving powers respectively of 1·33, 1·00, and 1·05, in terms of the cutoff period. The rectangular bandpass gives rise to the optimum antenna pattern and allows the Fourier components of a source from zero to cutoff frequency to be received with equal intensity and zero phase shift. Scanning curves of two equally intense point sources and a uniformly bright line are investigated.

"Resolving power of three antenna patterns derived from the same aperture." A. E. Covington and Gladys A. Harvey. *Canadian Journal of Physics,* **37**, pp. 1216-1229, 1959.