The Journal of

THE BRITISH INSTITUTION OF RADIO ENGINEERS

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

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IMMEDIATELY AFTERWARDS . . .

AS these notes are being written at the close of the 1959 Convention, it is too early to place on record considered reflections on its success. Judging by the many messages of appreciation which have already been received, however, it is believed that it was at least as successful as other post-war Conventions—perhaps even the most successful!

The Convention Handbook—which because of the printing dispute was nearly not produced at all!—was considered by every delegate to be a most useful publication. Indeed, many members asked if the notes of introduction in the booklet could be reproduced in the *Journal* as a record of historical fact. The Introduction stated that:

"When the British Institution of Radio Engineers was founded just over thirty-five years ago the prospect of visual communication was not then very tangible; indeed, audio communication between two given points without visible link was a comparatively new development of science and was still largely in the experimental stage. It was, in fact, still the era of the British Broadcasting *Company*.

"Nine years after its formation the Institution held the first Television Convention ever to take place in Great Britain, and two years afterwards the British Broadcasting *Corporation* started the world's first television service.

"It is impossible to pinpoint the stage at which the theory of television engineering was adapted or branched off into fields other than the transmission of what was at one time termed 'the radio picture.' At least it was British originality in the development and introduction of a television service which paved the way for the use of radar engineering in more urgent circumstances.

"Since that first Convention the Institution's meetings have achieved the reputation of being occasions for demonstrating and discussing new developments in radio science. Indeed, where the application has tended to extend beyond the field of oral, visual or recorded communication, the term 'electronics'—now popularly adopted—has crept in.

"The term 'electronics' was, of course, first mentioned in Great Britain in the Institution's post-war report. It was featured in the 1951 Convention in which industrial electronics and particularly electronic instrumentation in nucleonics formed a major part of the proceedings.

"The theme of this year's Convention also marries television engineering in the broadest sense with the application of television principles to wider industrial uses. How far it it is possible to use any application of science depends, to a very large extent, on disseminating existing knowledge and discussing that information with the requirements in other branches of science or industry. That is indeed one of the purposes of any scientific meeting, and the Convention may be regarded as a series of such meetings."

Once again the Convention programme featured the Clerk Maxwell Lecture. On the first evening the fourth of these Lectures was given by Dr. V. K. Zworykin to a capacity audience in the Clerk Maxwell Lecture Theatre and relayed over closed circuit television to delegates in adjoining theatres.

The Institution always takes pleasure in welcoming to its Conventions engineers and scientists from overseas. Eleven of the papers read were by authors from the United States, France, and the U.S.S.R. The four papers presented by the Russian delegates are believed to be the first accounts of television work in the Soviet Union to be presented before technical meetings in Great Britain.

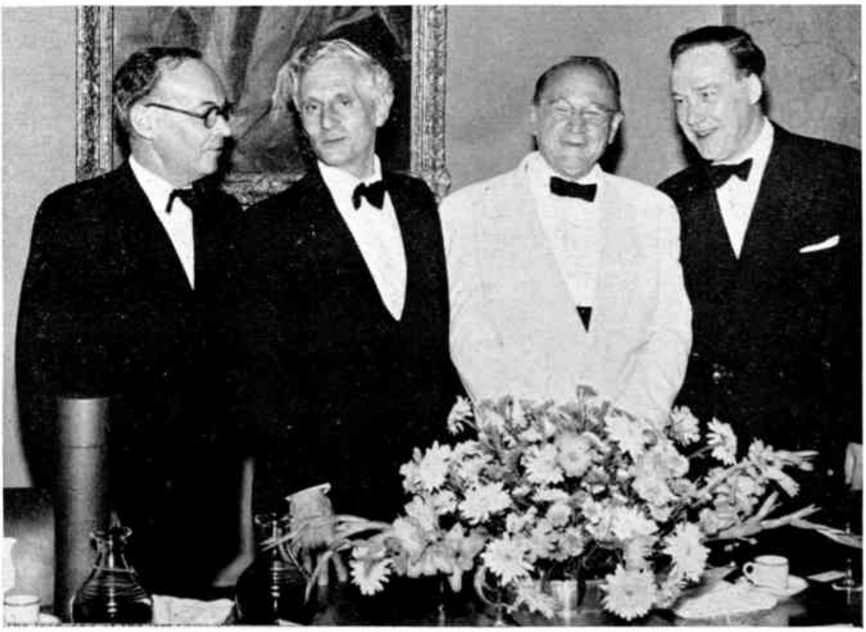
A particularly notable feature of the proceedings was the bestowal of Honorary Membership of the Institution on two men who have been prominent in the field of television engineering. Election to Honorary Membership is a rare occasion in the Institution's life: the ceremony which took place in Cambridge however, was unique in that for the first time ever two Honorary Members were elected in the same year, and the elections were confirmed in the Clerk Maxwell Lecture Theatre. The two new Honorary Members—Mr. Eric K. Cole, C.B.E., and Dr. Vladimir K. Zworykin—were warmly welcomed into membership of the Institution and received their Certificates after signing the Roll of Honorary Members at a Banquet held in their honour in Downing College. A photograph taken during this occasion is shown below, and a fuller account will be published later.

The addition of these two distinguished names to the Roll of Honorary Members brings the total of such elections to six since the end of the last war. The previous four were:

Leslie McMichael ... ... elected in 1949
H.R.H. Prince Philip, Duke of Edinburgh ... ... elected in 1951
Sir Noel Ashbridge ... ... elected in 1953
Professor G. W. O. Howe ... elected in 1956

Reference has already been made to the difficulties encountered in the production of the Convention Handbook. The preparation of preprints was similarly affected. Nevertheless, 36 of the 51 papers presented were available in complete or nearly complete form at the start of the Convention—an achievement by the Institution and its printers which greatly helped the running of the Convention and led to many favourable comments by delegates.

G.D.C.



After the signing of the Roll of Honorary Members.
Left to right: Mr. Eric K. Cole, C.B.E., The President, Professor E. E. Zepler, Dr. Vladimir K. Zworykin and Mr. Graham D. Clifford (General Secretary).

1959 Convention Diary

Tuesday, 30th June

The work which the Institution's staff had previously done in re-arranging part of the "D.C. Lab." at the Cavendish Laboratory as a Convention Office and Reception Lounge was completed just in time to accommodate the first delegates who registered on Tuesday evening.

A display of Institution publications, and tables and comfortable chairs were set up in the Reception Lounge. The remainder of the laboratory was given over to demonstrations of colour television, and black and white television relays from the lecture theatre, as well as for refreshments during the morning and afternoon.

Wednesday, 1st July

Each delegate received a copy of the first issue of "Brit.I.R.E. Convention News" which contained the latest information on additions to the programme and lists of delegates attending. It was announced on the first day that over 300 delegates had registered.

Well before 2 p.m. the Clerk Maxwell Lecture Theatre was filled and the overflow theatre was being pressed into service as the President, Professor E. E. Zepler, greeted members and delegates and formally opened the Institution's post-war Convention.

In his opening remarks the President said:

"Such is the increasing scope and application of radio and electronic science, that I sometimes wonder whether we could exhaust Convention material even if we had a Convention every six months! It was not easy, therefore, to decide on what branch of our profession we should concentrate this year's Convention. In the event the number of papers that have been submitted and, what is more important, the registrations for attendance at the Convention, have proved that our choice has at least met with general approval.

"It is, I think, of equal importance that we should make publicly known the tremendous amount of work that is going on in this field of television engineering, which is by no means confined to the broadcasting in the entertainment sense. I will not catalogue here these applications, but I am sure that the presentation of papers and the ensuing discussions will help to make more widely known the increasing scope and responsibility of the work of the radio and electronics engineer.

"The arrangements for a Convention like this are at times difficult and this year they have been complicated by such matters as the printing strike. I would therefore like to take this very early opportunity of congratulating the Convention Committee who have been responsible for the arrangements, and to thank in particular the Chairman of the Committee, Mr. Cooper, for all the work that he has done for the Institution."

All the opening day's papers were mainly general surveys of television engineering systems, but immediately led to a most useful discussion.

During the evening Dr. Vladimir K. Zworykin gave the Clerk Maxwell Memorial Lecture to an audience which packed the lecture theatre as well as two overflow theatres. (The lecture will be published in the September *Journal*.)

The first day's activities culminated in a reception in Downing College arranged by the Directors of the Pye Group. By courtesy of the Master of Downing, who attended the reception, guests were entertained in Hall, as well as in a large marquee erected on the lawn. The customary background music for such an occasion was provided by stereophonic equipment reproducing quieter and perhaps more appropriate musical entertainment than that normally given in such demonstrations!

Mr. and Mrs. C. O. Stanley and Officers of the Institution received the guests who included the Mayor and Mayoress of Cambridge, Councillor and Mrs. Wallace Cole.

This function provided an opportunity for delegates to meet, renew old acquaintances and make new friends in congenial surroundings.

The remaining days of the 1959 Convention provided much material for the "Brit.I.R.E. Convention News" and further extracts will be given next month.

INSTITUTION NOTICES

Institution Premiums for 1958

The Council of the Institution announces that the following awards are to be made for outstanding papers published in the *Journal* during 1958:—

Clerk Maxwell Premium:

To C. Powell and D. A. Hendley (Associate Members).

"Dectra: A Long-range Radio-navigation Aid," (May 1958).

Heinrich Hertz Premium:

To K. Foster, M.A.

"The Characteristic Impedance and Phase Velocity of High-Q Triplate Line," (December 1958).

Louis Sterling Premium:

To A. van Weel, Dr. Techn.Sc.

"Design of Detector Stages for Signals with Symmetrical or Asymmetrical Sidebands," (September 1958). (*This is the third successive year that Dr. van Weel has been awarded this Premium*)

J. C. Bose Premium:

To B. Ramachandra Rao, M.Sc., M. Srirama Rao, M.Sc., D.Sc., and C. Abhirama Reddy, M.Sc.

"Magneto-Ionic Fading in Pulsed Radio Waves Reflected at Vertical Incidence from the Ionosphere," (November 1958).

Brabazon Premium:

To Professor D. G. Tucker, D.Sc., V. G. Welsby, Ph.D. (Members), R. Kendell, M.Sc., and D. E. N. Davies, M.Sc.

"Electronic Sector Scanning," (August 1958) and "Radar Systems with Electronic Sector Scanning," (December 1958).

Marconi Premium:

To M. B. Prince, Ph.D., B.Sc. (Associate Member) and M. Wolf, Dipl. Phys.

"New Developments in Silicon Photovoltaic Devices," (October 1958).

These Premiums will be presented by the President at the Annual General Meeting in London on Wednesday, 2nd December next.

There are altogether thirteen Institution Premiums which may be awarded annually. Four of these were established by the Council at the end of last year. Papers published in the *Journal* during the current year, and falling within the appropriate terms of reference, will be eligible for these new Premiums. Details were published in the March *Journal*.

The President to visit America

The President of the Institution is visiting America in August. He is due to arrive in New York on 26th August and will be staying in the U.S.A. for about five weeks.

Professor E. E. Zepler will be very pleased to meet members of the Institution in America and may be contacted at 7501 Ridge Boulevard, Brooklyn, New York (Telephone number: SHore Road 5-6574).

F.C.G.I.

The Council of the City and Guilds of London Institute has conferred Fellowship of the Institute (F.C.G.I.) upon Mr. Paul Adorian (Past President of the Institution).

The award is cited as being in recognition for his pioneer work in relay broadcasting and for distinguished services to the profession.

In addition to a number of other appointments, Mr. Adorian is also managing director of Associated-Rediffusion Ltd.

New Appointment

Another Past President of the Institution, Mr. William Edward Miller, has recently been appointed managing director of the Trader Publishing Co. Ltd., a company associated with Iliffe & Sons, Ltd.

Corrections

The following correction should be made in the Medical Electronics Group report on "Constant Voltage Electro-diagnostic Stimulators" published in April 1959.

Page 247: the denominator of eqn. (4) *should read*

$$(R_o + R_s + R_p)\,(R_o + R_s)$$

The following corrections should be made in the paper "Multiplicative Receiving Arrays" which was published in the June issue.

Page 373: Equation (9) *should be renumbered* equation (7).

Page 375: In equation (9) which defines the directivity factor, $1/2\pi$ *should replace* $\pi/2$ in the denominator.

Transmitting Aerials for Television Broadcasting in the United Kingdom†

by

A. BROWN, ASSOCIATE MEMBER‡

A paper read at the Television Engineering Convention in Cambridge on 4th July 1959.

Summary: The development of transmitting aerials for television broadcasting in Band I is surveyed. Reference is made to the increase of gain, and to the splitting of the aerial system for increased flexibility and reliability. Aerials having directional radiation characteristics have been required for filling gaps in the service areas of other stations, and giving adequate protection against interference with other services. Measurement techniques and the permissible variation of reflection coefficient and radiation pattern over the video band are considered. Future developments discussed include the use of higher frequencies to obtain higher gains. The problem of giving an adequate service in areas near the mast will thus be increased and the maximum possible useful gain which can be obtained from an aerial of given height, and the degree to which a practical aerial can approach this condition, are discussed. Measurements on scale models are discussed.

1. Introduction

In 1946 the B.B.C. public television service was re-commenced, after having been closed down for the period of the war. Since this date a continuous steady expansion has been taking place, and the Independent Television Authority has been established, until at the present time it is estimated that approximately 98 per cent. of the population of the United Kingdom is within an area served by at least one television programme. It may be of interest to record some of the developments in the design of transmitting aerial systems which have taken place during the past 13 years, with particular reference to Band I (41-68 Mc/s), and to attempt to predict possible future developments.

The first phase of the post-war expansion was the construction of four high-power transmitting stations. These required a vertically-polarized aerial, having an omni-directional radiation pattern in the horizontal plane. At each station an essentially similar aerial design[1] was employed. The next phase in the development programme was the provision of a number of medium- and low-power stations to fill in the main gaps in the service coverage. In order to minimize co-channel interference with existing stations, horizontal polarization was used in several cases, necessitating a new aerial design. At some stations, directional radiation patterns were required. Another factor which became of increasing importance was the possibility of another aerial being required on the same mast to work on one of the higher frequency bands. In the event of such an aerial being required, it would be desirable to place it as high on the mast as possible. This meant that the Band I aerial would have to be mounted lower down the mast, and hence the cross-section of the supporting structure would be much greater than on the earlier models. As a result of these changing requirements, a number of different aerial designs was evolved.

† Manuscript received 2nd April 1959. (Paper No. 507.)

‡ British Broadcasting Corporation, Research Department, Kingswood Warren, Surrey.

U.D.C. No. 621.396.677: 621.397.61

2. Design Techniques

2.1. *Factors influencing the specification of pattern and gain*

The specification of the transmitting aerial is inevitably a compromise between various, sometimes conflicting, requirements. The radiation pattern is determined by the shape of the required service area, due consideration being given to the avoidance of interference with other stations working on the same channel. The gain is normally limited by the aperture† which is available on the mast. Any increase in this aperture is normally accompanied by an increase in mechanical stresses, thus leading to a larger, and hence more expensive, mast. It is often the case that more than one aerial has to use the same supporting mast, and the resultant design must be a compromise between the requirements of each aerial.

2.2. *Impedance characteristics*

In deciding the required specification of the permissible reflection coefficient of a transmitting aerial, a number of factors must be considered. These are:

(1) The loss of radiated power.
(2) Distortion of the primary signal.
(3) Increased current and voltage stresses on the feeder system.
(4) Possible effects upon the radiation pattern of the aerial.
(5) Possible effects upon combining filters, etc.
(6) The production of ghost images.

Considering (1), Fig. 1 shows the variation of power loss with reflection coefficient. It may be seen that if we arbitrarily allow a loss of 0·2 db due to mismatching, the reflection coefficient can be permitted to be as great as 21 per cent. at the carrier frequency. It will be seen that other factors require a more stringent specification than this.

Factor (2) may be important if the reflection coefficient varies over the video band. Referring again to Fig. 1 it may be seen that if a good match is obtained at the carrier frequency, the amplitude response over the video band will not vary by more than 0·2 db if the reflection coefficient does not exceed 21 per cent. at any frequency.

Factor (3) is not normally a limitation, as adequate safety factors are allowed on the ratings of the feeders. The normal limitation upon the power-handling capacity of the feeders is heating of the inner conductor rather than flashovers due to high voltage. Owing to the high thermal conductivity of copper the effect of increased currents at various points due to standing waves tends to be spread evenly along the feeder.

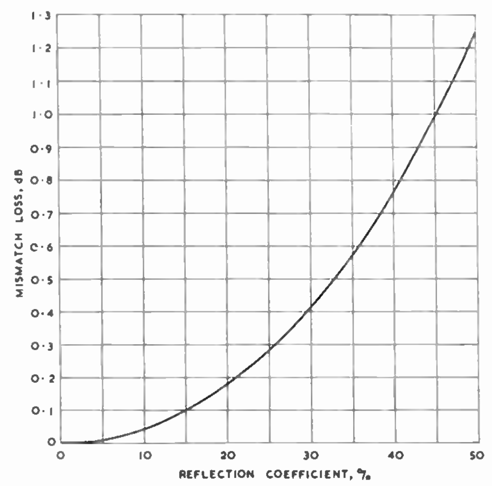


Fig. 1. Curve showing variation of mismatch loss with reflection coefficient.

Factor (4) depends upon the particular design of the aerial system. The effect may be important in a system using phase rotation of the aerial feeds, but it is impossible to state any general conclusions. Each aerial design has to be considered on its own merits in this respect.

It is difficult to make a general statement regarding factor (5), as the effect depends upon the design of the filter. However, it may generally be said that if the reflection coefficient is sufficiently low to meet the other specifications, no difficulties are obtained with the combining filters in this respect.

Factor (6) is the one which imposes the most stringent requirements upon the reflection coefficient. If a reflection takes places at the aerial terminals, the reflected wave may travel back down the feeder line to the transmitter, where it will in general be again reflected, and will be re-radiated at the aerial after suffering a time delay produced by the double traversal

† In this paper, the term "aperture" means the overall vertical length of the aerial system, expressed in terms of wavelengths.

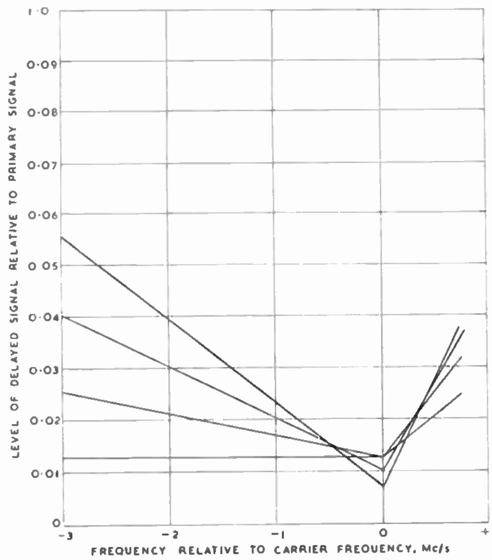


Fig. 2. Maximum level of radiated delayed signal for imperceptible distortion. If the relative amplitude of the delayed image is plotted against frequency, the curve must be wholly below one of the four lines shown.

of the feeder system. A typical time delay is 2 microseconds.

In order to find the maximum permissible reflection coefficient, a series of subjective tests was carried out. In these tests, a signal delayed by 2 microsec was deliberately introduced, and its amplitude relative to the primary signal was adjustable. The video signal employed was an electronically-generated test pattern, and also a series of typical pictures obtained from telecine equipment. A high-quality picture monitor was used for the tests. In general, the visibility of the ghost image is dependent upon the manner in which the amplitude of the image signal varies over the video band, and to take account on this fact the delayed image in the tests was passed through three types of distorting networks, giving delayed-image distortion corresponding to typical aerial impedance characteristics. The results of these tests are summarized in Fig. 2. Referring to this figure, if the relative amplitude of the delayed image is plotted against frequency, the curve must be below one of the four lines shown, in order that the distortion due to the ghost image is imperceptible. It may be seen that a small reduction in the level of the delayed image at the carrier frequency permits a considerable increase in the level towards the extremes of the video band. This fact is of considerable importance in specifying the reflection coefficient of the aerial, as it is usually much simpler to achieve an impedance characteristic in which the reflection coefficient is allowed to increase towards the ends of the band, rather than to achieve a low reflection coefficient over the whole band. In deriving the required specification of reflection coefficient from Fig. 2, allowance may be made for the attenuation undergone by the reflected signal in the feeder and in vestigial and combining filters and the loss due to incomplete reflection at the transmitter. All these factors help to permit some relaxation of the specification, but the amount depends upon the particular case involved. However, considering all these factors, a specification has been drawn up for a typical case, as shown in Fig. 3. From this figure it may be seen that if the reflection coefficient is approximately 1·8 per cent. at the vision carrier frequency, this value must be maintained over the whole band, but if the reflection coefficient can

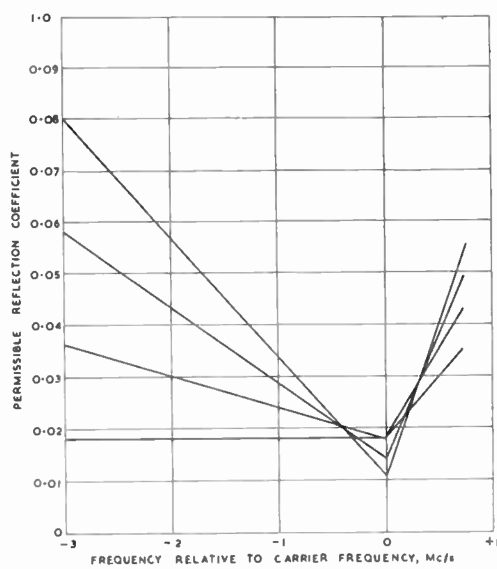


Fig. 3. Alternative characteristics for the maximum reflection coefficient for the aerial at a typical television station. If the reflection coefficient is plotted against frequency, the curve must be wholly below one of the four lines shown.

be reduced to 1 per cent. at the carrier frequency, it can be permitted to rise to 8 per cent. at the end of the band.

2.3. *Factors influencing the mechanical design*

The physical form of the aerial is determined by the polarization used, and also by the size and shape of the supporting structure. The first post-war design[1] consisted of two tiers of folded dipoles mounted on a square support pole of 1·5 ft. side. The assembly was fixed at the top of the cylindrical structure forming the slot aerial for the sound transmission in Band II (88-100 Mc/s). This design of aerial proved very satisfactory in service, but at later stations differing requirements precluded its use.

At two stations, an omni-directional horizontally-polarized aerial was required. In this case, the well-known " Super-turnstile " aerial was employed.[2] Fundamentally, this consists of two crossed horizontal elements fed in phase quadrature. Each element can be regarded as a stack of dipoles of varying lengths, connected in parallel. Alternatively, it can be considered as a slot in a finite plane. The resultant structure has a very large bandwidth, and is therefore very suitable for a television transmitting aerial.

At this stage in the development programme, the possibility of requiring another aerial on the mast to work on one of the higher-frequency bands became of increasing importance, with the result that the later Band I aerials had to be placed on a supporting structure having a larger cross-section than in the case of the earlier aerials. In a typical case, the mast may be of 4 ft square cross-section. This, in Band I, is of the order of $0{\cdot}2\lambda$ square, and the " Super-turnstile " type of aerial is no longer suitable for producing the required radiation pattern. It is also interesting to note that under this condition a uniform horizontal radiation pattern can usually be more easily achieved by using a ring of elements around the mast fed in phase, rather than with a progressive phase rotation. The advantages of using a phase rotation of aerial feeds are in any case not great. One advantage is that a very good overall impedance characteristic can be obtained, but against this must be weighed the fact that the shape of the horizontal radiation pattern is dependent upon the degree of matching achieved for each element.

When the Super-turnstile type of aerial was superseded, advantage could still be taken of

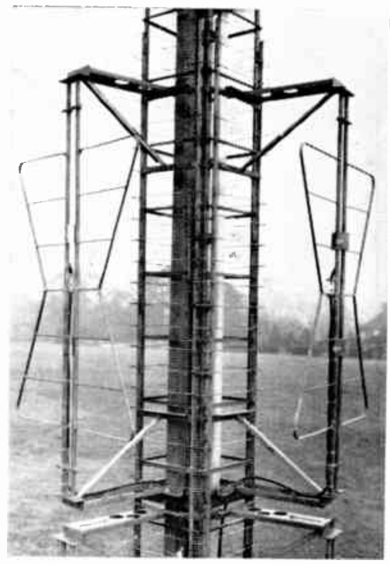


Fig. 4. Photograph of model of Band I aerial using tangential batwing elements.

the good impedance characteristics of the " batwing " type of radiating element which is used with the Super-turnstile, and several aerials were designed in which these elements were used instead of dipoles mounted on a mast of large cross-section.[3] Figure 4 shows a model of one such arrangement, in which the batwing element is mounted tangentially to the mast. Generally, however, it was found that these batwing elements proved to be rather cumbersome to install, whilst the windage is greater than an equivalent dipole system. On later aerials, therefore, wide-band folded dipoles were used as the radiating elements, reactance compensation being applied as necessary to achieve a suitable impedance characteristic.

2.4. *Split aerial systems*

At the earlier high-power stations, the total power output was obtained from one transmitter; in the event of failure of this transmitter, a reserve transmitter of lower power was brought into operation. More recently this practice has given way to an alternative scheme, in which two transmitters of equal power each feed into half the aerial system. Thus, if one transmitter should fail, a drop in signal strength of approximately 6 db will occur. It is usually arranged that, at a convenient time, the one remaining transmitter can be switched to the two halves of the aerial connected in parallel, so that the loss in signal strength can be reduced to 3 db. This practice requires two independent feeders up the mast, but the arrangement provides great flexibility and reliability. In designing such a system, however, there is a possible difficulty which has to be considered. If the receiving site is in a null of the vertical radiation pattern of the aerial, there is the possibility that different relative phase shifts through the two transmitters over the video band may cause a deterioration in the picture quality, and this effect may vary with time. However, this effect will be observed only in small areas. Under emergency conditions, when only one half of the aerial is in use, it is possible that "ghost" images will be produced. These may arise in two ways. In general, there will be mutual coupling between the two halves of the aerial system. If one half is de-energized, a signal will be induced from the half still radiating; this signal may travel down the feeder line, be reflected at the transmitter, and so be re-radiated as a ghost image. In addition to this effect, the impedance of the driven half of the aerial will in general change when the other half is de-energized, and the reflection coefficient may become great enough for a ghost image to be produced. However, this condition only occurs under emergency conditions; the design is normally such that the effect is very small, and it has not been found necessary to fit filters to absorb the reflected signals.

2.5. *Gain*

It is well known that the gain of a practical aerial is very closely connected with the available aperture,[4] and in the case of an aerial having an omni-directional radiation pattern in the horizontal plane, an increase in gain can usually be obtained only by vertical stacking of the elements. The gain of one tier of the aerial system could be increased by a horizontal extension of the aerial (e.g. a ring of Yagi-type aerials), but in practice this is precluded for mechanical reasons. The gain can also be increased by using a ring of elements, fed in a progressive phase rotation around the ring. As an illustration of this, a ring of four vertical dipoles fed in phase rotation and spaced on a circle of diameter $0{\cdot}3\lambda$ has a gain of approximately 1 db relative to a single vertical dipole. It has already been mentioned, however, that in some cases this type of aerial may not produce a satisfactory radiation pattern in the horizontal plane.

If more than one tier has in any case to be employed, the advantage of having an inherent power gain in each tier is lessened. For an aerial system having a cylindrical symmetry, consisting of a number of equi-spaced tiers fed in phase, the overall gain is proportional to

$$1 \Bigg/ \int_0^{\pi/2} \Big[F(\theta) \Big]^2 \left[\frac{\sin (n\pi s \sin \theta)}{n \sin (\pi s \sin \theta)} \right]^2 \cos \theta \, d\theta,$$

where θ is the angle from the normal to the axis of the aerial, $F(\theta)$ is the vertical radiation pattern of the individual tier, n is the number of tiers and $s\lambda$ is the spacing between tiers. As n increases, $\dfrac{\sin (n\pi s \sin \theta)}{n \sin (\pi s \sin \theta)}$ changes rapidly with θ, and unless the vertical pattern of each tier is very sharp, the term $F(\theta)$ becomes of less importance, i.e. the gain of the whole aerial tends to become independent of the gain of individual tiers. One consequence of having a power gain in each tier is that the optimum inter-tier spacing is increased This may be illustrated graphically as follows

Referring to Fig. 5, the vertical radiation pattern of a two-tier aerial, each tier consisting of an isotropic radiator, is plotted on a special scale. This scale is arranged so that the power gain of the aerial relative to an isotropic source is given by the ratio of the total area to the area under the plotted curve. Curves are plotted for inter-tier spacings of $0{\cdot}5\lambda$ to $1{\cdot}0\lambda$. By inspecting these curves, it is clear that the optimum spacing, (i.e. the spacing producing the minimum area under the curve) lies between $0{\cdot}7\lambda$ and $0{\cdot}8\lambda$. The actual optimum spacing has in fact been calculated to be $0{\cdot}72\lambda$

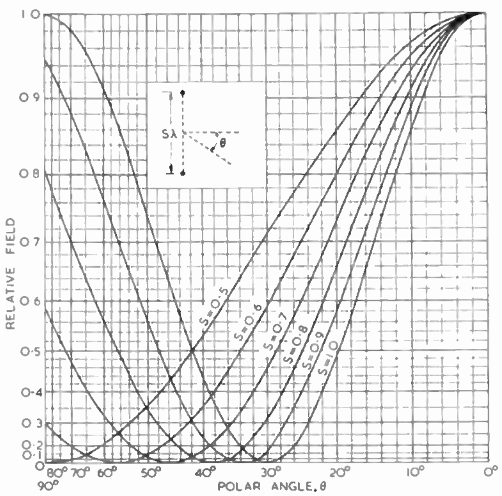


Fig. 5. Radiation pattern of two tiers of isotropic sources.

for this case. Supposing, however, each tier has a vertical radiation pattern such that the radiation decreases with increasing polar angle. Then, the lobe at the left-hand side of the minimum becomes reduced, and the optimum inter-tier spacing increases. In the limiting case, we may say that the resultant lobe on the left-hand side of the minimum becomes zero, and the power gain is determined completely by the area of the lobe on the right-hand side of the minimum. This decreases as s increases, and as the vertical directivity of the individual tier increases to infinity, the optimum inter-tier spacing tends to infinity. This is, of course, a consequence of the fact that the power gain obtainable from a linear aerial is governed very largely by the physical length of the aerial.

Up to the present, the gains obtained from omni-directional aerials in Band I have been between 3·7 db and 8·0 db. In Band III, gains up to 13·8 db have been achieved.

3. Measuring Techniques

3.1. *Impedance*

For impedance measurements a v.h.f. bridge is used, in conjunction with a taper for connection to the large coaxial feeders. Corrections can be applied for discontinuities in the taper. Greater accuracy of measurement can, however, be obtained by measuring the impedance at the end of the feeder line remote from the aerial. By making measurements at a number of frequencies in the band, the reflection coefficient at the aerial terminals can be separated from that due to discontinuities at the bridge and taper terminals.

3.2. *Radiation patterns*

The radiation pattern is measured using scale models, working at an appropriately higher frequency. It has been found that these measurements give good agreement with the results obtained from measurements of the radiated field of the final aerial.

In the case in which the aerial design requires elements being fed with currents having different amplitudes and phases, the radiation pattern of one element mounted on the mast is measured in amplitude and in phase, and it is then a relatively straightforward matter to calculate the resultant radiation pattern when a number of such elements are used, fed with various amplitudes and phases of currents. This system avoids the difficulties of having to ensure that the correct amplitudes and phases are achieved on the scale model.

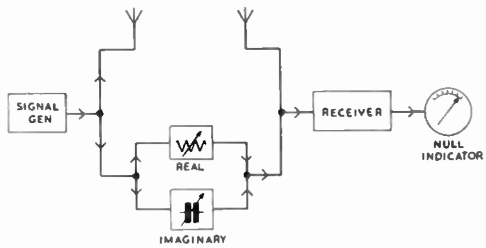


Fig. 6. Arrangement used for the measurement of radiation patterns in amplitude and phase.

The principle of the amplitude and phase measurement is illustrated in Fig. 6. The output from an oscillator is fed to a receiver via two paths—one of which consists of the aerial under test, together with the associated transmission path and receiving aerial, and the other is through the measuring instrument. This measuring instrument is a device which produces two components in phase quadrature, each component being adjustable in amplitude. The two components are adjusted to achieve complete cancellation of the signal through the path containing the aerial under test.

4. Overcrowding Problems

4.1. *National and International Protection Problems*

In order to produce as complete a service coverage as possible, it has been necessary to erect a large number of stations of varying powers. The initial B.B.C. plan was to provide five high-power stations and five medium-power stations in Band I, supplemented by stations in Band III (176-216 Mc/s). A number of channels in Band III have been allocated to the Independent Television Authority, but none has so far been made available to the B.B.C. The B.B.C. has therefore been faced with the difficult problem of achieving national coverage using Band I alone. Considerable ingenuity has been necessary to permit a number of additional low-power stations to be fitted into Band I without causing serious co-channel interference with the other stations using the same channels. There are already twenty-two stations in Band I, and further low-power stations are planned.

International co-operation is necessary in order to avoid excessive interference between stations in neighbouring European countries, and this work is co-ordinated through the International Telecommunications Union.[5] The channels allocated under the original B.B.C. plan were in conformity with the Stockholm Agreement of 1952. The later stations not provided for in the Agreement conform with internationally agreed standards of protection, and problems of mutual interest are discussed with the Administrations concerned.

4.2. *Cross-polarization*

One method of reducing interference between two stations working on the same frequency is to use different polarizations. Measurements have shown that in Band I a reduction in interference level of at least 10 db is obtained by using horizontal polarization for one station, and vertical polarization for the other station. It is possible that the use of circular polarization (using opposite directions of rotation for the two stations) might be beneficial, but this would lead to complicated aerial designs, and as far as the author knows, no measurements have been made in this country to assess the degree of discrimination which could be obtained in practice.

4.3. *Directional radiation patterns*

Additional protection against co-channel interference can be obtained by using aerials with directional radiation patterns at the receiving and transmitting sites. Up to the present, in planning the service coverage in Band I, it has been assumed that no protection is provided at the receiving site. This is in accord with the general policy of reducing the complications at the receiving end as much as possible. At higher frequencies, it would probably be reasonable to assume that some protection is provided by the receiving aerial, because at these frequencies, directional aerials are in any case more widely used for other reasons.

At the transmitting aerial, reduction of co-channel interference can be obtained by restricting the radiation in certain specified directions. As the number of transmitting stations increases, more such restrictions have to be placed upon the radiation pattern of the aerial. Figure 7 shows one example of the specification for an aerial for a low-power station. In this case, for instance, the required specifica-

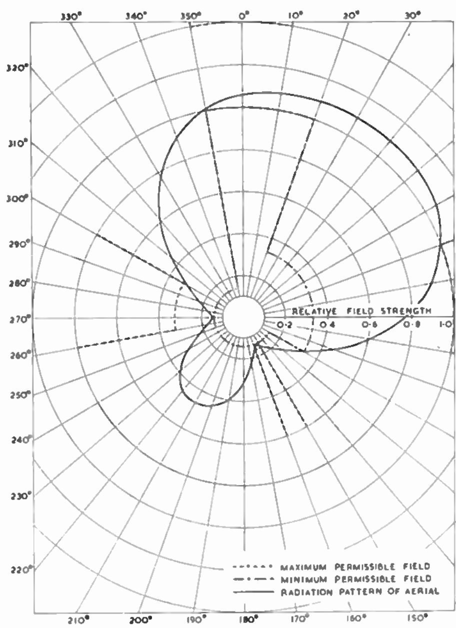


Fig. 7. Diagram illustrating a typical specification for a radiation pattern having restrictions placed upon the maximum radiation in certain directions.

tion could be met by the use of six dipoles per tier, carrying suitable amplitudes and phases of currents.

The calculation of the required distribution of radiating elements has been largely carried out by somewhat tedious "trial and error" methods. The calculation of the radiation pattern of a given system of dipoles mounted on a supporting mast is a relatively straightforward matter,[6, 7] but the converse problem of calculating the simplest arrangement required in order to produce a given radiation pattern is more difficult.

5. Future Developments

It is difficult to foresee exactly future developments, as these will no doubt depend to some extent upon factors which are not merely of an engineering nature. Perhaps the most obvious development will be the use of higher frequencies, in Bands IV and V. At these higher frequencies, higher gains will certainly be required. No great difficulty is foreseen in this connection, since the aerials will no doubt have a greater electrical aperture available than is available for the present lower-frequency aerials. However, increasing the gain merely by increasing the aperture of the aerial will produce a large number of nulls in the vertical radiation pattern, some of these nulls occurring at a considerable distance from the transmitting site. These nulls will, in general, have to be " filled in ", in order to ensure that an adequate signal is obtained throughout the service area.

It should be noted that in an area near a null of the vertical radiation pattern, a bad picture can be received even when the signal strength appears to be adequate. This effect can be caused by considerable variations in the phase of the received signal over the video band. Figure 8 shows a photograph of the type of picture which was obtained by simulating conditions which could exist at such a point. It can be seen that the high video frequencies are heavily accentuated, and the grey scale is badly distorted.

One result of having a larger electrical aperture available is that a greater measure of control is possible over the shape of the vertical radiation pattern. The filling-in of the nulls of the pattern could be regarded as a form of such deliberate shaping of the pattern. It may be of some interest to consider to what degree this process might be extended in the event of Bands IV and V being used.

As a starting point, the optimum shape required for the vertical pattern of a broadcasting aerial will be considered. This will, of course, depend in a particular case upon circumstances such as the type of terrain, location of heavily populated areas, etc. For a preliminary survey, however, it is reasonable to assume perfectly-conducting ground, and that the service area requires equal signal strength at each point. At present, of course, this condition is not achieved, the signal strength becoming weak at the fringe of the service area, and being unnecessarily great at relatively short distances from the transmitting site.

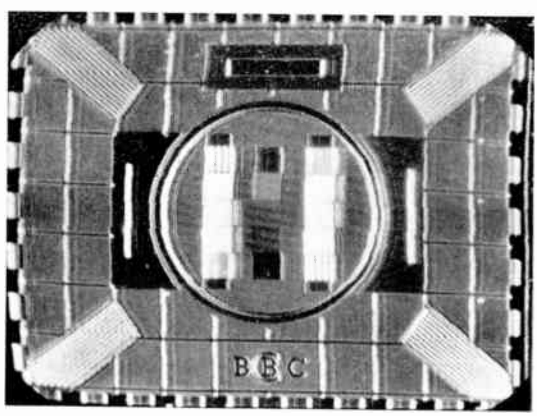


Fig. 8. Photograph illustrating the distortion which may be obtained at a site in the null of the radiation pattern of an aerial.

For the purpose of this discussion, it is necessary to decide upon some criterion with which to define the service area. It seems reasonable to assume that the service should be restricted to the area covered by line-of-sight transmission. Thus, referring to Fig. 9, it is required to produce an equal signal strength from T to P. Assuming that the earth is a perfect conductor, and that $h_1 + h_2 \ll d$ (this implies that the angle α is small), then the relative signal strength received at a point S from an isotropic radiator at point T′ varies as

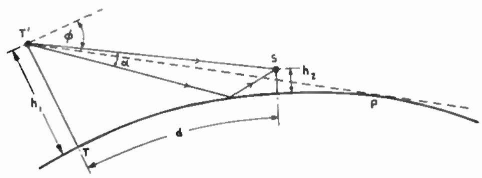


Fig. 9. Calculation of service area of line-of-sight transmission.

shown in Fig. 10. In deriving this curve, it is assumed that the effective radius of the earth is 10,000 km, and that the ratio $h_1/h_2 = 25$. The wavelength is assumed to be 0·5 m and the height of the transmitting aerial is 229 m (i.e. 750 ft). It may be seen that for a fixed height of receiving aerial, the field strength fluctuates cyclically to a distance of about 12 km. Taking a mean value of the fluctuation, the field decreases linearly with distance up to 12 km, and thereafter it decreases more rapidly, finally tending to decrease according to the square of the distance. It may be noted that the onset of the square-law characteristic takes place at a much greater distance from the transmitter than at lower frequencies. This is because the resultant field strength is the difference of two vectors at an angle η, where $\eta \cong \dfrac{4\pi\, h_1 h_2}{\lambda d}$ (h_1 and h_2 being the height of the transmitting and receiving aerials respectively). The position of the final maximum on the field strength is given by $\dfrac{4\pi\, h_1 h_2}{\lambda d} = \pi$, i.e. $d = \dfrac{4\, h_1 h_2}{\lambda}$, so that as λ is reduced, the distance of the last maximum is pushed further out from the transmitter.

Measured values of field strengths from the experimental Band V transmitter at Crystal Palace indicate that in fact the mean value of field strength decreases with the square of the distance even relatively close to the transmitter (apart from effects due to the vertical radiation pattern of the transmitting aerial).

If it is assumed that the field strength from an isotropic source decreases linearly with distance, then in order to equalize the field strength over the whole service area, it may be shown that the transmitting aerial should have a vertical radiation pattern which varies as cosec θ, where θ is the angle with respect to the normal to the aerial. A $\text{cosec}^2\theta$ pattern is required, however, if the field from an isotropic source decreases as the square of the distance. It should be noted that ideally the direction of the main lobe of the radiation pattern should be tilted down towards the horizon by the angle $\varnothing$ (in Fig. 9). It is easily shown that $\varnothing = \sqrt{(2h_1/R)}$, where h_1 is the height of the transmitting aerial, and R is the effective radius of the earth. One result of this is that there is a definite theoretical limit upon the maximum power gain which can usefully be obtained from an aerial of given height, even apart from practical considerations. Thus, for an aerial having a cosec θ pattern, for positive angles greater than $\varnothing$, and zero radiation for smaller positive and all negative angles, the power gain, relative to an isotropic source is given by

$$\text{Gain} = 2 \text{ cosec} \sqrt{\frac{2h_1}{R}} \; .$$

For an aerial having a $\text{cosec}^2\theta$ pattern, the gain becomes

$$6 \text{ cosec} \sqrt{\frac{2h_1}{R}}$$

It is unlikely that these idealized radiation patterns would be obtained in practice, but they form a useful standard with which to compare the performance of a practical aerial.

It is instructive to consider the degree to which a practical aerial may approach the ideal. Theoretically, it is possible to obtain

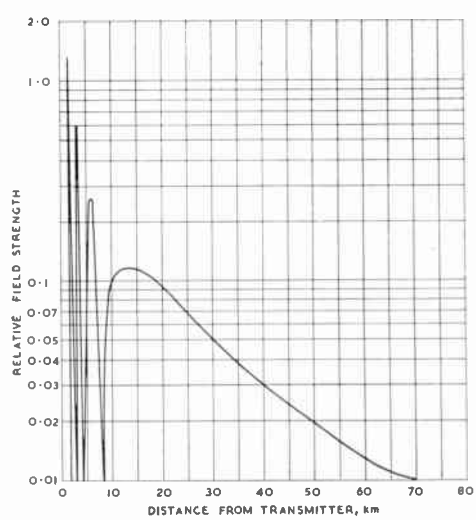


Fig. 10. Variation of field strength with increasing distance from transmitter.

any required radiation pattern from a linear aerial of finite size, but practical limitations preclude any appreciable increase of gain over that obtained from an equi-amplitude co-phased aperture. It therefore seems reasonable to use such an aperture as the basis of design, and then to make adjustments as required to improve the shape of the pattern. In discussing this problem, it will be assumed that there is a continuous distribution of energy across the aperture.

In "building up" the aperture distribution required to produce the necessary radiation pattern, it is useful to use the principle of superposition, i.e., if two aperture distributions are added, the resultant radiation pattern is the complex sum of the radiation patterns of each distribution taken separately. Taking as our basis a linear, equi-amplitude co-phased aperture, the radiation pattern is given by

$$F(\theta) = \cos\theta \frac{\sin\left(\frac{\pi a}{\lambda} . \sin\theta\right)}{(\sin\theta)\left(\frac{\pi a}{\lambda}\right)}$$

where θ = angle with respect to the normal to the aperture

a = length of the aperture.

This expression is "normalized" so that the field is unity at $\theta = 0$. This pattern is of course symmetrical about $\theta = 0$, and the positions of the nulls are given by $\sin\theta = n\lambda/a$, where n = 1, 2, 3 .... In order to "fill in" these nulls, it is therefore desirable to superimpose another distribution having a radiation pattern with maxima at $\sin\theta = n\lambda/a$. It may also be noted that the complex value of the "basic" co-phased radiation pattern is purely real, and that it alternates between positive and negative. It follows from this that if it is desired to have no nulls in the final radiation pattern, the superimposed pattern should be "imaginary" i.e. in quadrature with the basic pattern. In investigating possible types of distribution to achieve this, we may note that the required "null-filling" pattern has a shape corresponding to the rate of change of the basic pattern, and this suggests that a "triangular" distribution is required, i.e., a distribution in which the amplitude steadily increases along the aperture, with a phase reversal at the centre of the aperture. Such a distribution has a radiation pattern $F(\theta)$ given by

$$\frac{\frac{\pi a}{\lambda} . \sin\theta.\cos\left[\frac{\pi a}{\lambda}\sin\theta\right] - \sin\left[\frac{\pi a}{\lambda}\sin\theta\right]}{\left[\frac{\pi a}{\lambda}\sin\theta\right]^2} . \frac{a^2}{2}$$

Finally, it is desirable to tilt the main lobe by the angle $\sqrt{(2h_1/R)}$. This can be achieved by producing a continuous phase rotation along the aperture. Thus, if β is the ratio of the phase velocity of the feed along the aperture to that in free space, then the radiation pattern of the constant-amplitude aperture becomes

$$\cos\theta \frac{\sin\left[\frac{\pi a}{\lambda}(\beta + \sin\theta)\right]}{\frac{\pi a}{\lambda}(\beta + \sin\theta)}$$

In practice, the angle of tilt is not likely to exceed 1 deg., and for such small angles there is negligible change in the shape of the pattern. It may be seen that in order to tilt the angle of the main lobe by α deg., we require that $\beta = -\sin\alpha$ deg. As an illustration of the results of the application of these principles, Fig. 11 shows the calculated patterns obtained from an aperture $21{\cdot}3\lambda$ long (this corresponds to a length of 32 ft at 655 Mc/s). This figure shows the pattern obtained from a linear co-phased aperture, and also the pattern obtained when the nulls have been completely filled, and the main lobe tilted by 0·5 deg. As a comparison, the pattern of an aerial having an ideal cosec θ pattern from $\theta = 0{\cdot}5$ deg., and zero radiation for

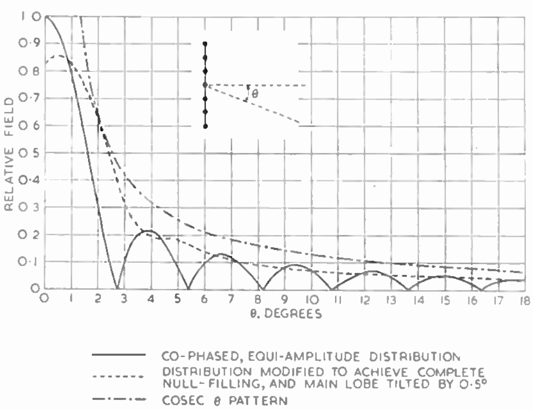


Fig. 11. The vertical radiation pattern of a linear aperture, $21{\cdot}3\lambda$ long.

$\theta < 0 \cdot 5$ deg., is also shown. The curves are scaled to allow for the different gains obtained in each case.

The extent to which the theoretical requirements will be met in practice depends very largely upon the amount of complexity that can be tolerated in the feeding arrangements. It must also be remembered that in a particular case, it may not be necessary to achieve complete null-filling.

The physical form of the aerial may be one of many types. In this country, a side-fire helical aerial has been used for the experimental transmissions. In America, stacks of slots have been used to achieve gains of up to 17·8 db and a waveguide form of aerial[8] has also been produced. Whatever form is adopted, careful consideration will have to be given to the mechanical rigidity of the structure, because with the high gains employed, an angular deflection of 0·5 deg. could be important. This may be demonstrated by considering Fig. 11. It may be seen that the relative field (in the gap-filled case) at an angle of 2·5 deg. is 0·48. If this angle changes by $\pm 0 \cdot 5$ deg., the field varies from 0·64 to 0·32, i.e. a variation of $+2 \cdot 5$ db to $-3 \cdot 5$ db.

5.1. *Scale models*

In developing a transmitting aerial, it is very convenient to carry out measurements upon scale models, the frequency being scaled accordingly. It is therefore of some interest to consider any difficulties which might arise in this connection.

Possibly the most important factor is that owing to the much shorter wavelengths, greater attention will be required to the scaling of details of the mast structure, etc. When making a scale model of, say, a Band I aerial, it is usually permissible to omit many small details, but for a Band V aerial, much more accurate scaling will be required. It is also possible that smaller scaling factors will have to be used. In the past, it has been found convenient to use models working at a frequency of about 400-600 Mc/s, this representing a scale factor of about 10:1 for a Band I aerial. In view of the more accurate scaling which will be required for a Band V aerial, and also because of possible practical difficulties associated with using frequencies of more than, say, 2000 Mc/s, it is probable that scaling factors of not more than 3:1 will be used. This may imply physically larger models than are normally used at present. If it is desired to carry out vertical radiation pattern measurements on a scale model of an aerial in which beam shaping of any kind is employed, then great care will have to be taken with the feeding arrangements on the model, to ensure that the pattern is the same as on the final full-scale aerial.

6. Conclusions

The development of television transmitting aerials for use in Band I up to the present day has been surveyed. In the future, the use of higher frequencies will enable higher gains to be achieved, and it will probably be desirable to attempt to shape the radiation pattern to achieve a more uniform field in the service area.

7. Acknowledgments

The author wishes to thank the Chief Engineer of the British Broadcasting Corporation for permission to publish the paper. Thanks are also due to colleagues in the Research Department of the B.B.C., upon whose work much of the information contained in the paper is based.

8. References

1. H. Cafferata, C. Gillam and J. F Ramsey, " Television transmitting aerials ", *Proc. Instn Elect. Engrs*, **99**, Part IIIA, p. 215, April-May 1952.
2. R. F. Holz, " Super-turnstile antenna ", *Communications*, **4**, p. 40, 1946.
3. G. J. Phillips, " V.h.f. aerials for television broadcasting ", *Proc. Instn Elect. Engrs*, **102**, Part B, p. 687, Sept. 1955.
4. P. M. Woodward and J. D. Lawson, " The theoretical precision with which an arbitrary radiation-pattern may be obtained from a source of finite size", *Proc. Instn Elect. Engrs*, **95**, Part III, p. 363, September 1948.
5. R. L. Smith-Rose, " International radio organisations: some aspects of their work ", *J.Brit.I.R.E.*, **18**, p. 631, November 1958.
6. P. Knight, "Methods of calculating the horizontal radiation patterns of dipole arrays around a support mast ", *Proc. Instn Elect. Engrs*, **105**, Part B, p. 548, November 1958.
7. P. S. Carter, " Antenna arrays around cylinders ", *Proc. Inst. Radio Engrs*, **31**, p. 671, 1943.
8. O. M. Woodward and J. Gibson, " The omniguide antenna—an omnidirectional waveguide array for u.h.f. television broadcasting ", *R.C.A. Review*; **17**, p. 13, March 1956.

# of current interest . . .

Communications and Radar Installations for the Port of London

On May 1st, 1959, the Port of London Authority opened a new Thames Navigation Service aimed at greatly increasing the safety and speedier working of Thames shipping, particularly during bad weather. Basically, the service consists of a comprehensive radio-telephone system extending from London Bridge to the outer limits of the port beyond the Nore, backed up by a harbour radar covering some five miles of river around Gravesend Reach. The purpose of the Service is to provide ships with information on berthing and all necessary complementary navigational data.

The five radio-telephony channels in the band 156-174 Mc/s, allocated under international agreement to the service, are handled by individual transmitter/receivers at Shooters Hill in South-East London and at All Hallows, Kent. These stations are remotely operated from the control room at Gravesend.

Pye Marine Limited have supplied and installed v.h.f. f.m. transmitting and receiving radio-telephone equipment and have also installed at the control centre a special tape recorder capable of simultaneously recording, on all five channels, conversations and messages for future reference. The operations room can carry on two way conversations with ships entering the Port and an additional facility enables a ship to be connected directly to a Dockmaster's office by land line.

The first Harbour Radar for the Service is a Decca Harbour Radar Type 33, a high definition, X-band equipment using an aerial of 6 foot span which is mounted on the roof of the control room building. There are two separate transmitters and receivers with remotely-controlled changeover switching arrangements. A 15 inch diameter display of fixed coil type is sited in the operations room which is designed so that, as a result of experience, further radar stations can be set up both above and below Gravesend, feeding additional displays. Eventually, if the demand necessitates, the entire river from the seaward limit of the port up to the Royal Docks will be shown on eight radar displays at one central position.

At Gravesend three operators sit at a control desk facing a 20 feet long illuminated map of the river which is marked with all information of nautical significance. Existing tide gauges on the river are also linked to record directly into the operations room.

Thermoelectric Generator for Operation in Outer Space

Long mission satellites and manned space vehicles of the future will be able to tap a limitless supply of electric energy according to engineers of the U.S. companies of Westinghouse and Boeing who recently showed a working model of a solar-powered thermoelectric generator. The generator itself weighs three pounds and measures 20 inches in length, and is capable of converting the energy of the sun into 2·5 watts of power—enough to operate a transmitter broadcasting a strong signal back to earth.

The direct conversion of heat to electricity through a static device involves the Seebeck effect whereby a current flows continuously in a closed circuit composed of two different metals so long as the junctions of the metals are maintained at different temperatures. Basically the problem is first to get heat from the sun into the system and then dissipate the waste heat from the "cold" junction in the vacuum environment of space.

In order to evaluate these problems a "bench model" generator which can be easily instrumented but does not represent an attempt to attain the optimum flyable system has been developed. The sun's energy is collected and concentrated by a mirror-like device similar to a solar furnace. This will eventually lead to mechanical problems because the collector must have a large area which will have to be in a collapsed form during the ascent of the system through dense air and must be extended or erected at space altitudes. The development generator has a reflective aluminium sheet mounted in a 20 by 50-inch semi-cylindrical shape to collect the sun's energy and concentrate it on the thermoelectric material.

A Negative Resistance for D.C. Computers†

by

P. V. INDIRESAN, B.SC., ASSOCIATE MEMBER‡

Summary: A d.c. negative resistance suitable for use in analogue computers is described. The circuit uses two transistors and eight resistances and is compensated against unbalance in the two transistors. A nomogram has been constructed to simplify the design of the circuit for any required value of negative resistance. There is also a discussion of the properties of, and the differences between series and shunt type of negative resistances. The circuit described gives a shunt type of negative resistance.

1. Introduction

A number of engineering problems require the computation of various types of fields, such as stress fields, thermal fields, flow fields, electric fields, etc. These are often solved by the method of finite difference equations.[1] To compute this mathematically would be laborious. On the other hand, in electrical networks, the result can be obtained directly, by measuring the voltage at the required point. This advantage may be extended to other types of fields also, by using an electrical analogue of the field to be studied.

In constructing the electrical analogue, it is preferable to work with direct currents using resistances only. This would not only be more economical, it would also be easier to operate and construct; there would be no difficulty due to radiation; no errors due to the finite power factor that occur when inductances and capacitances are used and there would be no difficulties due to distributed capacitance or inductance. However, for many computations, positive resistances alone are not sufficient; negative resistances also would be required. (A negative resistance may be defined as one whose voltage-current characteristic has a negative slope at the voltages and frequencies used.) For example, negative resistances are required in the solution of the wave equation. In solving aerodynamical problems at supersonic speeds, all elements in the direction of supersonic flow, should be represented by negative resistances only. The object of this investigation is to devise a negative resistance suitable for this purpose.

Swenson and Higgins[2] have described one method of obtaining such a negative resistance. The circuit is shown in Fig. 1. When the voltmeter V reads zero, the voltage between the points a and b is twice the input voltage E_{in}, and

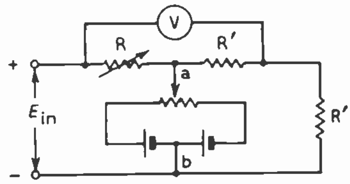


Fig. 1. A negative resistance circuit.

a current E_{in}/R flows into the positive input terminal. That is, at the terminals 1 and 2, the circuit has a negative voltage/current ratio, or is behaving as a negative resistance of magnitude R. This circuit is simple to construct, but suffers from the serious defect. that its adjustment is critically dependent on the external network connections. If there are a number of these negative resistances in this circuit adjusting one would affect the operation of all the others. In consequence, they will all have to be adjusted repeatedly until all the voltages are zero. Not only is this time consuming but there is no proof that this process of iterative correction is always convergent. If it is not so, it would become impossible to use this system.

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The problem may now be stated: It is required to devise a negative resistance, which

(a) will operate at d.c.,

(b) will be independent of external connections,

(c) is stable and is also, if necessary, adjustable, and

(d) is compact, economical in cost and in power consumption, so that it can be used in large numbers.

2. Theory of Negative Resistance

There is a large volume of literature[3-17] available on the theory, design and operation of negative impedances, and a number of phenomena are known to exhibit the characteristic of negative impedance. The earliest to be utilized was perhaps the series generator. The other examples include the arc discharge, the dynatron, the retarding field type of negative resistance, the space charge grid tube, the split anode magnetron, the low gas pressure triode, etc. None of these however, are linear nor are they stable in operation or magnitude. At present, the almost universal method of obtaining negative impedances is by the use, in some form or other, of stabilized positive feedback.

2.1. *The concept of negative resistance*

In the study or measurement of positive resistances no distinction is made, whether it is the current that causes the voltage drop or it is the voltage that makes the current flow: it is immaterial. On the other hand, in the case of negative resistances, this distinction must be made. This leads to two different types of negative resistances. These are (a) the current actuated or series type of negative resistances, and (b) the voltage actuated or shunt type of negative resistances.

A series type of negative resistance may be defined as one in which a decrease in current, instead of causing a voltage drop, causes a voltage rise. This may be considered to be a negative of voltage drop. Then

$$R_N \text{ (series)} = \frac{(-V)}{I} \qquad \text{.........(1)}$$

The shunt type of negative resistance may be defined as one in which an increase in voltage causes a decrease in current and vice versa. In this case, it is the current that is negative. Then

$$R_N \text{ (shunt)} = \frac{V}{(-I)} \qquad \text{.........(2)}$$

It is possible to define negative resistances in terms of power also. A negative resistance may then be described as one which instead of consuming power, delivers power. If this happens when a current is passed through it, it would be of the series type and the magnitude of the negative resistance would be given by

$$R_N \text{ (series)} = \frac{(-P)}{I^2} \qquad \text{.........(3)}$$

A shunt type of negative resistance returns power to the associated circuit when a voltage is applied to it. Its magnitude would be given by

$$R_N \text{ (shunt)} = \frac{E^2}{(-P)} \qquad \text{.........(4)}$$

The fact that a negative resistance delivers power does not mean that it is equivalent to, say, a battery. The e.m.f. of a battery is independent of the current flowing through it. The voltage across a negative resistance on the other hand is directly proportional to the current flowing through it. In other words, a negative resistance and a generator are in no way interchangeable.

2.2. *The positive feedback method of obtaining negative resistance*

The type of negative resistance obtained by positive feedback depends on the type of feedback. Current feedback will give the series type of negative resistance and voltage feedback the shunt type of negative resistance.

In a current feedback circuit, let i be the input current and k the feedback factor. Then the total current flowing between the input terminals is $i(1-k)$. If R is the total effective resistance connected across the input terminals, the resultant input voltage will be $iR(1-k)$. This makes the input resistance to be

$$R_{in} = e_{in}/i = R\,(1-k) \qquad \text{.........(5)}$$

When k is greater than unity, R_{in} will be negative and correspond to a series type of negative resistance.

For the case of voltage feedback, it is more convenient to think in terms of conductances rather than in terms of resistances. Let Y be the total effective conductance connected across the input terminals, and e the voltage applied to it. Then, if k is the feedback factor, the net voltage across the input terminals will be $e\ (1-k)$ and the resultant current will be $e\ (1-k)\ Y$. That is, the input conductance Y_{in} is equal to

$$Y_{in} = i_{in} / e_{in} = (1-k)\ Y \qquad \text{........(6)}$$

If k is greater than unity, the conductance becomes negative. This is the case of a shunt type of negative resistance.

It must be noted that it is not essential that only a resistance be connected across the input terminals. It can in fact be any type of impedance. This means that it would be possible to obtain by this method not only negative resistances but also negative capacitances or negative inductances or for that matter any negative impedance. The range of operation of the resultant negative impedance in frequency or in amplitude, as also its stability (in magnitude) depends on the characteristics of the amplifier circuit.

The magnitude of the negative impedance depends on the factor $(1-k)$. To obtain a large negative impedance effect k must be as large as possible. But the feedback factor k is given by

$$k = \frac{A}{1-AB} \qquad \text{........(7)}$$

where A is gain of the amplifier and B the fraction of the output that is fed back. So, in order to make k large AB must be brought close to unity. But, if AB approaches unity, the slightest changes in the gain of the amplifier will cause large changes in the magnitude of k and hence in that of the negative impedance. On the other hand if AB is made large k tends to become small and even small changes in k may cause large changes in $(1-k)$. The best compromise is therefore to make the gain A large, stabilize it by negative feedback and adjust so that k is reasonably high.

In any electronic circuit, it is impossible to eliminate completely the shunting effect of stray capacitances. The effect of this stray capacitance is to add a shunt negative-capacitance type of reactance to shunt type of negative resistances and a series negative inductance reactance to series type of negative resistances.

Another point of interest to be noted is that an increase in the gain of the amplifier increases the series negative resistance but decreases the shunt negative resistance.

2.3. *Amplifier gain and stability*

In a series negative resistance circuit, if the amplifier gain is zero, the input resistance becomes equal to R. As the gain increases from zero, $(1-k)$ decreases and the input resistance also decreases. When k equals unity the input voltage equals the feedback voltage and the total voltage drop becomes zero. This corresponds to zero input resistance and the current is limited only by the internal impedance of the input voltage source. When the gain increases further, the voltage increases and the input resistance becomes negative. When the gain becomes so large, that the current increase is sufficient to overcome all the losses in the circuit, the circuit becomes unstable. At this stage the magnitude of the negative resistance is greater than the total positive resistance in series with it. For stable operation, therefore, the series type negative resistance must have a *larger* positive resistance in series with it.

In the shunt negative resistance circuit, when the amplifier gain is zero, the input conductance equals Y. As the gain increases $(1-k)$ decreases and the magnitude of the input conductance also decreases proportionately. When k equals unity, the input conductance drops to zero. When the gain increases further, the current reverses and the input conductance becomes negative. When the gain becomes so large that the increase in the voltage output is able to overcome all the losses in the circuit, the circuit becomes unstable. At this stage the shunt negative conductance is greater than the total positive conductance in parallel with it; or, the shunt negative resistance is smaller than the total positive resistance in parallel with it. That is, for stable operations, the shunt type negative resistance must be associated with a *smaller* positive resistance in parallel with it.

The conditions of stability for the two types of negative resistances are, therefore, exactly opposite to each other. Because of this it is *impossible* to predict the stability of a circuit containing a negative resistance, *unless* it is definitely known how the negative resistance has been obtained. Hence, a distinction must always be made between resistances which are negative due to a negative voltage component (series or current operated type) and those which are negative because of a negative current component (shunt or voltage operated type).

2.4. *Negative resistance and equivalent circuits*

Often it is not realized that Thévénin's and Norton's theorems are correct only so far as the external voltages and currents are concerned and do *not* give a correct picture of the total power consumed in the circuit. For example consider the two circuits shown in Fig. 2. They are exactly equivalent to each other as far as the output resistance ($-R_N$) is concerned, and they both give

E_N = voltage across negative resistance

$$= -\frac{R_N}{R_P - R_N} E_g \qquad \text{.........(8)}$$

I_N = current through the negative resistance

$$= E_g / R_P - R_N \qquad \text{.........(9)}$$

$P_N = E_g^2 R_N / (R_p - R_N)^2$ = power in the negative resistance ....(10)

However if the positive power loss is calculated it appears from Fig. 2(a) to be equal to $E_g^2 R_P / (R_P - R_N)^2$ whereas Fig. 2(b) indicates it to be $E_g^2 R_N^2 / (R_P - R_N)^2$. Evidently, the equivalence between the two circuits holds good only so far as the external circuit is concerned and not for the entire circuit. Due to this, from Fig. 2(a) it appears as though the circuit will be stable only when R_P is greater than R_N, whereas Fig. 2(b) indicates that it should actually be smaller. This ambiguity will be avoided if, and only if, the voltage generator and mesh system of analysis alone is used for series negative resistances and the current generator and nodal system of analysis only is used for shunt type of negative resistance circuits. However, if the voltage and currents and not the stability of the negative resistance is required, either method may be used.

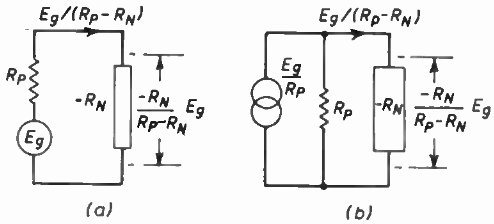


Fig. 2. Thévénin's and Norton's theorems applied to a circuit containing a negative resistance.

Table 1 summarizes the characteristics of series and shunt type negative resistances.

3. Negative Resistance Circuit for Direct Currents

3.1. *The Waldmann-Bieri circuit*

Although a number of circuits are known that will give a negative impedance for a.c., there appears to be only one described which is suitable for d.c., namely, that due to Waldmann and Bieri.[18] Their circuit uses two valves as shown in Fig. 3. This may be seen to be a modification of the well-known multivibrator circuit. The main difference lies in the fact that the cross-connection between the valves is by means of the two batteries B1 and B2. The negative resistance characteristic appears between the points A and B.

Table 1

Comparison of Series and Shunt Type of Negative Resistances

Parameter	Series Type	Shunt Type
Independent variable	Current operated	Voltage operated
Negative component	Voltage drop	Current
Condition of stability	Open circuit stable	Short circuit stable
Effect of internal gain reduction	R_N decreases	R_N increases
Associated reactance	Series inductance	Shunt capacitance
Proper method of analysis	Mesh method	Nodal method

The two types of negative resistances are circuit duals.

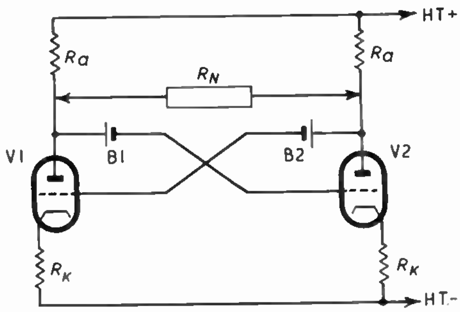


Fig. 3. Waldmann and Bieri's Circuit.

A physical explanation of this circuit would be as follows: when a d.c. voltage is applied between the two points A and B, a current is drawn by the anodes of the two valves. At the same time, through the batteries B1 and B2, the same voltage is passed on to the grids of valve 2 and 1 respectively. This change in grid voltage causes a change in the anode current of the two valves and this is opposite in sign to the current drawn by the anodes from the voltage source V_{AB}. This corresponds to a negative current. If the magnitude of this is greater than that of the positive current drawn by the anode circuit, the net current drawn by the circuit from the voltage source V_{AB} becomes negative. In other words, the circuit behaves as a shunt type of negative resistance at the two points A and B.

3.2. *Disadvantages of the Waldmann and Bieri circuit*

Although the Waldmann and Bieri circuit does give a shunt type of negative resistance that can operate down to d.c., it still possesses a few serious disadvantages from the point of view of computer application. These are:

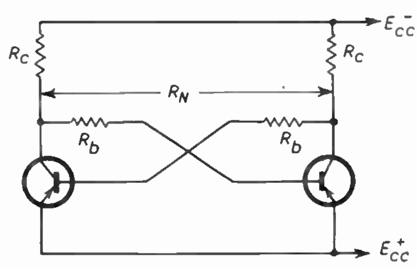


Fig. 4. A transistor d.c. negative resistance circuit.

(1) Each negative resistance circuit requires four batteries, one for anode supply, one for heater and two for interconnection between the two valves.

(2) The use of the valves themselves is unsuitable as they occupy space and consume a lot of power.

(3) The circuit is costly: in first cost because of the batteries, and in running cost because of the large power consumption.

The answer to these difficulties is the use of transistors in place of valves. It will be shown that the use of transistors does not only cut down the size and power consumption, but also reduces the number of batteries required from four to one only.

3.3. *A transistor shunt negative resistance circuit*

The version of the Waldmann-Bier circuit using transistors is shown in Fig. 4. Two p-n-p junction transistors replace the two triode valves. The bias batteries B1, B2 are no longer required, their place being taken by two resistances R_b. The only source of power required is for the collectors. But for these differences, the circuits of Fig. 3 and Fig. 4 are exactly similar.

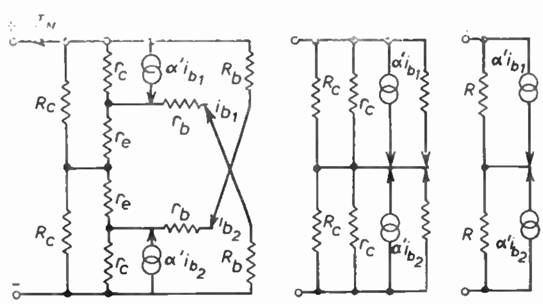


Fig. 5. Equivalent circuit for transistor negative resistance.

Figure 5(a) shows the exact equivalent of the transistorized negative resistance circuit. As the emitter resistance r_e is much smaller than any of the other resistances in the circuit, it may be to a first approximation neglected. Then the equivalent circuit reduces itself to the form shown in Fig. 5(b) which in turn can be reduced to the form shown in Fig. 5(c).

The current I drawn from the external voltage source V_{AB} is given by Fig. 5(c) as

$$I = i_R + \alpha' i_{b1} \qquad \text{.........(11)}$$

$$i_R = \frac{V_{AB}}{2R} \qquad \text{.........(12)}$$

where $$\frac{1}{R} = \frac{1}{R_L} + \frac{1}{r_d} + \frac{1}{R_b + r_b} \qquad \text{.........(13)}$$

R_L = collector load resistance.

r_d = collector resistance for common emitter configuration.

r_b = base resistance.

But $$i_{b1} = -V_{AB}/2R_b' \qquad \text{.........(14)}$$

where

$$R_b' = R_b + r_b.$$

From eqns. (14), (12) and (11).

$$I = V_{AB}\left(\frac{1}{2R} - \frac{\alpha'}{2R_b'}\right) \qquad \text{.........(15)}$$

or the conductance between the terminals A and B is

$$Y_{AB} = Y_N = \frac{1}{2R} - \frac{\alpha'}{2R_b'} \qquad \text{.........(16)}$$

This is negative when $\alpha'/2R_b'$ is greater than $1/R$.

Equation (16) shows that the negative resistance obtained by this transistor circuit depends mainly on the current amplification factor α' and on the base circuit resistance R_b'. The voltage/current characteristic of this circuit is shown in Fig. 6. It may be seen that the line does not pass exactly through the origin and that beyond a certain voltage level it bends off until the resistance becomes positive. Any unbalance in the circuit will cause a differential voltage to be set up between the points A and B, and this shifts the $V-I$ characteristic so that it does not pass through the origin. The voltage limits to the operation of the circuit are generally determined by the supply voltage E, and the resistance R_b. It should also be noted the magnitude of the negative resistance is affected by changes in α'.

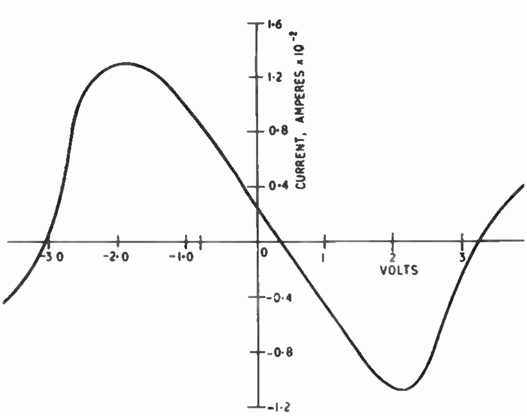


Fig. 6. The negative resistance characteristic of the circuit of Fig. 4, using OC 71 transistors with $R_b = R_c = 1000$ ohms.

3.4. *Stability of the magnitude of the negative resistance*

Equation (16) shows that the magnitude of Y_N depends on α', r_b the base resistance and R_b the resistance connected in the base circuit. By using high stability close tolerance resistors R_b can be held very constant. But r_b is liable to vary considerably with temperature and from one transistor to another. Still the effect of its variations on Y_N may be made small by choosing a high value of R_b. This, however, reduces the magnitude of the negative resistance obtained.

The current amplification factor α' also is far from a constant. Even among similar makes of transistors, it is liable to vary over a wide range. A range of 30 to 80 is a typical one, but it does not vary appreciably with temperature.

Hence, it can be stated that the magnitude of the negative resistance is critically dependent on the individual transistors used and should be expected to vary over a range of 3:1 or more even with identical circuits. It is also to a certain extent dependent on temperature, mainly due to variations in r_b.

3.5. *The effect of asymmetry in the circuit*

The negative resistance circuit is a symmetrical one and any asymmetry in the components used would cause unequal drops across the two collector load resistances. This in turn will cause a voltage to appear across the negative resistance terminals A B, even when no external voltage is applied between them. If high-stability, close tolerance resistors are used, this unbalance will be almost entirely due to the difference in the two transistors. This difference in the transistors may be due to (a) unequal values of α', and (b), unequal values of collector back current I_{CO}.

4. A Stabilized Negative Resistance Circuit

A modification of the above circuit which gives a high degree of stability is shown in Fig. 7.

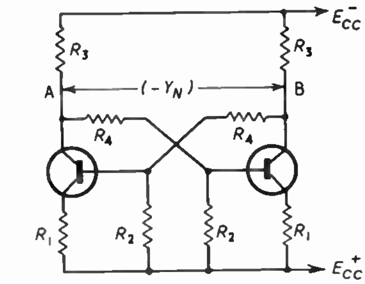


Fig. 7. The stabilized negative resistance circuit.

The circuit can be analysed if the transistors are replaced by their equivalent circuits as shown in Fig. 8. For calculating the magnitude of Y_N the circuit is assumed to be exactly symmetrical. Then the two changes in the two collector voltages, when an external voltage is applied to the negative resistance terminals, would be equal and opposite. The same will be true of the base and emitter voltages also. From Fig. 8 the following nodal equations may be written down:

$$\begin{aligned} e_1(Y_3+Y_4)+e_2(0)+e_3(Y_4) &= I+\alpha i_e \\ e_1(0)+e_2(Y_1+Y_b)+e_3(-Y_b) &= -\alpha i_e \\ e_1(Y_4)+e_2(-Y_b)+e_3(Y_2+Y_4+Y_b) &= 0 \end{aligned} \qquad \text{.........(17)}$$

and $$\alpha i_e = -e2Y_1 \qquad \text{.........(18)}$$

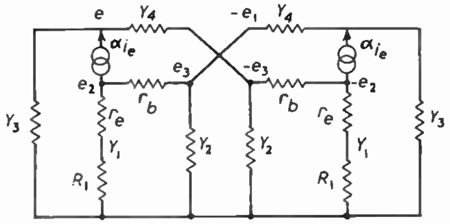


Fig. 8. Equivalent circuit for the stabilized negative resistance.

The negative conductance appears at the terminals A and B and is equal to $I/2e_1 = Y_N$ (say). Then Y_N is given by

$$\frac{(Y_3+Y_4)\,(1-\alpha)Y_1+Y_b\,(Y_2+Y_4+Y_b)-Y_b^2-Y_4\,(1-\alpha)Y_1+Y_b\,Y_4+Y_1Y_b}{2\,(1-\alpha)Y_1+Y_b\,(Y^2+Y_4+Y_b)-2Y_b^2} \qquad \text{.........(19)}$$

As α is very nearly equal to unity, it may be assumed without much error that $(1-\alpha)\,Y_1 \ll Y_b$. Then the expression simplifies to

$$2Y_N = Y_3 - Y_4 \,.\, \frac{\alpha Y_1 - Y_2}{Y_2 + Y_4} \qquad \text{.........(20)}$$

This expression will be quite accurate if R_1 is large. A comparison with experimental values is given in Table 2.

Table 2

Experimental Verification of Formula for Y_N (all conductances in micro-ohms)

Y_1	Y_2	Y_3	Y_4	Y_2'	Y_N calculated	Y_N experimental
500	10,000	1000	1000	7750	580	550
500	10,000	1000	666	7750	1100	1050
500	10,000	1000	2000	7750	2000	2000
500	10,000	1000	4000	7750	2350	2500
500	10,000	400	1000	7750	2320	2150
500	10,000	666	1000	7750	1900	1820
500	10,000	2000	1000	7750	725	760
500	3,330	1000	1000	2950	235	220
500	5,000	1000	1000	4325	580	550
500	20,000	1000	1000	13,000	2750	2525
200	10,000	1000	1000	7750	1590	1470
667	10,000	1000	1000	7750	1360	1300

Equation (20) shows that Y_N depends mainly on α and r_e. α is always nearly unity and r_e is small compared to the resistances used in the circuit. Incidentally, r_e is one transistor parameter that is virtually independent of temperature.[19] Equation (20) also shows that to increase Y_N, Y_3 and Y_2 must be decreased or Y_1 and Y_4 must be increased. It is also necessary to make $Y_1 > Y_2$.

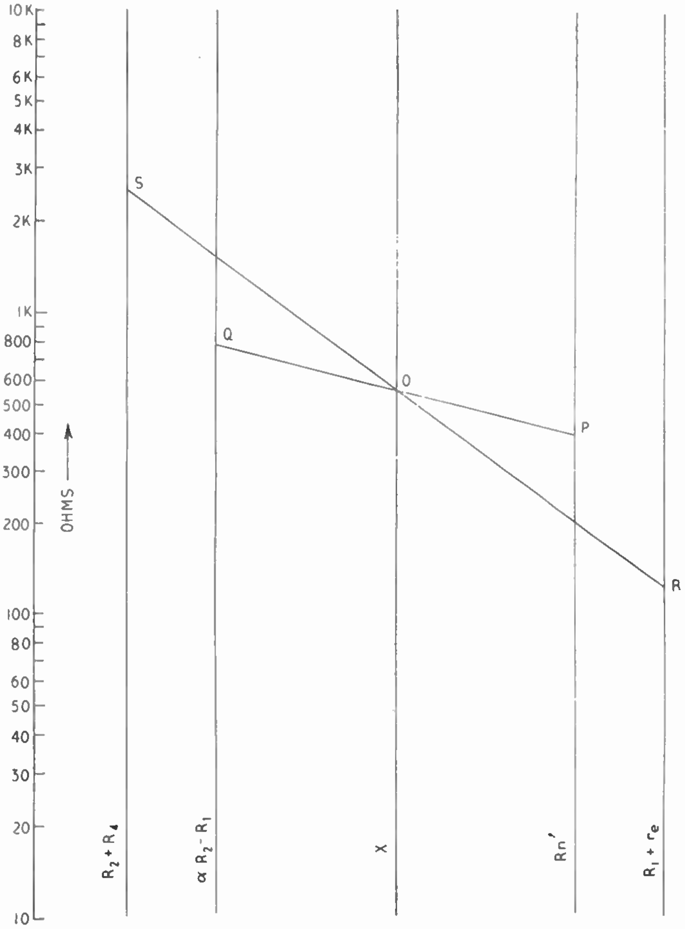
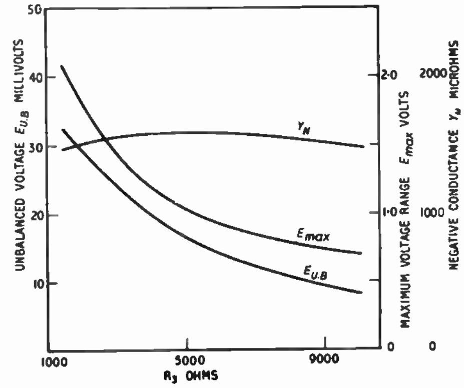


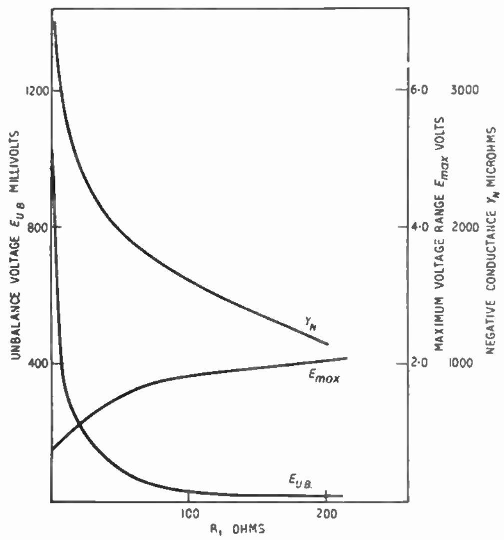
Fig. 9. Nomogram for designing the negative resistance circuit.

4.1. Nomogram for the design of the circuit

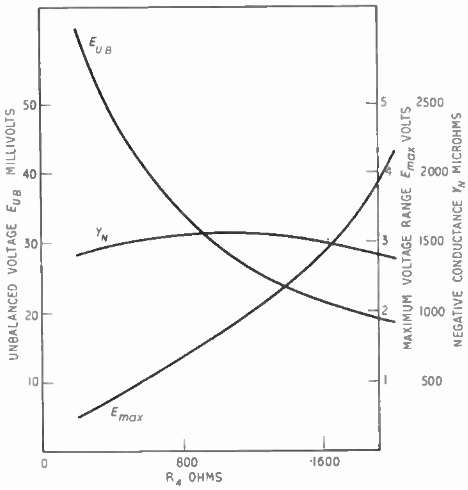
Figure 9 shows a nomogram that can be used to design the circuit. The procedure for using the nomogram is as follows: Let Y_n be the required value of the negative conductance. The value of the collector load resistance is at first assumed. This is not a severe limitation as its value is not critical, as can be seen from Fig. 10(c). A reasonable assumption for R_3 would be $2/Y_n$. From these values of Y_n and R_3, the factor $R'_n = R_3/(1+2Y_{n3})$ is calculated. This is then plotted on the R'_n scale on the nomogram. Let it be at the point P. Through P a straight line POQ is drawn intersecting the X scale at O and the $\alpha R_2 - R_1$ scale at Q. Through the point O another straight line ROS is drawn intersecting the scale of R_1+r_e and of R_2+R_4 at R and S respectively. Then the point R gives the value of R_1+r_e; r_e the emitter resistance can be obtained from the table of transistor constants and thus R_1 is known. The point Q gives the value of $\alpha R_2 - R_1$. From this R_2 can be determined as α is a transistor constant (which is always nearly unity) and R_1 is known. Finally R_4 can be obtained from the point S. The lines POQ and ROS may be so drawn as to get suitable values for the resistances R_1, R_2 and R_4. The nomogram can also be used to determine the values of Y_n for a given set of components. For $R_1 = 100\Omega$; $R_2 = 1000\Omega$; $R_3 = 2000\Omega$ and $R_4 = 1500\Omega$, the nomogram gave $Y_n = 1000\Omega$. Experimentally this was found to be 1080Ω. In designing the circuit either R_1 or R_2 may be



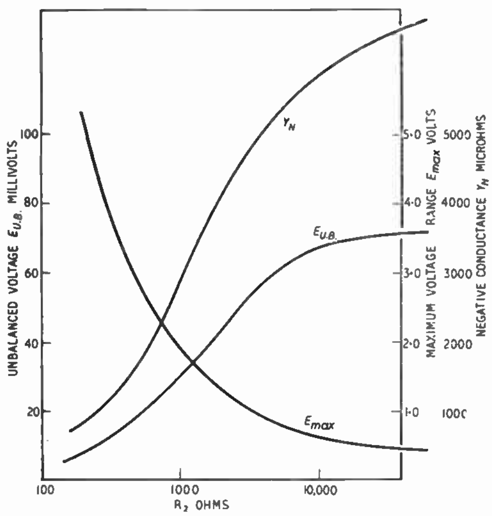
(c) Effect of varying R_3.



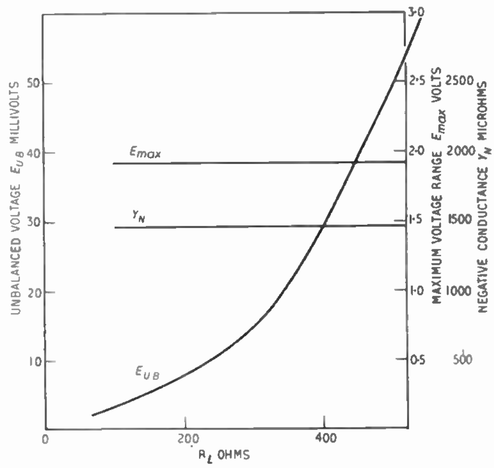
(a) Effect of varying R_1.



(d) Effect of varying R_4.



(b) Effect of varying R_2.



(e) Effect of varying R_L, the load across the negative resistance.

Fig. 10. Negative resistance characteristics

made adjustable if an exact value of Y_n is required.

The variation of (a) Y_n, (b) the unbalanced voltage at the negative resistance terminals, and (c), the maximum voltage range of operation, for various values of R_1, R_2, R_3, R_4 and of the load resistance R_L are shown in Fig. 10.

5. Conclusion

A shunt type of negative resistance circuit for d.c. voltages has been described. The circuit configuration is the same as that of a multivibrator. It is much more advantageous to use transistors rather than valves, although transistors make it more difficult to balance the circuit. A method of stabilizing against variations in temperature and in transistor parameters has been described. A nomogram has also been drawn, so that the negative resistance of any desired value may be obtained. A series of experimental curves for various values of circuit parameters, using type OC 71 transistors, is also given. This type of negative resistance is particularly suitable for d.c. analog computers, as it is not affected by external connections. (See Fig. 10(e)).

6. Acknowledgments

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7. References

1. W. W. Soroka, "Analog Methods in Computation and Simulation," p. 269. (McGraw-Hill, New York, 1954).
2. G. W. Swenson, Jr. and J. J. Higgins, "A direct current network analyser for solving wave-equation boundary-value problems," *J. Appl. Phys.*, **23**, pp. 126-31, January 1952.
3. J R. Tillman, "A negative impedance converter for use as a two-terminal amplifier" *P.O. Elect. Engrs J.*, **48**, pp. 97-101, July 1955.
4. E. L. Ginzton, "Stabilized negative impedances," *Electronics*, **18**, 1945: Pt. 1, pp. 140-150, July; Pt. 2, pp. 138-148, August; Pt. 3, pp. 140-144, September.
5. C. Brunetti and L. Greenhough, "Some characteristics of a stable negative resistance," *Proc. Inst. Radio Engrs*, **30**, pp. 542-546, December 1942.
6. J. T. Bangert, "The transistor as a network element," *Bell Syst. Tech. J.*, **33**, pp. 329-352, March 1954.
7. J. L. Merrill, Jr., A. F. Rose and J. O. Smethurst, "Negative impedance telephone repeaters," *Bell Syst. Tech. J.*, **33**, pp. 1055-1092, September 1954.
8. J. G. Linvill, "Transistor negative-impedance converters," *Proc. Inst. Radio Engrs*, **41**, pp. 725-729, June 1953.
9. J. L. Merrill, Jr. "Theory of the negative impedance converter," *Bell System Tech. J.*, **30**, pp. 88-109, January 1951.
10. L. C. Verman, "Negative circuit constants," *Proc. Inst. Radio Engrs*, **19**, pp. 676-681, April 1931.
11. C. Brunetti, "The transistron oscillator," *Proc. Inst. Radio Engrs*, **27**, pp. 88-94, February 1939.
12. F. E. Terman *et alii*, "Some applications of negative feedback with particular reference to laboratory equipment. *Proc. Inst. Radio Engrs*, **27**, pp. 649-655, October 1939.
13. S. P. Chakravarti, "The bandpass-effect," *Wireless Engineer*, **18**, pp. 103-111, March 1941.
14. G. Crisson, "Negative impedances and the twin 21-type repeater," *Bell Syst. Tech. J.*, **10**, pp. 485-513, July 1931.
15. E. W. Herold, "Negative resistance and devices for obtaining it," *Proc. Inst. Radio Engrs*, **23**, pp. 1201-1223, October 1935.
16. J. R. Tillman, "A note on electronic negative resistors," *Wireless Engineer*, **22**, pp. 17-24, January 1945.
17. D. M. Tombs, "Negative and positive resistance," *Wireless Engineer*, **19**, pp. 341-346, August 1942.
18. L. Waldmann, and R. Bieri, "A negative resistance for direct and alternating current," *Z. Naturforsch*, **10**A, No. 11. pp. 814-820, 1955 (In German).
19. E. Keonjian, "Temperature-compensated d.c. transistor amplifier," *Proc. Inst. Radio Engrs*, **42**, pp. 661-671, April 1954.

Ultraharmonic and Subharmonic Resonance in an Oscillator†

by

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Summary: Different order subharmonic and ultraharmonic resonance curves for as oscillator with a cubic non-linearity are obtained theoretically by van der Pol's method. Their distinctive features are discussed. Experimental results obtained with the help of an electronic differential analyser are compared with the theoretical curves.

1. Introduction

A self-excited oscillator may exhibit harmonic, ultraharmonic or subharmonic resonance when it is subjected to a periodic signal.[1-5] In harmonic resonance which occurs when the frequency of the external signal is approximately equal to that of the oscillator, the free oscillation is suppressed and the oscillator executes only the forced oscillation. On the other hand in ultraharmonic or subharmonic resonance the free oscillation is not suppressed but its amplitude is altered and the frequency modified so as to be equal to the multiple or submultiple of that of the external signal. In ultraharmonic resonance the modification is due to the ultraharmonic of the forced oscillation having a frequency nearly equal to that of the oscillator. Its amplitude is exclusively determined by that of the external signal for small non-linearities in the oscillator and hence the general characteristics of ultraharmonic resonance are similar to those of harmonic resonance, there being variations only in details. However, in subharmonic resonance it is the difference frequency signal having the frequency approximately equal to that of the oscillator which modifies the free oscillation. Since the amplitude of the difference frequency signal is determined by the amplitude of both the free and forced oscillation, the characteristics of subharmonic resonance are quite distinct from the harmonic or ultraharmonic resonance.

Harmonic resonance curves have been studied in detail by several workers but ultraharmonic and subharmonic resonance curves have received comparatively less attention. In the present paper different order ultraharmonic and subharmonic resonance curves of an oscillator with a cubic non-linearity are obtained following the method used by van der Pol. The criteria for determining the stability of the resonance condition are deduced and the details of the characteristic features of the resonance curves are studied. Experimental data obtained with the help of a differential analyser are also discussed.

2. Theoretical Derivation of Resonance Curves

Theoretical study of the resonance characteristics of a self-excited oscillator subjected to an external signal involves the solution of its describing differential equation. The exact form of the equation is dependent on the circuit arrangement of the oscillator. However, the representative differential equation may be taken to be of the form

$$\ddot{x} - a\dot{x} + \mu f(x, \dot{x}) + \omega_0^2 x = E \sin \omega t \qquad \ldots\ldots\ldots(1)$$

The oscillator has the frequency ω_0 and initial negative damping coefficient a. $\mu f(x, \dot{x})$ is the non-linearity in the oscillator. The external periodic signal is assumed to be sinusoidal having the frequency ω and amplitude E.

There is no analytical method for obtaining the accurate solution of the above equation. However, if a and μ are assumed to be small, approximate solutions may be obtained starting with a periodic solution and choosing its phase and amplitude so as to satisfy the equation. The starting solution has to be chosen from physical considerations.

In general, the oscillator is expected to execute a forced oscillation of frequency ω and a

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free oscillation the frequency of which will be close to ω_0. When ω is approximately equal to $n\omega_0$, where n is an integral ratio, it is observed that the free oscillation frequency is modified to ω/n. The locking of the frequency of free oscillation to an integral multiple of that of the external signal is the resonance phenomenon and the range of the external frequency over which it occurs is the zone of resonance.

Thus when $\omega \cong \omega_0/n$, the starting periodic solution may be taken to be

$$x = A \sin(\omega t+\phi)+A_n \sin(n\omega t+\phi_n) \quad \dots\dots\dots(2)$$

The first term represents the forced oscillation and the second the modified free oscillation. In general, A, ϕ, A_n and ϕ_n are time varying functions. But there may be equilibrium states for which they attain constant values.

Differentiating (2) twice:

$$\dot{x} = \omega A \cos(\omega t+\phi)+\dot{A}\sin(\omega t+\phi)+$$
$$+A\dot{\phi}\cos(\omega t+\phi)+n\omega A_n \cos(n\omega t+\phi_n)+$$
$$+\dot{A}_n \sin(n\omega t+\phi_n)+A_n\dot{\phi}_n \cos(n\omega t+\phi_n)$$

$$\ddot{x} = -\omega^2 A \sin(\omega t+\phi)+2\omega\dot{A}\cos(\omega t+\dot{\phi})-$$
$$-2\omega A\dot{\phi}\sin(\omega t+\phi)-n^2\omega^2 A_n \sin(n\omega t+\phi_n)+$$
$$+2n\omega\dot{A}_n \cos(n\omega t+\phi_n)-2n\omega A_n\dot{\phi}_n \sin(n\omega t+\phi_n).$$

Second order terms like $\ddot{A}$, $\ddot{\phi}$ have been neglected since a and μ are assumed to be small.

Substituting in eqn. (1), neglecting second order terms and equating the co-efficients of like terms on the two sides:

$$2\omega A\dot{\phi} = -E\cos\phi+(\omega_0^2-\omega^2)\,A+a_1 \quad \dots\dots\dots(3a)$$

$$2\omega\dot{A} = -E\sin\phi+a\,\omega A-b_1 \quad \dots\dots\dots(3b)$$

$$2n\omega A_n\,\dot{\phi}_n = (\omega_0^2-n^2\omega^2)\,A_n+a_n \quad \dots\dots\dots(3c)$$

$$2n\omega\dot{A}_n = an\omega A_n-b_n \quad \dots\dots\dots(3d)$$

where

$$a_1 = \frac{1}{\pi}\int_{-\pi}^{\pi}\mu\,\mathrm{f}(\quad)\sin(\omega t+\phi)\,\mathrm{d}(\omega t)$$

$$b_1 = \frac{1}{\pi}\int_{-\pi}^{\pi}\mu\,\mathrm{f}(\quad)\cos(\omega t+\phi)\,\mathrm{d}(\omega t)$$

$$a_n = \frac{1}{\pi}\int_{-\pi}^{\pi}\mu\,\mathrm{f}(\quad)\sin(n\omega t+\phi_n)\,\mathrm{d}(n\omega t)$$

$$b_n = \frac{1}{\pi}\int_{-\pi}^{\pi}\mu\,\mathrm{f}(\quad)\cos(n\omega t+\phi_n)\,\mathrm{d}(n\omega t)$$

$$\mathrm{f}(\quad) = \mathrm{f}\,[A\sin(\omega t+\phi)+A_n\sin(n\omega t+\phi_n),\ \omega A\cos(\omega t+\phi)+n\omega A_n\cos(n\omega t+\phi_n)]$$

In the equilibrium state $\dot{A} = \dot{\phi} = \dot{A}_n = \dot{\phi}_n = 0$. Hence

$$E\cos\phi = (\omega_0^2-\omega^2)A+a_1 \quad \dots\dots\dots(4a)$$

$$E\sin\phi = a\,\omega A-b_1 \quad \dots\dots\dots(4b)$$

$$(\omega_0^2-n^2\omega^2)A_n = -a_n \quad \dots\dots\dots(4c)$$

$$an\omega A_n = b_n \quad \dots\dots\dots(4d)$$

For ultra or subharmonic resonance $(\omega_0^2-\omega^2)$. A is much larger than the other terms in eqns. (4a) and (4b); hence A is approximately given by $\dfrac{E}{\omega_0^2-\omega^2}$. Thus for obtaining A_n and ϕ_n, only the last two equations have to be solved. Plots of A_n, obtained on solving the equations, against $\dfrac{\omega_0^2-n^2\omega^2}{n\omega}$ give the theoretical resonance curves. All parts of the resonance curves do not represent physical solution since for a physical solution it is required that the equilibrium state $\dot{A}_n = \dot{\phi}_n = 0$ be stable. $\dot{A}$, $\dot{\phi}$ may be assumed to be zero to the first order of approximation.

It is known[5] that the equilibrium state $\dot{A}_n = \dot{\phi}_n = 0$ is stable if

$$\frac{\partial\dot{A}_n}{\partial A_n}+\frac{\partial\dot{\phi}_n}{\partial\phi_n}<0 \quad \dots\dots\dots(5a)$$

and

$$\frac{\partial\dot{A}_n}{\partial A_n}\cdot\frac{\partial\dot{\phi}_n}{\partial\phi_n}-\frac{\partial\dot{A}_n}{\partial\phi_n}\cdot\frac{\partial\dot{\phi}_n}{\partial A_n}>0 \quad \dots\dots\dots(5b)$$

The differentials are to be evaluated at the equilibrium values of A_n and ϕ_n.

Details of the resonance curves may be obtained only when $\mu\mathrm{f}(x, \dot{x})$ is known. Let it be considered that $\mu\mathrm{f}(x, \dot{x}) = b\dot{x}^2+c\dot{x}^3$. It represents the type of non-linearity that is encountered in a properly adjusted triode oscillator. Evidently, in this case ultraharmonic resonance of orders 2 and 3 and subharmonic resonance of order $\frac{1}{2}$ and $\frac{1}{3}$ may be obtained.

2.1. *Ultraharmonic Resonance Curves*

Ultraharmonic resonance of order 3: It can be shown that when $n = 3$,

$$b_3 = \tfrac{3}{4}c[(3\omega A_3)^2+2(\omega A)^2+\tfrac{1}{3}\frac{(\omega A)^3}{(3\omega A_3)}\cos\Phi]\times(3\omega A_3) \quad \dots\dots\dots(6a)$$

$$a_3 = \tfrac{1}{4}c(\omega A)^3 \sin \Phi \qquad \text{.........(6b)}$$

where $$\Phi = \phi_3 - 3\phi \qquad \text{.........(6c)}$$

On substituting in eqns. (3c) and (3d) and putting

$$\frac{4a}{3c} = \rho_0^2, \frac{3\omega A_3}{\rho_0} = \rho_3, \frac{\omega A}{\rho_0} = \rho \text{ and } \frac{\omega_0^2 - 9\omega^2}{a\ 3\omega} = \delta_3$$

$$2\dot{\rho}_3 = a[1 - \rho_3^2 - 2\rho^2 - \tfrac{1}{3}\frac{\rho^3}{\rho_3}\cos \Phi]\,\rho_3 \qquad \text{.........(7a)}$$

$$2\dot{\phi}_3 = a[\delta_3 + \tfrac{1}{3}\frac{\rho^3}{\rho_3}\sin \Phi] \qquad \text{.........(7b)}$$

The equilibrium value of ρ_3 is, therefore, given by

$$\left(\frac{\rho^3}{3\rho_3}\right)^2 = \delta_3^2 + (1 - 2\rho^2 - \rho_3^2)^2 \qquad \text{.........(8)}$$

The conditions of stability as obtained from the inequalities (5a) and (5b) are

$$\rho_3^2 + \rho^2 > 0{\cdot}5 \qquad \text{.........(9a)}$$

$$3\rho_3^4 - 4\rho_3^2(1 - 2\rho^2) + (1 - 2\rho^2)^2 + \delta_3^2 > 0 \qquad \text{.........(9b)}$$

The signal producing resonance has the amplitude $\frac{1}{3}\rho^3$ which is determined by E only. Curves for ultraharmonic resonance have, therefore, characteristics similar to those of harmonic resonance Three cases may be clearly distinguished. For $\rho < 0{\cdot}5864$, though ρ_3 is triple valued for the low values of δ_3, only the highest value is stable. As δ_3 is increased from zero ρ_3^2 decreases and becomes critical when the tangent to the resonance curve is vertical. For further increase in δ_3, $\frac{dA_3}{d\phi_3}$ has an unstable singularity, hence ultraharmonic resonance is not possible but limit cycles in the A_3-ϕ_3 plane may be exhibited. For $0{\cdot}5864 < \rho < 0{\cdot}6067$, two of the three possible values of δ_3 are stable and therefore jump phenomenon and the associated hysteresis[3] may be exhibited. For $\rho > 0{\cdot}6067$, resonance curves are single valued and stable harmonic resonance is exhibited up to a value of δ_3 up to which inequality (9a) is satisfied. However, for $\rho > 0{\cdot}707$ ultraharmonic oscillation is possible for all values of δ_3. In fact, it should be regarded as the ultraharmonic of the forced oscillation, which suppresses the free oscillation. The features noted above are illustrated by the resonance curves shown in Fig. 1 in which ρ_3^2 is plotted against δ_3 with ρ as parameter. The unstable parts of the resonance curves are shown by dotted lines. The lower branch of the curves for $\rho < 0{\cdot}58$ are not shown.

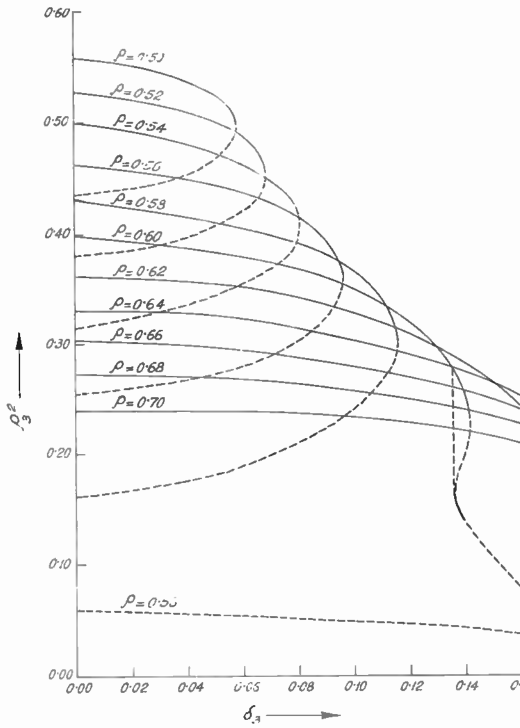

Fig. 1. Ultraharmonic resonance curves of order three.

It should be noted that the resonance curves in this case have the distinctive feature that ρ_3^2 decreases with increase in the external signal. This arises from the fact that with increase in E, ρ increases; thus though the signal producing resonance is increased, the damping for the ultraharmonic oscillation is also increased. It is also observed that ρ_3^2 is always less than 1 in the resonance condition; hence the power available for ultraharmonic oscillation is less than that for free oscillation.

Ultraharmonic resonance of order 2: In this case the Fourier coefficients are:

$$b_2 = a\rho_0[\rho_2^2 + 2\rho^2 + \frac{2b}{3\rho_0 c}\,\frac{\rho^2}{\rho_2}\cos \Phi]\rho_2 \qquad \text{......(10a)}$$

$$a_2 = a\rho_0[\frac{2b}{3\rho_0 c}\,\rho^2 \sin \Phi] \qquad \text{.........(10b)}$$

where

$$\Phi = \phi_2 - 2\phi,\ \rho_2 = \frac{2\omega A}{\rho_0},\ \rho_0^2 = \frac{4a}{3c}, \frac{\omega A}{\rho_0} = \rho$$

Substituting in eqns. (3c) and (3d) and putting

$$\frac{\omega_0^2 - 4\omega^2}{a.\,2\omega} = \delta_2$$

$$2\dot{\rho}_2 = a\left[1-\rho_2^2-2\rho^2-\frac{2b}{3\rho_0 c}\frac{\rho^2}{\rho_2}\cos\Phi\right]\rho_2 \qquad \text{.........(11a)}$$

$$2\dot{\phi}_2 = a\left[\delta_2+\frac{2b}{3\rho_0 c}\frac{\rho^2}{\rho_2}\sin\Phi\right] \qquad \text{.........(11b)}$$

The equilibrium values of ρ_2 are given by

$$\left(\frac{2b}{3\rho_0 c}\frac{\rho^2}{\rho_2}\right)^2 = \delta_2^2+(1-\rho_2^2-2\rho^2)^2 \qquad \text{.........(12)}$$

The equilibrium is stable if

$$\rho_2^2+\rho^2 > 0{\cdot}5 \qquad \text{.........(13a)}$$

and

$$3\rho_2^4-4\rho_2^2(1-2\rho^2)+(1-2\rho^2)^2+\delta_2^2>0 \qquad \text{......(13b)}$$

Plots of ρ_2^2 against δ_2 with ρ as a parameter for $\frac{2b}{3\rho_0 c}=1$ are shown in Fig. 2. The characteristics are identical to those of the previous case. However, the signal causing the resonance has the amplitude $\frac{2b}{3\rho_0 c}\rho^2$ instead of $\frac{\rho^3}{3}$. Hence the zone of resonance as well as the maximum power at the ultraharmonic frequency for a fixed value of ρ may be controlled by adjusting the asymmetry in the non-linearity.

2.2. *Subharmonic Resonance Curves*

Subharmonic resonance of order 1/3: On evaluating a_n and b_n for $n=\frac{1}{3}$ and substituting

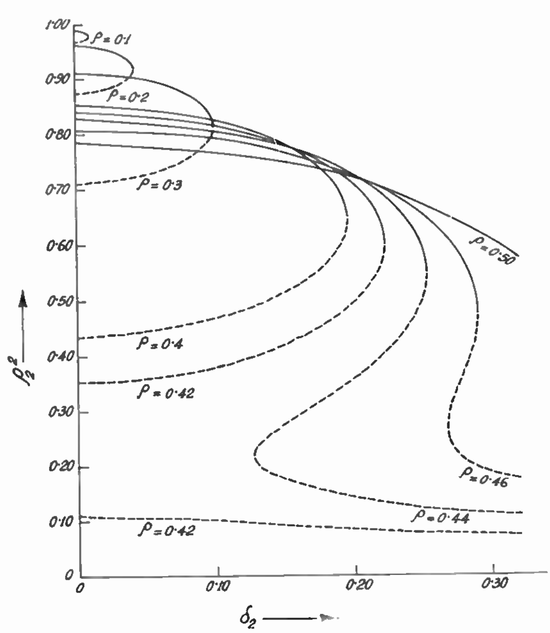

Fig. 2. Ultraharmonic resonance curves of order two.

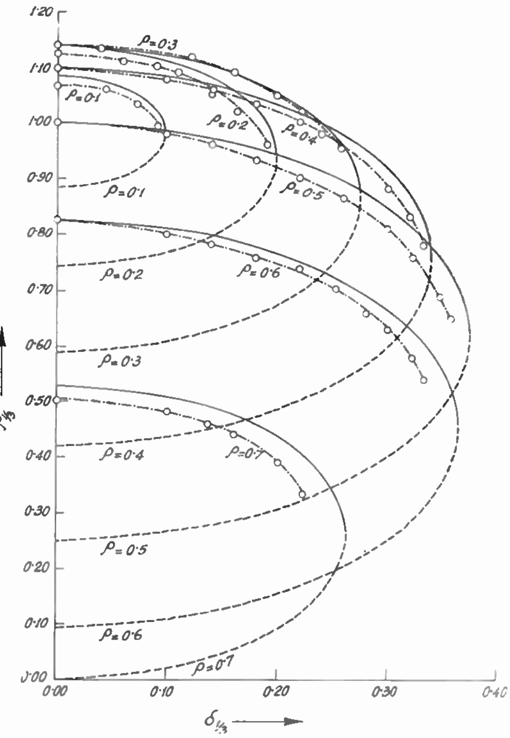

Fig. 3. Subharmonic resonance curves of order one-third.

in eqns. (3c) and (3d), one gets

$$2\dot{\rho}_{\frac{1}{3}} = a\left[1-\rho_{\frac{1}{3}}^2-2\rho^2-\rho\rho_{\frac{1}{3}}\cos\Phi\right]\rho_{\frac{1}{3}} \qquad \text{........(14a)}$$

$$2\dot{\phi}_{\frac{1}{3}} = a\left[\delta_{\frac{1}{3}}-\rho\rho_{\frac{1}{3}}\sin\Phi\right] \qquad \text{.........(14b)}$$

where

$$\Phi = \phi-3\phi_{\frac{1}{3}},\ \delta_{\frac{1}{3}} = \frac{\omega_0^2-\omega^2/9}{a\ \omega/3},\ \rho_{\frac{1}{3}} = \frac{\omega/3\ A_{\frac{1}{3}}}{\rho_0}$$

and, as before, $\rho_0^2=\frac{4a}{c}$, $\rho=\frac{\omega A}{\rho_0}$

Equilibrium values of ρ are given by

$$(\rho\rho_{\frac{1}{3}})^2 = \delta_{\frac{1}{3}}^2+(1-\rho_{\frac{1}{3}}^2-\rho^2)^2 \qquad \text{.........(15)}$$

$$\rho_{\frac{1}{3}}^2 = 1-\frac{3}{2}\rho^2\pm\sqrt{\rho^2-\frac{7}{4}\rho^4-\delta_{\frac{1}{3}}^2} \qquad \text{........(16)}$$

Conditions of stability of the equilibrium are

$$\rho_{\frac{1}{3}}^2+\rho^2>0{\cdot}5 \qquad \text{.........(17a)}$$

$$\sqrt{\rho^2-\frac{7}{4}\rho^4-\delta_{\frac{1}{3}}^2}\ \pm\ (1-\frac{3}{2}\rho^2)>0 \qquad \text{.........(17b)}$$

Plots of $\rho_{\frac{1}{3}}^2$ against $\delta_{\frac{1}{3}}$ with ρ as a parameter are shown in Fig. 3. For a fixed value of ρ, $\rho_{\frac{1}{3}}$ decreases with increase in $\delta_{\frac{1}{3}}$ till the tangent to the curve becomes vertical. With further increase in $\delta_{\frac{1}{3}}$ subharmonic oscillation vanishes and for $\rho^2 > 0{\cdot}5$ the oscillation occurs only at the external frequency while for $\rho^2 < 0{\cdot}5$ a combination oscillation for which the phase and amplitude vary over a cycle is executed. It should be noted that outside the zone of resonance $\frac{dA_{\frac{1}{2}}}{d\phi_{\frac{1}{2}}}$ does not have any singularity. Hence no limit cycle in the $A_{\frac{1}{2}}$ -$\phi_{\frac{1}{2}}$ plane is possible and the resultant oscillation is the combination of the free and harmonic ones. Resonance curves in this case have also the distinctive feature that the zone of resonance at first increases with increase in E, attains a maximum value and then decreases to zero. With increase in E, ρ increases; the increase of ρ at the same time decreases $\rho_{\frac{1}{2}}$. Hence the amplitude of the resonating signal which is equal to $\rho\rho_{\frac{1}{2}}{}^2$, at first increases but ultimately decreases to zero, the decreasing effect of $\rho_{\frac{1}{2}}$ offsetting the increasing effect of ρ.

Thus, it is evident that subharmonic resonance may be exhibited only within a limited range of ρ; for ultraharmonic or harmonic resonance there is no such limit. Further, in subharmonic resonance there is a maximum value of the zone of resonance which is $0{\cdot}286a$ and is dependent only on the initial damping. It may be seen that in the other two cases the zone of resonance increases with increase in the external signal. Hysteresis or the jump within the zone of resonance is also not possible in the subharmonic resonance of order 1/3. But sudden jump from a finite value of $\rho_{\frac{1}{2}}$ to zero on the border of the zone of resonance is possible, the jump being reversible. Maximum power available at the subharmonic frequency is greater than that for free oscillation for $\rho < 0{\cdot}5$.

Subharmonic resonance of order 1/2: On evaluating the Fourier coefficients and substituting in eqns. (3c) and (3d), one gets

$$2\dot{\rho}_{\frac{1}{2}} = a[1 - \rho_{\frac{1}{2}}{}^2 - 2\rho^2 - \frac{4b}{3\rho_0 c}\,\rho\cos\Phi]\,\rho_{\frac{1}{2}} \qquad \text{.........(18a)}$$

$$2\dot{\phi}_{\frac{1}{2}} = a[\delta_{\frac{1}{2}} - \frac{4b}{3\rho_0 c}\sin\Phi] \qquad \text{.........(18b)}$$

where $\Phi = \phi - 2\phi_{\frac{1}{2}}$, $\rho = \frac{\frac{1}{2}\omega A_{\frac{1}{2}}}{\rho_0}$, $\delta_{\frac{1}{2}} = \frac{\omega_0{}^2 - \frac{1}{4}\omega^2}{\frac{1}{2}a\omega}$,

and $\rho_0{}^2 = \frac{4a}{3c}$, $\rho = \frac{\omega A}{\rho_0}$ as before.

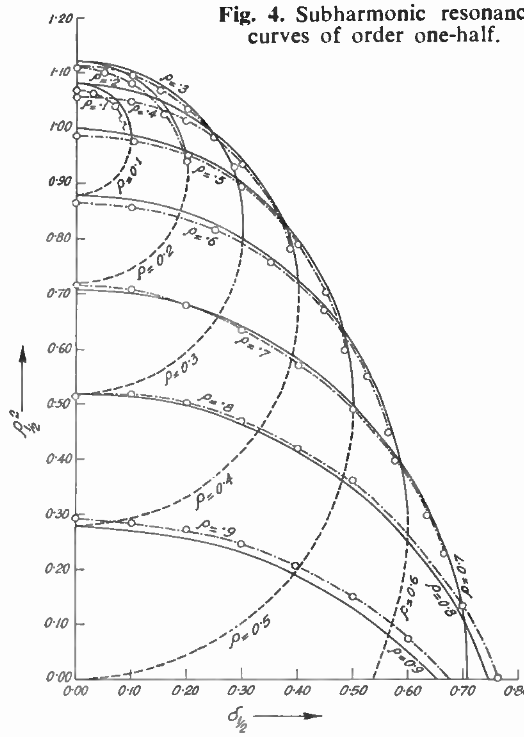

Fig. 4. Subharmonic resonance curves of order one-half.

The equilibrium values of $\rho_{\frac{1}{2}}$ are given by

$$\rho_{\frac{1}{2}}{}^2 = 1 - 2\rho^2 \pm \sqrt{\left(\frac{4b}{3\rho_0 c}\,\rho\right)^2 - \delta_{\frac{1}{2}}{}^2} \qquad \text{.........(19)}$$

The conditions of stability are

$$\rho^2 + \rho_{\frac{1}{2}}{}^2 > 0{\cdot}5 \qquad \text{.........(20a)}$$

$$\rho_{\frac{1}{2}}{}^2 + 2\rho^2 - 1 > 0 \qquad \text{.........(20b)}$$

Plots of $\rho_{\frac{1}{2}}{}^2$ as obtained from eqn. (19) are shown in Fig. 4 for $\frac{4b}{3\rho_0 c} = 1$, the stable values being indicated by full lines.

The characteristics of the curves are similar to those for subharmonic resonance of order 1/3. For $\rho^2 < 0{\cdot}5$, with increase in $\delta_{\frac{1}{2}}$, $\rho_{\frac{1}{2}}{}^2$ decreases and subharmonic oscillation vanishes when the tangent to the curve becomes vertical. But for $\rho^2 > 0{\cdot}5$, the resonance curves do not have a vertical tangent, $\rho_{\frac{1}{2}}{}^2$ decreases gradually to zero with increase in $\delta_{\frac{1}{2}}$. The zone of resonance increases with increase in ρ, attains a maximum value and decreases to zero. Also, the resonance is possible only within a limited range of ρ. No jump phenomenon, however, is possible. The maximum zone of resonance is dependent not only on a but also on b/c.

3. Experimental Resonance Curves

Experimental observations for verifying the resonance curves shown in Figs. 1-4 have been made with the help of an electronic differential analyser. The set-up of the analyser for solving eqn. (1) is shown in Fig. 5. The non-linear function was generated by a diode function generator. Different values of δ were set by varying ω_0^2, the external signal frequency being kept fixed and so adjusted that for $\omega_0^2 = 1$; $\delta = 0$. For measuring ρ_n^2, subharmonic or ultraharmonic oscillations were separated from the forced oscillations by combining properly the outputs at A and B (Fig. 5) representing respectively $\dot{x}$ and x. The mean square value was then obtained with the help of a thermocouple and a d.c. millivoltmeter.

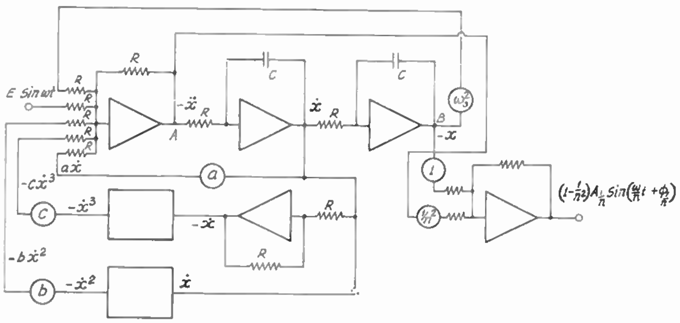

Fig. 5. Set-up of the Differential Analyser for obtaining the resonance curves.

Experimental points for subharmonic resonance are shown by circles on the theoretical curves of Figs. 3 and 4. It will be seen that the experimental results agree quite closely with the theoretical curves. The small discrepancy that exists may be accounted for by errors in the setting of ω_0^2.

On the other hand, ultraharmonic resonance curves obtained experimentally showed only rough agreement with the theoretical curves. A probable cause of the departure may be the effect of the sum and difference frequency terms, the contributions of which were neglected in deriving the theoretical curves. In ultraharmonic resonance the sum and difference frequencies are quite close to the frequency of free oscillation and therefore may have large effect. In subharmonic resonance, however, these frequencies are far removed from the frequency of free oscillation and have little effect.

4. Conclusion

The above analysis revealed the following distinguishing features of ultraharmonic and subharmonic resonance. The characteristic features of ultraharmonic resonance are: (a) zone of resonance increases with increase in the external signal; (b) jump and associated hysteresis effect may be exhibited; (c) amplitude of the ultraharmonic oscillation decreases with increase in the external signal; (d) the maximum power available for the ultraharmonic oscillation is less than that for free oscillation. Characteristic features of subharmonic resonance are: (a) subharmonic resonance occurs for a limited range of the amplitude of the external signal; (b) zone of resonance has a maximum value; (c) jump and hysteresis effect is not possible in the zone of resonance; (d) the amplitude of the sub-harmonic oscillation at first increases and then decreases with increase in the external signal; (e) maximum power available for subharmonic oscillations may be greater than that for free oscillation.

In addition it may be noted that the experimental subharmonic resonance curves agree closely with the theoretical curves while for ultraharmonic resonance the agreement is not so good.

5. Acknowledgments

The author is deeply indebted to Professor J. N. Bhar for his constant guidance during the progress of the work.

6. References

1. B. van der Pol, "Forced oscillation in a circuit with non-linear resistance," *Philosophical Magazine*, **3**, p. 65, 1927.
2. D. G. Tucker, "Forced oscillations in oscillator circuits and the synchronization of oscillators," *J. Instn Elect. Engrs*, **93**, Part III, p. 226, 1946.
3. M. L. Cartwright, "Forced oscillations in nearly sinusoidal systems," *J. Instn Elect. Engrs*, **95**, Part III, p. 88, 1948.
4. A. W. Gillies, "The application of power series to the solution of non-linear circuit problems," *Proc. Instn Elect. Engrs*, **96**, Part III, p. 53, 1949.
5. N. Minorsky, "Introduction to Non-linear Mechanics," pp. 321-340. (Edward Brothers, Inc., Ann Arbor, Michigan, 1947.)

Photo-Electric Image Techniques in Astronomy†

by

B. V. SOMES-CHARLTON‡

A paper read at the 1959 Convention in Cambridge on 4th July.

Summary: The underlying scientific principles of light detectors such as the human eye, photographic emulsion or photo-electric device, are explained and their quantum efficiencies compared in terms of signal/noise ratio. Television techniques can be applied to aid astronomical observations and help to overcome, in some measure, the fundamental problems in observing celestial objects through the semi-transparency of the Earth's atmosphere. The television techniques in the detection of threshold and extremely low light level stellar and planetary images are described. Achievements to date are summarized and the results illustrated by photographs of the Moon and of Mars obtained by the author in 1956. On the latter occasion a low light-level television system was used to record pictures showing the fine surface structure of the planet. The paper concludes with a survey of proposals for the construction of a space orbiting astronomical telescope.

1. Introduction and Historical Outline

The application of photo-electric principles to quantitative measurements in astronomy commenced in 1910 when the American astronomer Stebbins used the selenium cell in his new photometer and with this device discovered the secondary eclipse of the binary star Algol. The accuracy of this method for obtaining stellar magnitudes§ (brightness values) far surpassed that of visual and photographic methods. Since that time. there has been a steady increase in sensitivity and spectral range of photo-electric cells and in 1946 the photometer employing a photomultiplier cell used on the 100-inch and 200-inch telescopes at the Mount Wilson and Palomar Observatories showed that the sensitivity of the device had advanced by a factor of 16, i.e. from magnitude 16 to magnitude 19.

A few years later the limit of sensitivity was further extended to stars of the 23rd magnitude using the 200-inch telescope at Mt. Palomar with the photo-multiplier cell in an electronic device known as a "pulse-counting photometer" which virtually counts the incident light quanta. This method was first adopted in astronomy by W. Blitzstein in America and by G. G. Yates at Cambridge and has since been used to determine the luminosity of the faintest stars ever recorded, namely, 23rd magnitude which represents a brightness range of about one thousand million to one.

The disadvantage of the photo-electric cell technique as compared with photography for recording very faint stars is that only one measurement can be made at a time whereas a single photograph can record several stellar magnitudes simultaneously even though an exposure time of many hours may be necessary.

To overcome this great disadvantage, and in order to take advantage of the merits of the photo-electric principle of light detection, experiments with electronic image devices have been conducted. As early as 1934, the French astronomer M. Lallemand at the Observatoire de Paris, commenced work on a

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‡ High-Definition Television Ltd., St. Andrews Road, Cambridge.

U.D.C. No. 621.397: 522.6

§ The magnitude sequence used to define apparent or absolute brightness of a star is based on a number of steps (n) bearing a constant logarithmic ratio of 2·512. The order of magnitude (2·512) is expressed as a positive, or negative (if a star is brighter than zero magnitude) whole number or decimal.

photo-electric image-converter tube capable of recording the image of an entire star field or the disc of a planet using an electron-sensitive photographic film placed inside a vacuum tube.

In 1951, the author, in collaboration with P. B. Fellgett of the Cambridge Observatories, carried out quantitative tests at the telescope using a television camera in order to ascertain the possible advantages, if any, of using an image orthicon in a standard television system for reproducing and recording the complete image of a planetary object in conjunction with photography. Encouraging results were obtained using the Moon as the subject. These tests were accompanied by a considerable amount of theoretical speculation regarding the relative sensitivities in the case of the image orthicon versus photography, and Fellgett carried out prolonged investigations with manufacturers of photographic materials in order to obtain and correlate data on the performance of various types of photographic emulsion. Similar research work was conducted in an attempt to collect performance data on photo-electric devices and ordinary photo-cathodes so that some comparisons could be made with photographic films and plates.

Following the series of practical tests between 1951 and 1955 on the Solar (Tunnel) telescope and the 28-inch Newall refractor at the Cambridge Observatories, sufficient evidence was available on the use of television techniques for consideration by a representative body of astronomers. It was therefore decided that a closed-circuit demonstration should be given at the 9th General Assembly of the International Astronomical Union held in Dublin in 1955, where a sub-commission on photo-electric image tubes would be meeting to discuss their development for astronomical applications. Television pictures of the Moon were produced by an image orthicon camera attached to the 12-inch refractor of the Dunsink Observatory for the purpose of the demonstration. During a meeting of the General Assembly a resolution was passed by delegates for the formation of a sub-commission to investigate the use of television techniques, thereby giving formal recognition of their potential value to astronomy.

Early in 1956 laboratory tests were conducted at Cambridge to obtain further comparisons in sensitivity at very low light levels between the image orthicon and a super-sensitive film emulsion which provided reasonably conclusive evidence of a factor of 3 gain in favour of the television camera. This conclusion led to a visit by the author to Bloemfontein, South Africa, for observations and recordings of the planet Mars during the 1956 " perigee opposition ", when it was at a distance of 35 million miles from the Earth. Experiments with television methods have also been undertaken in America at the Lowell Observatory, Arizona, and more recently in Russia at the Leningrad Observatory. Likewise, work on image converters of the Lallemand type, is proceeding at the Mt. Palomar Observatory, U.S.A. and the Observatoire de Paris. A new department of Instrument Technology under Prof. J. D. McGee at the Imperial College of Science, London, has been established to develop special photo-electric image tubes for astronomical applications.

2. Scientific Background

To understand and fully appreciate the roles that television and photo-electric image techniques may play in observational astronomy, it is necessary to have some knowledge of the under-lying problems confronting astronomers, and the theoretical considerations. Optical astronomers are faced with three main problems:

(a) To collect and detect light from extremely faint stars and nebulae against the night-sky threshold brightness.

(b) To obtain from telescopes and detectors sufficient resolution to clearly resolve detailed structure of planets and stars and to identify separately two stars which are very close together.

(c) To overcome the disadvantage of viewing through a semi-transparent atmosphere which is in a state of continual turbulence and spoils image definition.

In the past, the first and second problems have led to the policy of building larger telescopes, culminating in the construction of the

200-inch reflecting instrument at Mt. Palomar in California. The third problem presents a challenge which might be solved by television techniques in conjunction with the photographic methods using super-sensitive emulsions—failing this there may be no alternative to placing telescopes outside the dense region of the Earth's atmosphere by means of balloons, rockets or even satellites. An American project already exists for launching a space orbiting telescope as a robot completely under the control of a "command ground observatory", and a British proposal exists for a satellite-controlled telescope.

However, the photo-electric methods have, so far, shown promising results and the ways in which they have been achieved will be described later on in this paper.

The theoretical considerations have been based on the behaviour of light and the effects of quanta on different types of photo-sensitive surfaces, such as the retina of the human eye, photographic emulsion or photo-electric device.

2.1. *Quantitative Astronomical Measurements*

It is known that photons do not arrive at the surface of a detector at a steady rate, but randomly distributed in time. It is necessary to collect sufficient numbers so that these fluctuations are averaged if an accurate measurement is to be made. If a measurement has to be made for 10 per cent. accuracy, at least 100 photons will be required and for 1 per cent. accuracy 10,000 photons will be needed. Generally, four times as many photons are necessary to double the accuracy of measurement.

One of the principal aims in building large telescopes is to collect photons of star light from as large an area as possible. A star of zero magnitude sends about 10^6 photons of visible light per second on to each square centimetre of the collecting area of the mirror or object lens of a telescope. A star of such luminosity is easily seen with only the naked eye as a detector but a star of 20th magnitude sends only one photon in a hundred seconds on to each square centimetre and it requires a very large telescope to collect a sufficient number for detection. The 200-inch instrument at Mt. Palomar is able to collect approximately 2,000 photons per second from a 20th magnitude star.

2.2. *Performances of Light Detectors*

Since the extent and value of the information received from the universe depends upon the number of photons collected and detected, it is surprising that 99.9 per cent. entering the telescope are wasted. This is due to the fact that only a small fraction of the total number of photons available can be effectively used to produce a recognizable effect on the detector. Ideally, each incident photon should give rise to the reduction of one grain of silver halide in an emulsion or liberate an electron from a photo-cathode. The nearest approach to this quantum efficiency of unity is made by the photo-multiplier tube. Yet, even this type of detector requires 10 to 20 photons to eject the electron. Photographic emulsion is still less efficient as only 0·1 to 1 per cent. of the incident photons are effectively used. Moreover there is an additional serious loss in efficiency when very long exposures are made as is necessary in the case of recording light from faint stars or nebulae. The efficiency of a photocell does not depend on the rate of arrival of photons and the same is probably true of the fundamental processes whereby an emulsion reacts to light, but the crystal lattice of the photographic emulsion grains is continually excited by its own heat in a way exactly similar to the effect produced by light. If a commercial film or plate, normally used for relatively short exposures is to be successfully stored between manufacture and use, it must be protected in some way from such thermal excitations. This protection takes the form of an effect known as "reciprocity failure" whereby an emulsion is made less sensitive to excitations at a slow rate. This failure of the reciprocity law at low photon velocities means that as exposure time is increased to record faint star images there is no corresponding blackening of the photographic grains and with fast emulsions saturation and consequent overlapping of reduced silver grains is reached within a very short exposure time. "Reciprocity failure" is a serious disadvantage in astronomy where exposures may take several hours or a whole night's work. Astronomers therefore use emulsions in which "reciprocity failure" is minimized in manufacture and consequently plates thus treated have to be stored under refrigeration. Photo-electric materials do not, however, suffer in this way.

It is, therefore, in an effort to overcome the gross wastage of photons collected by the telescope that the possible uses of television principles are being considered by astronomers. Nevertheless, it is only within the last decade or so that systematic investigation of the fundamental limitations of radiation detectors and quantitative measurements of their sensitivities have revealed the great need to improve efficiency. Photocells and photomultipliers have already shown a marked improvement, but whereas a photographic plate can record the whole field of a telescope image simultaneously, a photocell can only be used on one star at a time. This method is obviously too laborious and the solution lies in the development of a device combining the quantum efficiency of the photocell with the complete image forming properties of photography. If an efficiency of 10 per cent., i.e. 100 times greater than photography, could be obtained with such a device and applied to the 200-inch telescope it would give it the light gathering power of a 2,000 inch; the construction of such a large telescope is beyond the bounds of possibility. Two photoelectric devices bring such a gain within the realms of possibility, namely, a special type of image converter and the television type of image tube.

2.3. *Quantitative Sensitivity Measurements*

In order to discuss why television techniques and image converters may give increased sensitivity as light receptors placed at the prime focus of a telescope, it is necessary to define the term sensitivity. It might appear that sensitivity can be adequately defined in terms of the faint-limit of a device but consideration shows that this approach is too elementary to take account of the diversity of needs in astronomical work. A more satisfactory viewpoint may be gained by considering these requirements in relation to the fundamental limits set by the quantum nature of light.

The problem is related to that referred to under (a) in Section 2 in cases where a measurement has to be made on faint starlight against the night-sky background illumination caused by scattered light and the "airglow" of the upper atmosphere. This effect produces about as much light per square minute of arc as a star of 12th magnitude. As a result, the main requirement of a photo-electric detector is the ability to show up low contrast images against a relatively high threshold brightness. A theoretical explanation is given by the following mathematical treatment.

Suppose that n_s photons per second are available for time t from the source to be measured and that n_b per second are received from the background (sky brightness) against which the desired source is observed. The statistical fluctuation Δ_n in the total number of quanta received is given by

$$\overline{\Delta}_n = (n_s + n_b)t \qquad \ldots\ldots\ldots(1)$$

or very nearly, provided the radiation is not intense. The mean squared signal-to-noise ratio in the radiation available is therefore

$$R_1 = \frac{n_s{}^2 t^2}{\Delta_n{}^2} = \frac{n_s{}^2 t}{n_s + n_b} \qquad \ldots\ldots\ldots(2)$$

If this radiation is, for example, measured photo-electrically by a cell that on an average liberates one electron of charge q for every $1/\Sigma$ quanta incident ($\Sigma \ll 1$) and the charge liberated is magnified M times, as in the case of an electron multiplier of a photocell or a television image tube of the image orthicon type, then the charge due to the signal n_s is

$$Q_s = M\Sigma q\ n_s\ t \qquad \ldots\ldots\ldots(3)$$

The statistical fluctuation $\overline{\Delta Q}$ in the total charge collected will be

$$\overline{\Delta Q}^2 = \Sigma t\ M^2\ q^2\ (n_s + n_b) \qquad \ldots\ldots\ldots(4)$$

and the signal-to-noise ratio is given by

$$R\Sigma_1 M = R\Sigma = \frac{Q_s^2}{\Delta Q^2} = \frac{\Sigma n_s^2 t}{(n_s + n_p)} = \Sigma R_1 \ \ldots\ (5)$$

These results can be expressed by stating that R_1 represents the performance of an ideal detector under given conditions, and R that of a detector that is imperfect only in having a quantum efficiency less than unity. It is to be noted that the signal-to-noise ratio is independent of M.

If under the conditions assumed, an actual detector gives a measurement of "signal-to-noise" ratio R, the performance D may be expressed by stating that it has an "Equivalent Quantum Efficiency" Σ_e, where Σ_e is the value of Σ obtained by substituting the value R into eqn. (5):

$$\Sigma_e = \frac{R\,(n_s + n_b)}{n_s^2} = \frac{R}{R_1} \qquad \text{.........(6)}$$

As applied to image detectors this concept is due to A. Rose of America. It has a simple interpretation; suppose that a detector yields a picture of given quality, in other words, the elements of the picture have given signal-to-noise ratios and mutual correlation representing the finite resolution of the picture. The amount of information in the picture and the number N_1 of quanta ideally required to yield this information, can be calculated. If the actual number of quanta available is N then

$$\Sigma_e = \frac{N_1}{N} \qquad \text{.........(7)}$$

that is to say, the detector needs $1/\Sigma_e$ times the brightness level, or $1/\Sigma_e$ times as long exposure as an ideal detector to yield the same picture.

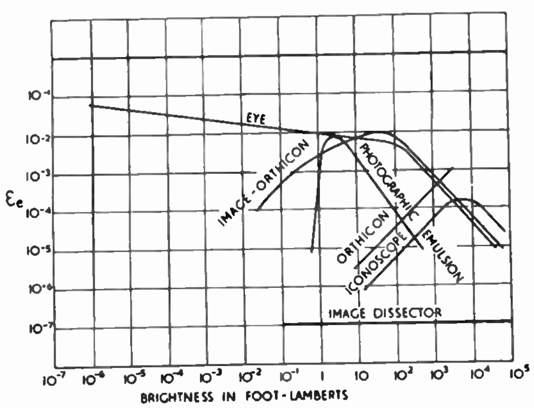


Fig. 1. Equivalent quantum efficiency curves, produced by A. Rose, for the eye, photographic emulsion and television image tube.

In general, Σ_e is a function of the conditions of use, and typical variations, with brightness level have been given by Rose for the human eye, photographic emulsions and television camera tubes. Because of these variations of Σ_e the "faint limit," even if it is precisely defined, gives little indication of the performance at other light levels. The sensitivity of the detector is represented by Rose as a group of curves (Fig. 1) showing the variation of Σ_e with the significant parameters in use, and from these curves the efficiency of the device for a given purpose may be readily obtained. The success of photo-electric methods in astronomy during recent years may be attributed to the availability of photocathodes with quantum efficiencies as high as $\Sigma = 0{\cdot}2$, and the feasibility of using photocells so that the equivalent quantum efficiency Σ_e of the measurement is only slightly less than Σ. A photomultiplier tube can have an Σ_e as high as 0·5.

Equation (5) shows that the signal-to-noise ratio of the photo-current can be controlled only through the quantum efficiency Σ of the photocathode, and is made as favourable as possible by using a cathode with a high efficiency. In the case of a television camera tube such as the Image Orthicon, if the noise in the photo-current (eqn. (5)) can be made to override all other sources of noise, the equivalent quantum efficiency will be nearly equal to ε, and the performance of the tube will be substantially that of its photocathode. There is, however, a loss in effective quantum efficiency in the Image Orthicon due to the target mesh intercepting some of the photo-electrons.

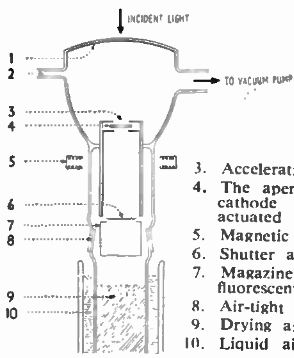
The human eye has a high maximum Σ_e of about 0·05 with the remarkable characteristic of maintaining it over a very wide range of brightness levels, whereas for the image orthicon Σ_e is somewhere between 0·02 to 0·03 but the tube excels the eye in performance at low contrasts. The fastest film Σ_e is about 0·01 or in the order of 2 to 3 times lower than that of the image orthicon. However, it is possible that some of the more recent emulsions have an Σ_e of over 0·1 and if such is case then modern films and photocathodes indeed possess comparable performances.

3. Electronic Imaging Devices

At the 10th General Assembly of the International Astronomical Union held in Moscow in 1958, the Sub-Commission on Photo-electric Image Tubes again discussed their astronomical applications and reported the image tube developments since the last meeting at Dublin in 1955, as a " field of rapidly growing activity ". The discussions were primarily concerned with image converters, employing electronography, and signal-generating image tubes as used in conjunction with standard closed-circuit television systems. The first experiments using an electronic image device for astronomical purposes were made as long ago as 1934 when M. Lallemand of the Observatoire de Paris produced his special image converter tube in which the usual phosphor viewing screen is replaced by a demountable photographic section containing an electron sensitive emulsion surface. With such a device placed at the prime focus of a big telescope, the image of a star can be focused on to the photocathode. The resulting flow of photoelectrons is accelerated to 20 keV (approx.) and magnetically focused on to the electron-sensitive plate or film. For reasons which will be explained in the following section, the tube required to be pumped continuously during operation on the telescope and this, together with other complications, has so far limited the use of the device in astronomical work.

3.1. *The Photographic Image Converter*

It must be acknowledged that with the image converter tube, Lallemand paved the way for the very much more recent experiments in the use of television techniques which were commenced by P. B. Fellgett and the author in 1951. Lallemand's original tube (Fig. 2) consists of a glass envelope with two separate compartments for the photocathode (1) and photographic film (7) respectively which are part of a demountable vacuum system. The formation of the sensitized surface on the inside of the tube face (1) is carried out at the telescope prior to operation by passing caesium vapour through an inlet tube (2) which condenses to produce a thin semi-transparent film. Unfortunately, photocathodes are contaminated by molecules of oxygen and water vapour liberated by photographic emulsions resulting in a loss of sensitivity, and for this reason, the two elements must be isolated until such time as exposure is desired. During the introduction of the caesium vapour through the inlet, the photographic compartment is completely sealed off by closing a magnetically-operated shutter (6) in the neck of the tube. Following the sensitization process, residual molecules of caesium and air are evacuated by continuous pumping. With this tube, stellar images of good definition and contrast are possible, nevertheless, the most significant feature is the 100-fold gain in sensitivity due to the higher quantum efficiency compared with unaided photography and the effective exposure time is reduced accordingly. The quantum efficiency is improved by the high energy given to a photoelectron emitted by the photocathode under the influence of the 20-30 keV acceleration; the energy obtained by a photoelectron being sufficient to produce one developable grain of silver halide in the electron-sensitive emulsion.

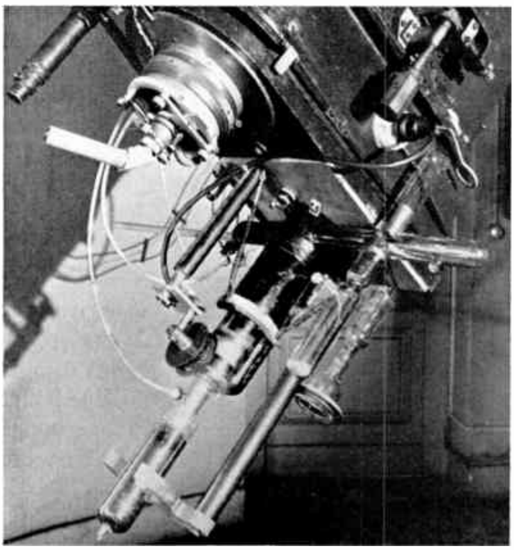


1. Photo-sensitive caesium layer (photocathode) which is semi-transparent.
2. Tube through which caesium is introduced when sensitising the photocathode.
3. Accelerating electron lens.
4. The aperture for isolating the photocathode during sensitising process is actuated by external magnet.
5. Magnetic lens for focusing.
6. Shutter actuated by magnet.
7. Magazine for photographic plates, and fluorescent screen for focusing.
8. Air-tight joint.
9. Drying agent.
10. Liquid air.

Fig. 2. The first type of photographic image converter tube developed by M. Lallemand for intensifying and recording faint star images on an electron sensitive photographic emulsion.

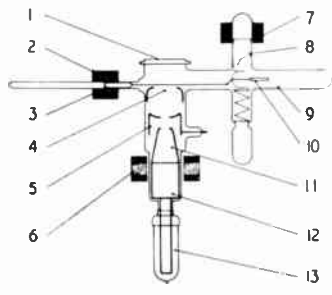
An astronomical plate, Kodak type 103a-o, which is commonly used, requires in the order of 1000 photons to result in one developable grain of silver giving only 0·1 per cent. efficiency, whereas some photocathodes have quantum efficiencies of better than 2 per cent. and electronic image tubes generally about 10 per cent. However, translated into advantages at the telescope, the 100-fold gain over long exposure photography using the photoelectric image technique would give to the 200-inch telescope the light-gathering power of a 2000-inch telescope and a corresponding increase in the case of smaller instruments. M. Lallemand is continuing experiments with image converters and his latest tube (Fig. 3) has many improvements. For example, the photocathode (10) is prefabricated and enclosed in a thin glass capsule (9) as illustrated in the diagram, thereby ensuring complete isolation from the film until an exposure is required. At this point, the capsule is broken by a hammer (8) energized by a solenoid (7) and magnetically drawn into its operative position (4) in the main part of the tube. Nevertheless, the problem of complete isolation is not entirely solved by this method and an alternative is being tried. This involves the introduction of a thin membrane less than 1000 Å thick just in front of the photographic emulsion so as to intercept the flow of gas molecules towards the photocathode during the actual exposure period. this is achieved, there will still be the practical disadvantages of the tube: (i) special servicing of the demountable parts of the tube is necessary each time it is used, (ii) visual identification of a faint star and telescope guiding become difficult, and, (iii) the motion of the telescope with respect to the Earth's magnetic field tends to cause smearing of the electronic image.

A promising method of overcoming the most serious problem of the desensitizing of the photocathode by contamination from the vapours of the film emulsion during a fairly long exposure is being tested by J. D. McGee and his team at the Imperial College of Science, London. As an alternative to the



(*a*)

Fig. 3.(a) The large refracting telescope at the Observatoire de Paris, showing the Lallemand image converter tube in position at the prime focus.



(*b*)

1. Light window or aperture.
2. Solenoid for positioning of photocathode.
3. Iron inertia block.
4. Working position of the photocathode.
5. Electrostatic lenses.
6. Magnetic coil for control of the electron plate holder.
7. Solenoid for operating the hammer.
8. Hammer for breaking the capsule of the photocathode.
9. Capsule for photocathode.
10. Photocathode.
11. Electrostatic lens.
12. Magazine for electron sensitive plate.
13. Liquid air.

Fig. 3.(b) M. Lallemand's image converter designed to reduce the extent of photocathode contamination and resulting loss in sensitivity due to vapours from photographic emulsion.

The membrane reduces the flow of molecules by a factor of 10^8 but efforts are still being made to reduce the flow further. However, until this problem has been satisfactorily overcome, the use of the image-converter combined with electronography will be limited since the 100-fold increase in sensitivity over photography cannot be fully realized in practice where *long* exposures are involved. Assuming image converter with a demountable photographic section for mounting plates and films in a vacuum tube, McGee is experimenting with a more normal type of converter incorporating a phosphor screen for providing an illuminated image. In this method, the phosphor is deposited on the inner surface of a thin mica window 25 mm in diameter which is fused into the end of the tube where the electron image is focused and will withstand atmospheric pressure. A "contact" photograph can be obtained by attaching a film firmly against the outside surface of the window and using the fluorescent light from the phosphor screen for the exposure. However, the resolution is limited by the grain of the phosphor and by the thickness of the mica window, which is about 15 microns. This type of tube, unlike the Lallemand converter which is continuously pumped during operation, develops a spurious background signal caused by residual molecules of caesium from the photocathode process coming into contact with the main chamber of tube. A similar approach to McGee's method has been proposed by Prof. V. I. Krassovsky of the Geographical Institute. Leningrad, using the mica window technique but coating the screen with a "storage" phosphor for integrating the electronic image, and erasing afterwards by irradiation with infra-red light or thermal treatment. Phosphors are known which will store an image for several months and they can "live" with photocathodes without contaminating them by exuding poisonous vapours. Both these methods obviate the need to replace the demountable parts, but the desire for a 100-fold gain in sensitivity over unaided photography cannot be achieved by any of the ordinary image converters without some internal intensification of the image so that each photoelectron can result in one or more reduced grains of silver in the photographic process.

3.2. *Image Intensifiers and Secondary Emission Multipliers*

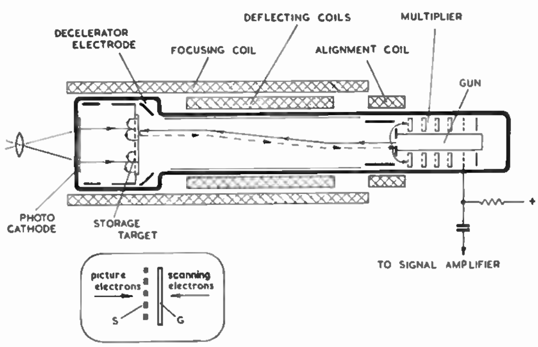
Two methods for intensifying the electronic image are being currently explored. One method involves a series of thin mica or glass membranes each coated with a phosphor on one side and a photoconductive material on the other, thus forming a chain of light amplifiers. Overall light gains as high as 1000 have been obtained with tri-alkali photocathodes. Secondary electron image intensifier tubes have been developed by E. J. Sternglass of the Westinghouse Company. In both of these devices the electrons must be focused from one element to another. One of the difficulties experienced in the image multiplier tube is that the dynode surfaces must not only be good secondary electron emitters but they have to have the property of retarding the speed of the secondaries so that their release energies are low compared with the inter-stage accelerating potentials. If this were not so, the secondaries emitted from one surface at widely differing velocities could not be sharply focused on to the surface of the next dynode. Westinghouse, however, have succeeded in making tubes possessing these essential features and have produced a six-stage multiplier with a gold photocathode for use in an image converter. The tube has an overall gain of greater than 1500 with a resolution of 6 line pairs per mm. Nevertheless, Sternglass and his team have found that the dynode materials when exposed to bright images suffer from fatigue spots resulting in permanent damage to the tube life. The image is virtually "burnt in" in a manner similar to the effect which occurs in the image orthicon tube at the end of its useful life in a television broadcast camera. It is considered that this basically old problem may now be amenable to a solution.

Comparing the resolution of these two methods for a given overall sensitivity, the phosphor-photocathode intensifiers are probably better and Morton of R.C.A. has quoted a figure of 300 to 400 lines if the intensifier were to be incorporated in a modified image orthicon type of tube to be described in more detail in the following section.

3.3. *Signal-generating Image Tubes*

3.3.1. Detection of threshold star images

Comparing television techniques with photographic image converters for aiding or replacing the direct photographic method of storing an image, the main difference between them is that the television camera operates on the principle of vision electrical signal generation using a "pick-up" image tube with a light storage type of target. No corresponding electrical signal is produced by the converter. This characteristic gives the television system a big advantage over the image converters since quantitative measurements in connection with the determination of star field brightness values can be obtained quickly and directly from the video waveform displayed by an oscilloscope. Whereas, using either direct photography or an image converter the resulting negative film of a star field has to be scanned after processing by a microphotometer in order to obtain brightness measurements of individual stars. The use of pulse-counting photometers with the relatively high quantum efficiencies of the photo-multiplier cell are commonly used for magnitude measurements but the process is very laborious and a single observation of a faint star or nebula may require a whole night's work on a big telescope. Yet, the television tube for this kind of astronomical observation is unlikely to be adopted until a higher quantum efficiency comparable with that of a cell is possible. The image orthicon tube with its secondary emission multiplier for improving the signal-to-noise ratio is certainly the most sensitive detector available for use with the television technique, but even so the quantum efficiency in the order of 0·02 to 0·03 is at least ten times lower than that of a photomultiplier cell. This is mainly due to the far greater complexity of the image orthicon and the limitations at low light levels imposed by the inherent " noise " set up by the scanning beam, which produces statistical fluctuations and limits threshold image identification. On the other hand, if the primary photoelectron emission can be substantially increased at low light levels by the introduction of an image intensifier of the phosphor-photocathode type in the image section of the image orthicon tube, the effects of beam or " shot " noise would be greatly minimized. Morton of R.C.A. has proposed such a modification to the standard type of image orthicon (Fig. 4) using a two-stage intensifier section placed behind the primary photocathode. In order to obtain sufficient charge on the target area receiving primary photoelectron emission at low light levels, the photoelectron image current must be amplified by a factor of 200 or 300 prior to reaching the target surface. The proposed intensifier image orthicon consists of two multiplier stages each consisting of a thin mica membrane coated on the image side with an aluminized phosphor layer and on the other side with a photocathode matching the spectral response of the phosphor material. At an acceleration potential of 10 kV between the primary photocathode and first intensifier phosphor layer each incident photoelectron liberates ten electrons from the other side of the screen. Each stage has, therefore, a gain of 10 and this results in the emission of 300 electrons from the actual target of the tube. Morton claims that these intensifier orthicons which have been made in the laboratory for X-ray studies, can produce images of 400 lines definition with a photocathode illumination of about 10^{-6} or 10^{-7} foot candles. Although such tubes have great possibilities for the



(*a*)

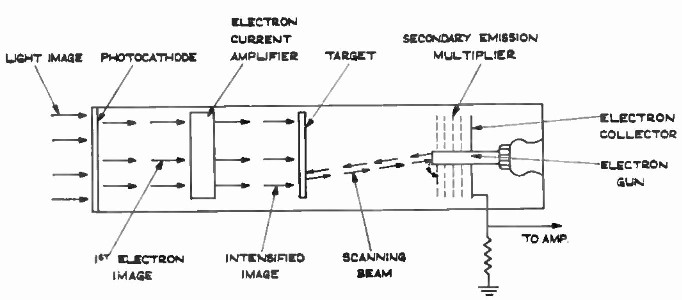


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Fig. 4. (a) The image orthicon tube as used for a standard type of television broadcast camera. This is the most sensitive image tube available.

Fig. 4. (b) Morton's image intensifier tube.

instantaneous reproduction of faint images at the normal 1/25 sec exposure time of the standard broadcast television system, they also have the ability to integrate faint light in the manner of the photographic emulsion. Experiments in America have shown that the image orthicon target is capable of storage times up to 40 minutes with only slight loss of resolution. Target leakage which affects integration time is due to both surface and volume conductivity effects and the latter is an exponential function of temperature. Therefore, in order to obtain long integration times, and also to maintain image resolution at shorter exposures, target refrigeration becomes necessary. However, McGee has proposed a charge integration tube based on a modified C.P.S. Emitron camera tube which has the addition of a high quantum efficiency photocathode and orthogonal scanning for target evaluation as in the orthicon tube. A double-sided target is used for charge integration and consists of a metal plate coated on both surfaces with a very thin, highly insulating layer of silicon or magnesium fluoride. The target output is fed directly into an amplifier via the metal signal plate which is mounted on a spindle so that it can be rotated by an external magnet to bring either side to face the photocathode. The photoelectrons bombard the target with energies of 500 to 1000 eV sufficient to liberate 5 to 10 secondaries for each incident primary. The high target insulation combined with large capacitance enables charges to be integrated for periods up to an hour or more so as to intensify threshold star images to a level where they can be distinguished from the night-sky background illumination. When integration is complete the target is reversed so that the charged surface faces the scanning beam which may be deflected at a slow rate with the advantage of being able to reduce bandwidth with consequent improvement in signal-to-noise ratio in the signal amplifier.

In the normal unaided photographic process of integrating light from a faint star against a relatively bright sky background, there is no way of intensifying the desired image to increase the extremely low contrast or means of subtracting the spurious "fogging" effect surrounding it. From the photographic viewpoint this would appear to be an insuperable problem but it is one which lends itself most aptly for solution by television methods. An intensifier type of orthicon tube with sufficient amplification of the electron current associated with the primary photoelectron emission will result in each electron producing an effect on the display screen of the television system. This single effect can then be integrated by the normal photographic recording process and obviates the problem of storage in the target of the image tube itself. The subtraction of the unwanted background signal resulting from the "night-sky glow" can be overcome by lowering the "black-level" of the cathode ray tube to a point where only the required signal pulses can be observed. These results would be similar to those obtained by the Lallemand photographic image converter. Neither of these photoelectric devices suffers from the "reciprocity failure" associated with long exposure photography.

3.3.2. Recording of planetary images

It is a requirement in planetary photography to obtain a recorded image on a film or plate emulsion with the *shortest* possible exposure time in order to overcome the effects of atmospheric turbulence or "seeing." Bearing in mind that the near planets including the Moon possess a fairly high albedo (ability to reflect sunlight) a bright image is obtainable at the prime focus of a large telescope and only a medium-speed-fine-grain photographic emulsion is required so as to record it in an exposure time of a few seconds. However, the continuous motion of the atmosphere produces a refraction differential which results in a gyration of the planet's image on the photographic plate at the focus of the telescope. This produces a superimposition of images on the final picture and spoils the definition to the detriment of the fine structure in the image. Even a few seconds' exposure is likely to show the "seeing" effects on the quality of the recorded image. The fastest photographic emulsions are unable to reduce the required exposure to the order of a small fraction of a second and even if they were adequate as regards sensitivity, the grain structure would be coarse. Here is another problem to which image converters and television techniques

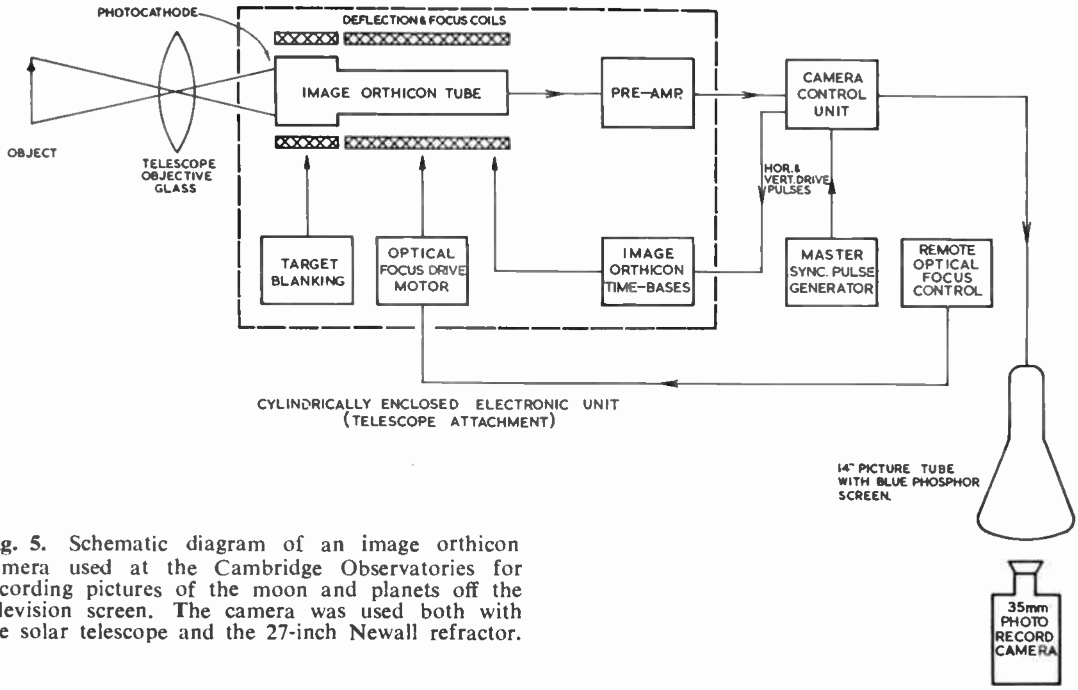


Fig. 5. Schematic diagram of an image orthicon camera used at the Cambridge Observatories for recording pictures of the moon and planets off the television screen. The camera was used both with the solar telescope and the 27-inch Newall refractor.

have been applied with a considerable measure of success, since their higher quantum efficiencies permit exposures as short as 0·01 seconds. For this application a 'charge integration' type of tube is unnecessary since the light flux available from a planet is of much greater intensity than that of starlight. The requirements can therefore be met by a high sensitivity tube such as the image orthicon, which possesses a secondary electron multiplier, operating at a frame time in the order of fractions of a second. By using this tube in a television camera operating at a high sequential frame repetition frequency, say 100 per second, an effective exposure time of 1/100 second is feasible; a photographic camera with a wide aperture lens and 1/100 second shutter speed is then used for recording pictures off the television display screen.

4. Television Methods in Practice

4.1. *Image Orthicon Camera*

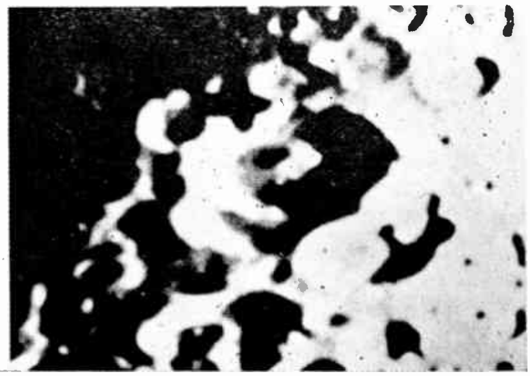
As remarkable and imaginative as the designs for image intensifier tubes may appear, such tubes for use in special television circuit techniques to meet specific kinds of astronomical application have not been available for tests with the telescope. The decision of Fellgett and the author to commence tests in 1951 with the standard type of image orthicon camera was made with the object of evaluating the performance of the image orthicon (Fig. 4(a), the most sensitive image tube available at that time, and to obtain a direct comparison with an average astronomical plate. Moreover, it was hoped that strong evidence in support of development on special television tubes and circuit techniques would be forthcoming. The manner in which the television equipment was first used at the Cambridge University Observatories is shown schematically in Fig. 5. A standard type image orthicon camera operating on a 625-lines 25 frames per second system standard was used on the Solar Tunnel telescope at the prime focus where an image of the Moon was focused on the photocathode of the image orthicon tube in the camera. The aperture of the Solar Tunnel is 12 in. and focal length 60 ft. Using this kind of astronomical telescope designed mainly for solar observations, it was possible to place the television camera in a stationary position in line with the optical axis and in the focal plane of the image formed by the collimating objective.

When the camera is used with a refractor it must be attached directly to the structure at the prime focus and aligned with the optical axis of the main objective. Photographs were taken off the television screen at a twentieth of the exposure time required to record the same image at more or less the same instant using an ordinary astronomical plate (Kodak 103a-o). The direct comparison of the two methods (Fig. 6) clearly shows the enhanced contrast of the image by the television technique. Greater detail can be seen in the television picture since the definition has been improved by shortening the exposure time and defeating some of the "seeing" effects. Similar tests were made on the Newall 27-inch refractor with Saturn and Jupiter as the subjects.

in the order of 0·02 foot-lamberts, gave the impression that improved signal-to-noise ratio might be possible by increasing the target-to-mesh spacing. Various spaced targets were tested and it was found that optimum results were obtained with target spacings of 8 mils, beyond which no further improvement was apparent. The performance obtained with the increased spacing was very much below that claimed by A. A. Rotow.[10] These specially modified tubes were tested in a television camera operating on 625-lines but with a sequential frame speed of 75 per second. This meant that the tube operated with an effective target exposure time of 1/75 second, and pictures could be recorded in the same time off the display screen using a camera shutter speed to correspond.

(*a*)

Fig. 6. (a) The moon recorded by direct photography at an exposure time of 4 sec. (Cambridge solar tunnel telescope).



(*b*)

Fig. 6. (b) The moon recorded by television with an exposure time of 0·2 sec off the screen.

Following the series of tests at the telescope using the 625-lines standard image orthicon camera, to obtain some assessment of the image tube performance, a series of simulated experiments was carried out in the laboratories of the author's organization. These experiments were made with the object of carrying out a more precise assessment of the image orthicon tube compared with the fastest photographic film available (Ilford HPS). However, by this time theoretical speculation regarding the exact behaviour of the image tube target, when operating at extremely low light levels

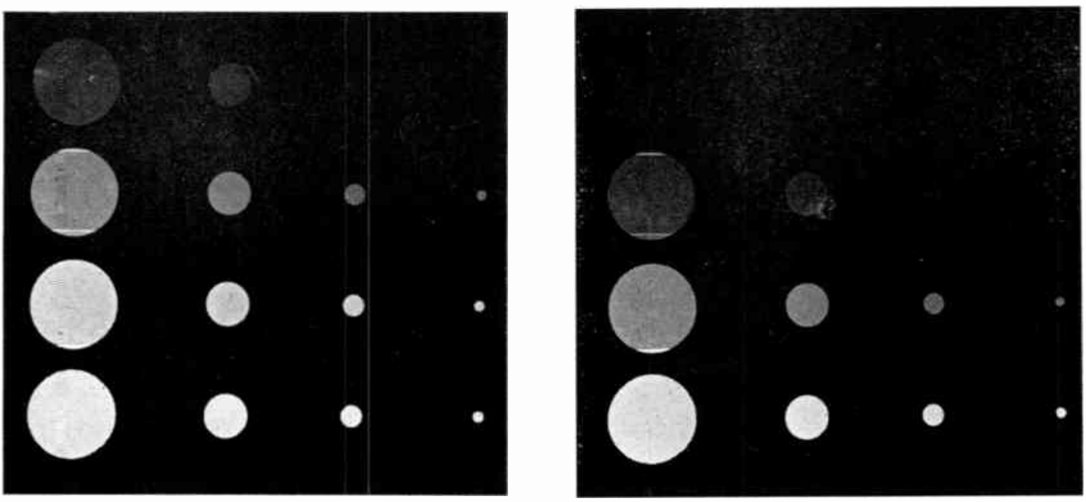
In seeking conclusive evidence of the gain in sensitivity of the image orthicon tube over HPS film it was also essential to compare image resolution in picture points at the lowest light levels.

4.1.1. Performance test chart

In one of the methods a chart was constructed on the lines proposed by A. Rose who devised a technique for giving direct evaluation of the performance of an image detector such as the eye, photography or television camera. A similar chart used by the author

Fig. 6. (c) The image orthicon camera assembly attached to the refracting telescope at the Dublin Observatory for the demonstrations to the I.A.U. in 1955.

consisted of a metal plate about 12 in. x 8 in. with rows of holes progressively reduced in area by a factor of 4 to 1 along each row. At the back of each row, proceeding downwards, strips of film each of which had been given an exposure time to produce a density ratio of 2 : 1 between any two successive strips, were placed over each row of holes to give a progressive variation in contrast from the top to the bottom rows of the chart (Fig. 7). The chart was illuminated evenly from the rear by a cold light source. The television and photographic cameras were set up at a given distance and the same optical conditions applied in each case to focus upon the chart. The appearance of the illuminated chart as viewed by the television or photographic camera showed the limit of visibility of each hole along a series of diagonal lines from one corner to the other. The number of holes registered by the television camera was noted and the chart photographed under the same conditions. After processing the film, using photographic materials to give optimum results, a direct comparison between the two image detectors could be made in terms of



(a) *(b)*

Fig. 7. Pictures obtained by direct photography of a chart similar to the one devised by A. Rose for obtaining a direct assessment of the performance of image detectors.

(a) The picture on the left illustrates the performance of the photographic film in terms of sensitivity and resolution. The effects of a low contrast on image recognition are clearly seen.

(b) The picture on the right was taken at a reduced rear illumination of the chart, and shows a shift of the diagonal towards the images of increased area and contrast.

sensitivity and resolution. The finite recognition was obtained for each detector at the point where the smallest circular image along the diagonal could be distinguished from the " noise " in the television image or grain density of the film. As the intensity of the illumination decreases, the " diagonal " shifts towards the holes producing higher contrasts. This particular test confirmed the factor of 3 gain in sensitivity of the modified image orthicon tube over the fastest film emulsion. Observation of the chart by the eye shows that it has more difficulty in resolving images of very low contrast, whereas the performance of the image orthicon is maintained to a higher degree.

4.1.2. Image orthicon target exposure tests

These tests were conducted in order to ascertain the image tube target charge and discharge characteristics at very low light levels, and thereby check the effective exposure time during the scanning cycle. For this experiment, the camera, working on 625-lines 75 frames per second, was set up and focused upon a flashing light strobe adjusted to a pulse repetition rate of 10 per second. Arrangements were also made to vary the flash intensity. The video output of the camera was fed into an

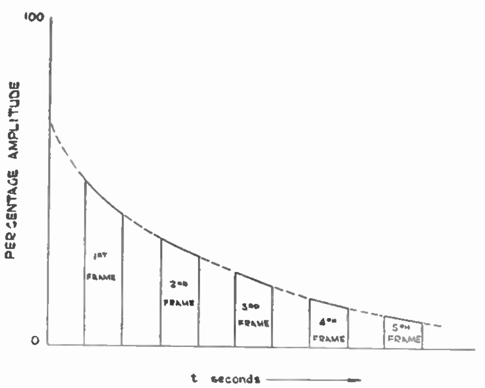


Fig. 8. Showing the charge decay of an image orthicon target at low light levels. The residual charge from one frame to another is due to failure of the scanning beam to completely discharge the target in a single television frame time.

oscilloscope with line selection facilities and a trace obtained of the target charge decay or persistence over ten picture frames. The results were somewhat surprising and disclosed the fact that at the lowest levels of illumination the scanning beam failed to completely discharge the target. In some cases, the residual charge over the 2nd and 3rd frames was as much as 50 per cent. The oscilloscope amplitudes and decay times were plotted as a curve showing the target charge persistence over a number of frames, (Fig. 8). From these tests it was concluded that if a photographic camera shutter set for 1/75 second were to be placed in front of the image orthicon photocathode and synchronized to operate with the television system frame rate, the result would be less favourable to the image tube in a performance test. Nevertheless, it was essential to take this result into account in comparing the performance of the image orthicon with film for which the photographic camera shutter gives a realistic exposure time.

5. Televised Images of Mars

5.1. *Equipment Set-up*

The results achieved in tests with the equipment at the University Observatories and in the Laboratory were sent to the Director of the Lowell Observatory (Dr. A. Wilson) in America at his request. In Dr. Wilson's opinion, the results were far more encouraging than those hitherto obtained with a similar American television system known as the " Lumicon ". This factor led to an invitation being extended to the author to join Mr. E. C. Slipher, a leading authority on the planet Mars, in an expedition to the Lamont-Hussey Observatory, Bloemfontein, South Africa, during the summer of 1956 in time for the " favourable opposition " on September 7th of that year. The expedition was sponsored by the National Geographical Society of Washington D.C. The equipment used was a 625-lines 75 frames (sequential) system, using Cathodeon wide-spaced image orthicon tubes in the camera. The camera itself was modified to provide for temperature control of the image orthicon target in order to maintain resolution of the planetary image over long periods of operation. This was necessary on account of the considerable amount of heat generated by the image tube focus and deflection coils, and surrounding valves. For cooling the image target, compressed air was blown evenly across the tube face. A temperature measuring circuit was built-into the image orthicon tube assembly for indicating the thermal conditions during operation of the camera on the telescope. This precaution for cooling the tube target proved well worth while as Dr. Slipher informed the author that, when making similar observations on the planet in 1954 using the Lumicon equipment, the image definition showed considered deterioration after it had been in operation for periods of over half an hour. Excessive temperature rise within the television camera might well have been the cause. The camera, weighing over 100 lb, was attached to the 28-inch refractor, of 42 ft focal length, by means of a rigid welded metal cradle which could be bolted on to the spectograph mounting at the prime focus. At "opposition" the disc of Mars had an apparent diameter of 25 seconds of arc and the image formed by the main telescope objective at the focus was about 3 mm. This image size was considered too small to focus upon the image tube photocathode in order to obtain maximum resolution from the television system. A small relay lens was therefore interposed and the image size of the photocathode was increased to 8 mm approximately. This produced an image 48 mm on the 7-in. flat screen on the television display monitor from which pictures were recorded by a 35 mm camera fitted with an f 1·9 lens and rapid action film transport mechanism. The camera had a film capacity of 200 exposures 24 mm x 24 mm in size. HPS film was used throughout the observations and over 1500 television pictures of Mars were recorded over a period of two months. Unfortunately, for most of these observations the disc of the planet was obscured by dense yellow clouds presumably due to a dust storm in the Martian atmosphere.

5.2. *Direct Observations*

The development of the disturbance of the Martian atmosphere has been very well illustrated in sketches of the planet made during Dr. Kuiper's observations at the 120-inch telescope of the McDonald Observatory, Texas. The dust storm first appeared over the central part of the Mare Sirenum on the 30th August, when five bright clouds appeared within the dust area above the Mare Cimmerium (Figs. 9(a) and (b)). By the 31st August the dust storm stretched into a giant "W" with the appearance of a jet-stream centred over the Mare Sirenum (Fig. 9(c)). During September 2nd and 3rd, the dust storm had completely enveloped most of the surface details which

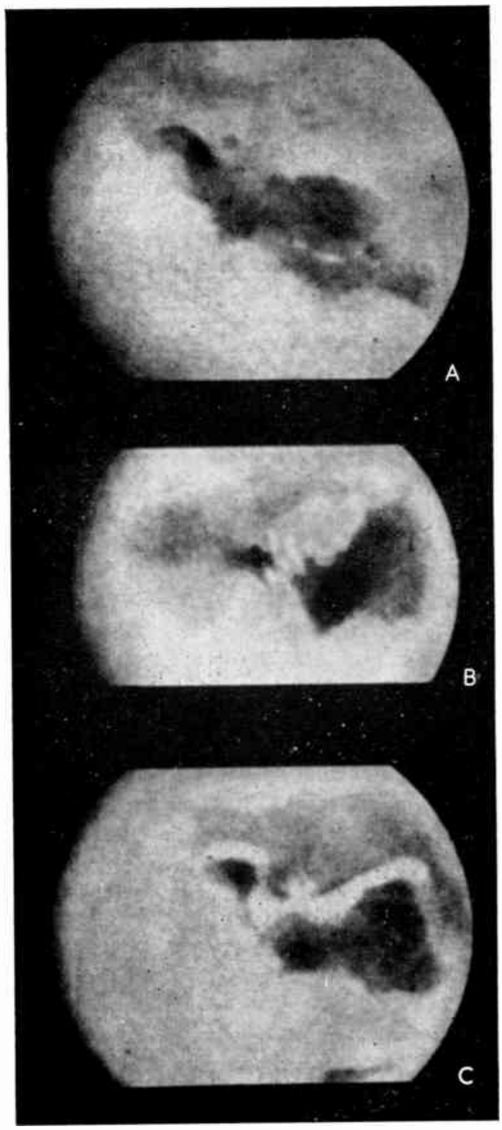


Fig. 9. Sketches of the planet Mars made by Dr. Kuiper in America during the 1956 "Opposition."
(a), (b) and (c) show development of the disturbance in the Martian atmosphere over the regions of the Mare Sirenum and Mare Cimmerium.

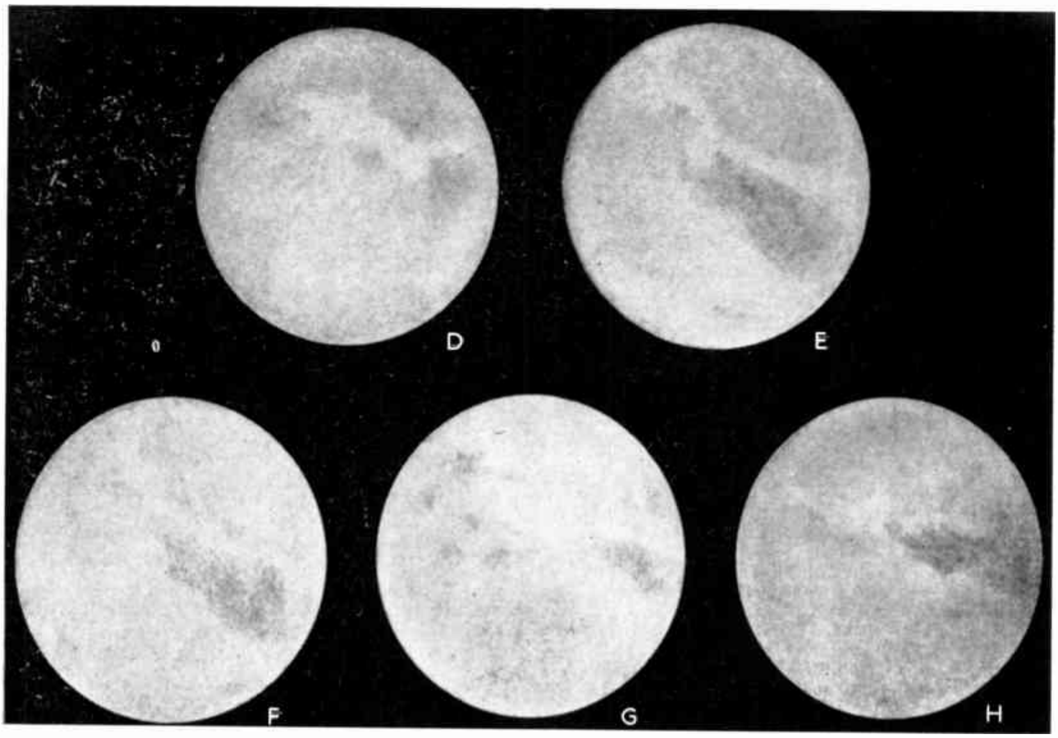


Fig. 9. (*contd.*) (d) and (e) show the features of Mars almost completely enveloped by the "dust storm." (f) and (h) show the re-deposited Polar Cap, probably due to snowfall, with a dark surrounding fringe.

appeared in extremely low contrast when viewed directly through the telescope (Figs. 9(d) and (e)). On the 4th September a new cloud appeared over the South Pole and several dust storms invaded the dark equatorial regions; by the 14th September, the pole cap again became visible and a new surface deposit, presumably due to a snowfall, surrounded the inner portion of the cap. From the sketch in Figs. 9(f) and (h) the dark zone surrounding the outer portion of the cap, appeared to be slightly brownish to Dr. Kuiper. The clearing of the atmosphere between the 9th and 23rd of September again revealed the fine surface structure of the planet with increasing contrast and television observations and recordings were continued (Fig. 10). On account of the low contrast of the Martian features throughout the period of observation, the use of television techniques played their most important role under the prevailing viewing conditions by making full use of the ability of the television system to enhance contrast electronically.

5.3. *Television versus Direct Photography*

Dr. Slipher has stated that in analysing and comparing the television pictures with those he obtained by direct photography, a total of 40,000 photographs over a period of six months, the latter were on the average slightly better. However, in making comparisons of the results obtained by the two methods, it should be borne in mind that the pictures were not recorded by direct photography and television at the same instant and therefore the "seeing" conditions were not identical in each case. However, statistics favoured photography since a far greater number of pictures were obtained direct over a very much longer period than were taken off the television screen, the proportion being about 25 : 1.

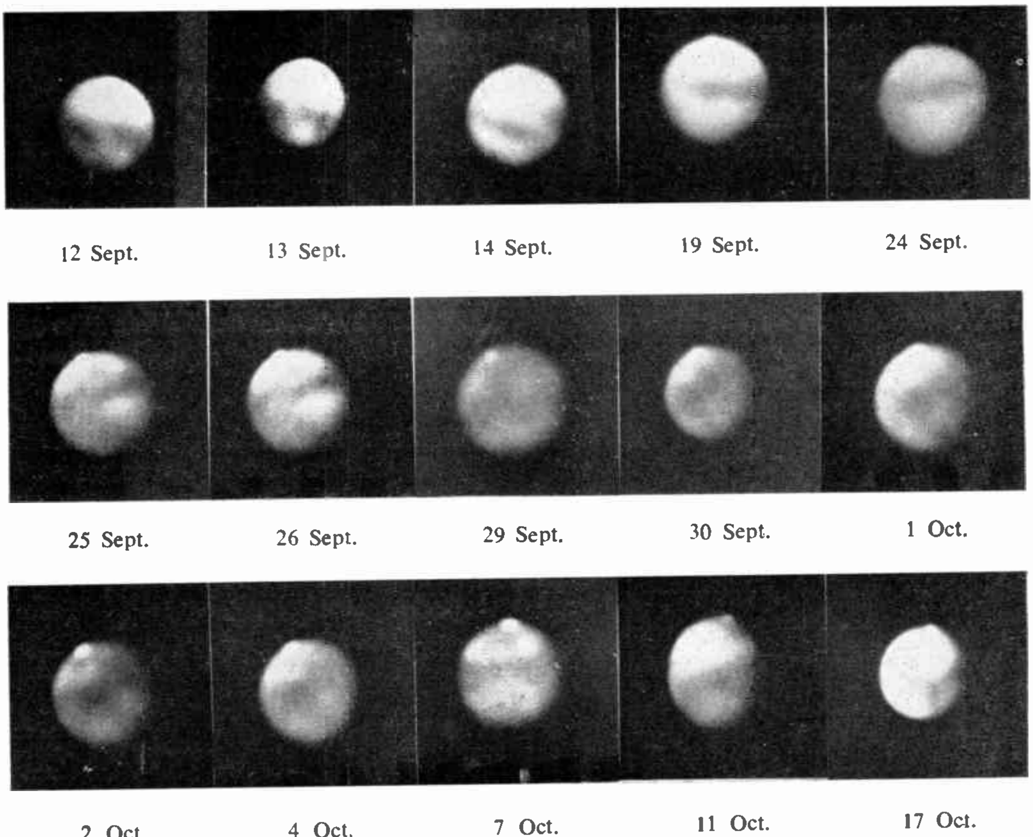


Fig. 10. Pictures of Mars photographed by the author off the television screen during the period following the "close opposition" on the 7th September, 1956. It will be seen that no features of the disc were visible until the 14th September when the Polar Cap and markings in the equatorial region reappeared. By October the contrast of the images had greatly improved.

Whatever assessment may be given to the value of the television observations made during the 1956 opposition, there is no doubt that it was the most serious attempt made to use television techniques to aid observational astronomy.

6. Space-orbiting Television Telescope

6.1. *Ground-controlled Space Telescope*

At the 9th Assembly of the International Astronautical Federation which took place in Amsterdam during 1958, F. Whipple, an astronomer at the Smithsonian Observatory in America, submitted a proposal for the construction of a space-orbiting astronomical telescope fitted with a television scanning device capable of being completely controlled from a ground observatory. If all other attempts to overcome the great disadvantage of viewing celestial objects through the Earth's atmosphere fail, this would provide a solution, but at enormous cost. The term " telescope " in this case refers to a device that will focus radiation from celestial objects so that an electronic image scanning system can record an area of the sky for transmission to a ground monitoring station. For this purpose, the space telescope would be operated by remote control and instruments provided to stabilize the angular rotation of the satellite to a minimum prescribed value. The optimum resolving power of the telescope would be dependent upon this tolerance and the maximum effective exposure time possible with the type of television equipment employed. Powered controls operated by command signals from the ground station would also have to be provided for focusing and orientation. The main advantage of placing such an astronomical telescope in a satellite orbit is that it brings into observation the far ultra-violet spectrum below 3000 Å.

6.2. *Observatory in Space*

Another proposal has been made by F. A. Smith of the British Interplanetary Society, London, for an astronomical telescope in space but controlled from a space station; a diagram of the instrument is shown in Fig. 11. The observatory consists of a Cassegrain telescope mounted in a framework with a segmented primary mirror, capable of being transported into space by a freight rocket of the future, to the controlling space station from which the telescope would be assembled in orbit. A chamber is provided for the observer at the prime focus and in considering the behaviour of the man and the telescope optics, the weightless conditions must be taken into account.

7. Conclusion

Underlying all the experimental work on the development of special photo-electric image tubes is the aim to produce a television system employing simple techniques which can eventually be used as a routine instrument in observatories throughout the World. The International Astronomical Union have pointed out that this state of affairs cannot be achieved by astronomers alone working on their own initiative, but the co-operation of electronics specialists, industry, and the necessary financial support, will be required. Unfortunately, industry is often reluctant to invest large sums of money in projects which, *prima facie*, may appear to be of purely scientific value. Yet, more often than not, the knowledge thereby acquired, has paved the way for major developments affecting the daily

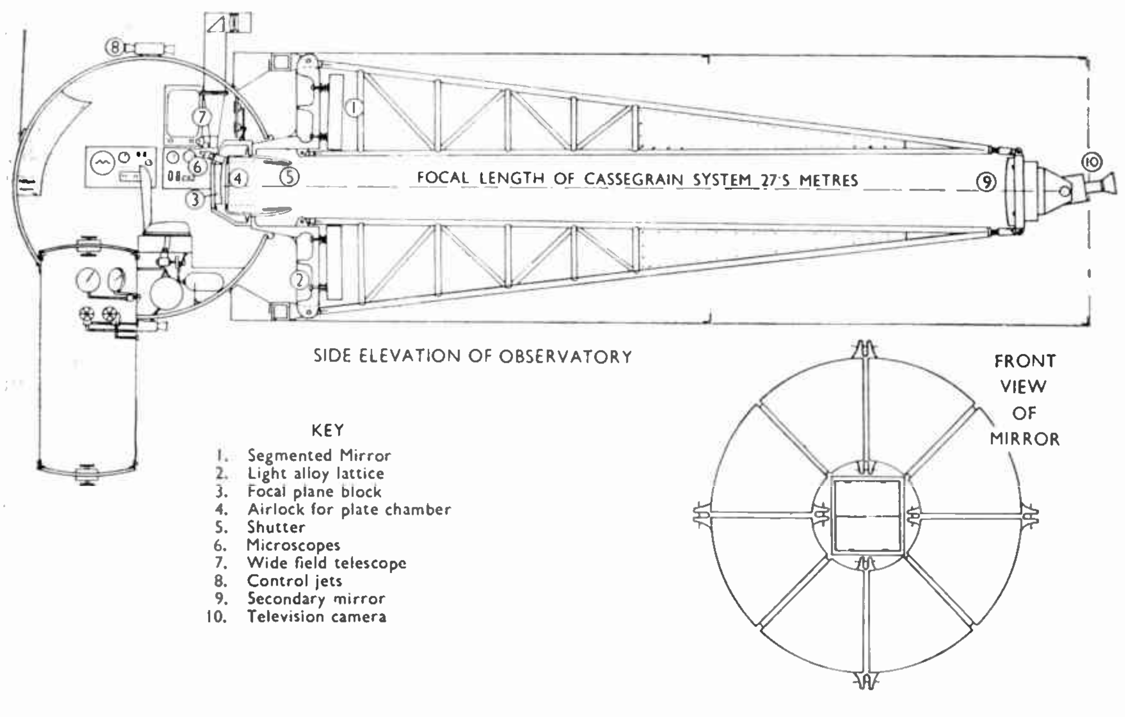


Fig. 11. Schematic diagram of a satellite telescope proposed by F. A. Smith of the British Interplanetary Society,[12] which could be controlled from a space station. The mirror is shown in segmented form for transport by freight rockets. A television camera is attached behind an aperture in the secondary mirror for transmission of stellar images to the main space station. Orientation of the telescope in space is controlled by the jet motors.

lives of men and nations all over the World. The discovery of the immense power contained by the atomic nucleus may be attributed directly to the work of astronomers and their solar observations using spectroscopy, rather than to the physicists.

Since the basic problem in developing image tubes for astronomy is one of faint light intensification, a successful solution and the production of a suitable device might well be applied to other applications where low levels of illumination are involved; for example, in radiology the low-contrast X-ray images shown on the fluoroscope screen are often too faint to be interpreted by the radiologist unless viewed after the eyes have become "dark-adapted." Experiments using the image orthicon in connection with X-ray equipment have already been conducted in America and Germany and have been partly successful.

It may well be that really significant gains in sensitivity will not be achieved by the present photo-electric techniques which may be superseded in the not too distant future by "solid state" image intensifiers with quantum efficiencies approaching unity. Initially, the performance of such devices may be limited by a poor signal-to-noise factor.

The use of television methods to defeat the adverse effects of "seeing" on recording fine details of, for example, the surface structure of Mars may well be superseded in the not too distant future by the placing of telescopes outside most of the Earth's atmosphere by means of balloons or even satellites as described. Nevertheless the applications of photo-electric image devices generally will probably become of increasing importance, especially in the field of faint star recognition by the separation electronically of night-sky background illumination. Increased performances of photographic emulsions and television image tubes, through greater quantum yields, will be of inestimable value to the astronomer and may well lead to secondary applications of considerable significance in other fields.

8. Acknowledgments

The author wishes to express his appreciation of the invaluable help and advice of his colleagues at Pye Ltd., and indebtedness to Dr. P. B. Fellgett for his close collaboration in all aspects of the work. Also, to Mr. R. Law for his assistance with the photography, and to Mr. B. J. Edwards, Technical Director, for his support and permission to publish results of the investigation referred to in this paper.

9. References and Bibliography

1. P. B. Fellgett, Proceeding of Symposium on Astronomical Optics, Manchester, April 1955.
2. A. Beer (Editor), "Vistas in Astronomy," Vol. 1. (Pergamon Press, London, 1954/55.)
3. M. Lallemand, Proceedings of Symposium on Photoelectric Image Tubes, General Assembly of the I.A.U., Moscow, 1958.
4. B. V. Somes-Charlton, "Television astronomy," *New Scientist*, 1st May 1958.
5. "Mars Symposium Report," *Sky and Telescope*, April 1957.
6. Gerard de Vaucouleurs, "The Planet Mars," (Faber & Faber, London).
7. Gerard de Vaucouleurs, "La Photographie Astronomique," (Albin Michel, Paris).
8. J. D. McGee, "Television in the service of science," *Journal of the Television Society*, **8**, No. 2, April/June, 1956.
9. B. V. Somes-Charlton, "Applications des techniques de television à l'astronomie d'observation," *Ciel et Terre*, May/June, 1956. (Journal of the Belgian Astronomical Society, Brussels).
10. A. A. Rotow, "Image orthicon for pick-up at low light levels," *R.C.A. Review*, **17**, No. 3, September 1956.
11. J. H. DeWitt, R. H. Hardie and Carl K. Seyfert, "A seeing compensator employing television techniques," *Sky and Telescope*, November, 1957.
12. F. A. Smith, "Satellite telescope," *Journal of the British Interplanetary Society*, **16**, No. 6, March/April, 1958.

DISCUSSION*

on

"Audio Frequency Selective Amplifiers"†

R. Hutchins‡: In connexion with some recent work on selective amplifiers, I referred to the above paper and feel I should point out an error which has been made.

The error occurs when dealing with the condition for optimum selectivity using the Wien bridge arm network (section 8 of the paper). For the Wien bridge arm shown in Fig. A, the voltage transfer ratio of the network should read:

$$\frac{V_{out}}{V_{in}} = \frac{1}{(1+n+m)+\mathrm{j}\left(n\omega CR - \dfrac{m}{\omega CR}\right)}$$

The real part of the denominator is given in the paper as $(1+n+m/n)$ and this error is carried through in subsequent calculations.

The expression for selectivity should become

$$Q = \frac{G_0\sqrt{m\,n}}{(1+m+n)^2}$$

For a given value of G_0, this expression has an optimum value for $n = m = \frac{1}{2}$ for which the selectivity becomes $Q = G_0/8$. It is coincidental that this agrees with the maximum value as obtained by Punnett. It should be pointed out that a small departure from these ratios is not critical in obtaining practical values for optimum Q.

The above error is continued in section 9 of the paper, dealing with the Wien bridge connected to the output of a phase splitting valve.

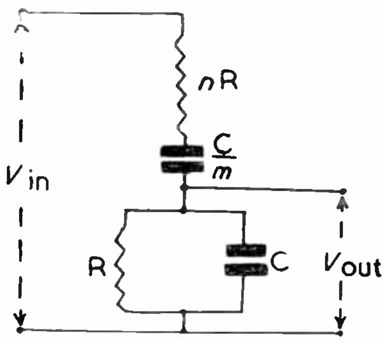


Fig. A.

The expression for the selectivity in this case should read:

$$Q = (G_0+1)\,\frac{\sqrt{m\,n}}{(1+m+n)}\quad .$$

For a given value of G_0, this has no maximum at any practical value of n or m; it increases to $Q = G_0+1/2$ as $n = m \to \infty$, but is reduced to $Q = G_0+1/3$ for $n = m = 1$. Here again the actual ratios are not critical in determining the selectivity.

Mr. S. W. Punnett has confirmed the above correction.

* Received by the Institution on 15th May 1959. (Contribution No. 20.)

† S. W. Punnett, *J. Brit.I.R.E.*, **10**, No. 2, pp. 39-54. February 1950.

‡ Northampton College of Advanced Technology, London, E.C.1.

U.D.C. No. 621.375.126

Parliamentary and Educational News

Parliamentary and Scientific Committee Conversazione

The Parliamentary and Scientific Committee is organizing a Conversazione which will be held on Wednesday, 9th December next, at Burlington House in the rooms of the Royal Society. The theme of the Conversazione will be "Science in everyday life," and a number of organizations who are members of the Committee are arranging appropriate exhibits. Developments in medicine, transport, building, communications, leisure, clothing and food, will be featured.

The Institution, which was a founder member of the Parliamentary and Scientific Committee, has been asked to sponsor suitable exhibits. An advance report of the exhibits will be published in a subsequent issue of the *Journal*. The Institution's recommendations include exhibits demonstrating recent advances in television and broadcasting and stereophonic reproduction.

Other Institutions in membership of the Parliamentary and Scientific Committee have also been invited to sponsor exhibits relative to their professional spheres.

Concentration of Advanced Courses in Science and Technology

One of the most difficult problems facing administrators of further education is to decide to what extent advanced courses should be concentrated at a limited number of centres.

This was stated recently at a meeting in Harrogate by Mr. A. A. Part, Head of the Further Education Branch of the Ministry of Education. Most of the courses at present were part-time and therefore had to be provided in many centres so that they were within daily travelling distance for the majority of students.

This had resulted in too many centres with very small numbers of students, and one of the reasons for prescribing minimum enrolment figures for existing courses was to close down courses that had persistently failed to establish themselves on a satisfactory basis. For advanced full-time and sandwich courses concentration was even more important, for they made particularly heavy demands on scarce staff and expensive equipment. Mr. Part said that we could not hope to create colleges comparable with the technical universities on the Continent unless we were prepared to concentrate to a considerable extent.

There were a few colleges other than colleges of advanced technology which could meet the stringent conditions rightly laid down by the Hives Council and it was sometimes right to approve a H.N.D. course at an area college. But, in general, we should only succeed in getting the quality that the country needed if we were to build up a limited number of really strong centres. For this reason he was glad to see that out of 2,500 students working for the Dip.Tech., 1,900 were in the eight colleges of advanced technology.

Development of Inventions

The President of the Board of Trade was asked recently in the House of Commons whether he could state the progress which had been made in the development of the Gabor cathode-ray tube for colour television, and whether he estimated that it would be ready for application at the time that colour television was generally introduced in this country. Sir David Eccles replied that this invention had been developed by the National Research Development Corporation, and its application was a matter for the radio industry.

Future of Television Broadcasting

The Postmaster General has declined to make available to Parliament the report of the Technical Sub-committee of the Television Advisory Committee until the main committee's report is published. He had been urged to circulate the sub-committee's report dealing with questions of television line standards, allocation of wavelengths and colour television in advance of the White Paper containing the T.A.C's second report. This procedure would have enabled the House to examine the basis of any proposed change likely to be put forward in the White Paper.

News from the Sections . . .

North Western Section

On June 23rd a party of 22 members of the Institution, visited the Semiconductor works of the General Electric Company Limited at Hazel Grove, Stockport, Cheshire.

After brief introductory talks by Mr. B. R. A. Bettridge (Member) and Mr. R. J. Horlock, the party was divided into small groups which were shown over the factory by operatives skilled in the many branches of semiconductor manufacture. During the tour, which began at 10.30 a.m. and finished at 4.15 p.m., the parties were shown all the stages of semiconductor manufacture (except glass-metal sealing) from preparation of raw materials to final testing of the finished product.

There were two operations which caused considerable interest, namely, the "pressing" of indium pellets to the germanium "slices" or discs, and the "cold-sealing" process after the cans were put on. The speed and dexterity with which the operatives carried out the first operation has to be seen to be believed; the "cold-sealing" process, which has been developed by G.E.C., is considered to be a great advance in semi-conductor manufacture.

Throughout manufacture the emphasis was placed on rigid control of quality and the visit to the quality control department readily demonstrated this.

The visit was extremely well organized from start to finish, and the Section is very grateful to the General Electric Company for their hospitality. P. A. B.

South Western Section

A party of members of the South Western Section and guests spent a very pleasant evening on June 18th delving into the technical and operational aspects of the Portishead shore-to-ship transmitting station. The party was conducted around the station by Mr. Hipkins, the deputy chief engineer, and members were able to see a range of working and standby equipment from an older Post Office 17 kc/s transmitter to modern Marconi and S.T.&C. c.w. short wave transmitters. A strong impression of reliable operational work was created in the minds of the visitors, who were also greatly interested by the 32 different arrays spread over a considerable acreage.

On the same evening another party from the Section visited the Headquarters of the B.B.C. West Region. Mr. Daly, the engineer-in-chief, provided technical guides at every stage of the tour, and members were most impressed by the high level of organization displayed. The recording section was the first call, where the party observed the recording of a standard news bulletin for retransmission on the Wenvoe f.m. channel. A typical television studio with three vidicon cameras was examined, the monitors providing some indication of "closed-circuit" television. Members had an excellent opportunity of comparing the vidicon with the image orthicons used in the rehearsal of "View from the West," which they attended. The control room for this rehearsal provided a sharp contrast with the calm of the technical arrangements—most members admitted that the post of television programme controller would not be their choice for a career!

The party was then conducted over the switching rooms for both sound and vision broadcasting the control of remote stations was explained and provided an absorbing study in line equalization and automation. The monitoring of the Channel 5 transmissions with high quality apparatus *and* a typical commercial receiver showed obvious discrepancies!

D.R.McC. and W.C.H.

Autumn Visit

The Technical Committee announces that, by courtesy of the directors of Mullard Ltd. a visit is being arranged to the Company's semi-conductor factory at Southampton. The visit, which will last all day, will be on Tuesday, November 3rd. The number of participants has to be strictly limited and members are advised that applications will be dealt with in order of receipt.

Section Programmes for 1959-60

The calendar of meetings arranged for the first half of the session, (September to January), will be circulated to all members in the United Kingdom during September.

The Use of Pre-pulse Techniques in High-Speed Oscillography†

by

F. E. WHITEWAY, B.SC., ASSOCIATE MEMBER‡

A paper read before the Institution in London on 29th January 1958
In the Chair: Mr. A. G. Wray, Associate Member

Summary: The recording of single transients of very short rise time and duration has necessitated the development of ultra-high-speed oscilloscopes. It is shown how the high-frequency range of conventional oscilloscopes and associated amplifiers can be extended by switching them into operating conditions of high current for only a very short part of the duty cycle. An amplifier is described which has a gain of 360, an output of 200V and bandwidth 140 Mc/s at −3 db; using only four amplifying valves. Pre-pulsed time-bases are discussed, covering the range 5 to 250 mμsec, and a high-speed pulse generator is described which may be used for testing pre-pulsed oscilloscopes.

1. Introduction

In order to record single high-speed transients in the milli-microsecond range, it is necessary to employ cathode-ray tubes possessing very high writing speeds. To achieve this, high e.h.t. potentials must be applied, and if the cathode-ray tube contains a conventional deflector plate system the resultant deflection sensitivities are poor. This means that the deflection amplifiers must have a high linear output voltage as well as a wide bandwidth. As the input signal amplitudes are often quite low, for instance the signal available from a photomultiplier, the amplifier must also provide high amplification. These requirements tend to be conflicting, but they may be met by the use of pre-pulse techniques in which the valves are switched into their operating conditions for only a short part of the duty cycle. A similar technique may be used for increasing the speed of time-bases and pulse generators. The oscilloscopes can normally only be used if a pre-pulse is available in the region of 50 microseconds before the signal is received. Random pulses of identical character, however, can be recorded, but this is likely to be tedious unless the average pulse repetition rate is high.

† Manuscript first received on 11th December 1958 and in final form on 10th April 1959. (Paper No. 511.)

‡ U.K. Atomic Energy Authority, Atomic Weapons Research Establishment, Aldermaston, Berkshire.

U.D.C. No. 621.317.755

2. Cathode-ray Tubes

The heart of any high-speed display is the cathode-ray tube, and this sets a limit to the performance of the display. This paper will deal only with cathode-ray tubes possessing a conventional deflector plate system; tubes possessing travelling-wave deflection systems are described in a complementary paper[1]. Since the oscilloscopes are often required for the recording of single transients, the traces are usually recorded photographically before making measurements. These photographs can be enlarged as required, but the ultimate resolution will depend on the trace width. For this reason it is more appropriate to define cathode-ray tube sensitivity in terms of volts per trace width, rather than volts per centimetre. The trace width depends on spot size, and this varies considerably from one design of tube to another. In a similar way it is more desirable to define the maximum writing speed of a cathode-ray tube in terms of trace widths per second than centimetres per second. (The "maximum writing speed" is the maximum velocity which can be recorded photographically for a single trace and for a given optical system and film.) The cathode-ray tube itself sets a limit to the maximum bandwidth of the display system, due to two characteristics. These are:

(a) Plate ringing frequency. There is a finite inductance between the deflector plates and the external connections which is in series with the capacitance between the two deflector plates, and the capacitance between deflector plates and the third anode, interplate shield, etc. These components form a series tuned circuit, the resonant frequency of which is often referred to as the "plate ringing frequency." This should be made as high as possible by keeping the inductance and capacitance to a minimum. It may be noted that reduction in the interplate capacitance usually entails loss of deflection sensitivity.

(b) Plate transit time. It takes a finite time for electrons to pass through the plates, and the signal at higher frequencies will have changed in phase during this time. The consequent reduction in amplitude is given by

$$\frac{A(\omega)}{A_0} = \frac{\sin \frac{1}{2}\omega t_r}{\frac{1}{2}\omega t_r}$$

where A_0 = deflection at very low frequency
$A(\omega)$ = deflection at frequency $\omega/2\pi$
t_r = plate transit time.

In order to obtain sufficient bandwidth, the length of the plates must be restricted, and this also reduces the sensitivity. Similar effects are present at other electrodes, setting a limit to the rise time of the beam brightening waveform and affecting the start of the time-base. In order to obtain sufficient writing speed, it is necessary to use a high overall e.h.t., often in the range 20-30 kV: this again means reduced sensitivity.

The VCRX434 cathode-ray tube represents a good compromise between the conflicting factors of writing speed, sensitivity, transit time, etc., and is the cathode-ray tube which will be considered for the remainder of this paper. It possesses the characteristics shown in Table 1.

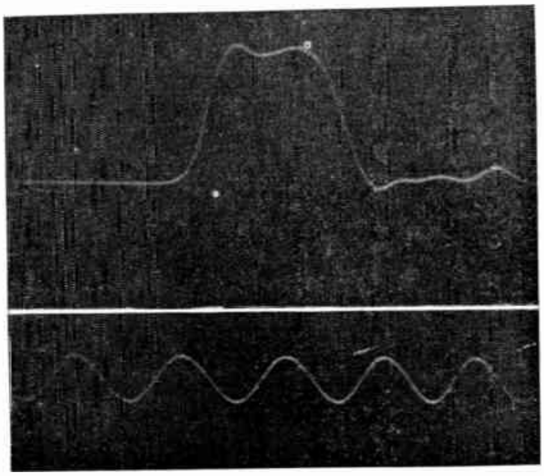


Fig. 1. Step response of VCRX434.

Figure 1 shows a single-shot photograph of a pulse from a relay pulse generator fed directly to the plates of a VCRX434 cathode-ray tube through a 100-ohms damping resistor. The rise time is governed mainly by the characteristics of the tube and is about 0·7 mμsec.

3. Amplifiers

For a Y deflection of 100 trace widths (about 3 cm on the screen), the output required from the amplifier to drive a VCRX434 cathode-ray tube is 180V. The required gain of the amplifier will vary considerably. For example, when a photomultiplier is used the signal level will

Table 1

E.h.t. on gun	7·5kV
Overall e.h.t.	30kV
Y sensitivity	1·8V/trace width
X sensitivity	2·0V/trace width
Trace width	$\frac{1}{3}$ mm approx.
Maximum writing speed	2×10^{11} trace widths/sec, using a Wray F1 lens, 4:1 reduction ratio, & R60 film.
Phosphor type	P.11.
Transit time through Y plates	0·6 mμsec.
Plate ringing frequency	700 Mc/s approx.

often be in the region of 1V, since low impedance cables must be used in order to preserve bandwidth. This means that a gain of 180 or so is required. The bandwidth also depends on requirements, although an upper limit is set by the transit time and plate ringing frequency of the cathode-ray tube. The usual method adopted when designing amplifiers with a bandwidth in excess of 30 Mc/s is to make use of distributed line techniques. However, a wideband amplifier with a bandwidth greater than 100 Mc/s and a gain of several hundred would consist usually of about five distributed stages, each stage consisting of about six valves. The early stages are relatively easy to design up to a frequency of about 200 Mc/s, but an output stage required to supply a push-pull deflection of 180V at bandwidths in excess of 100 Mc/s is a very different proposition, and the h.t. consumption is high because of the low impedance lines necessary to retain the bandwidth. The total number of valves required in the whole amplifier would quite likely be between 40 and 50 for bandwidths in excess of 100 Mc/s.

An alternative system employs pre-pulse techniques in which a pulse which precedes the signal is used to switch the valves into their operating conditions.[2] As is well known, the

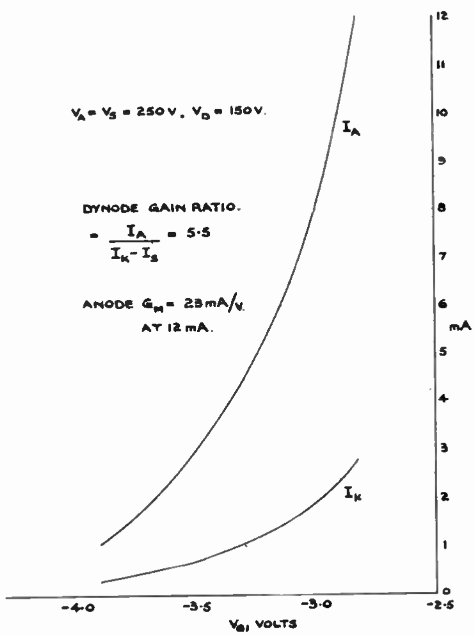


Fig. 2. Static characteristics of EFP60.

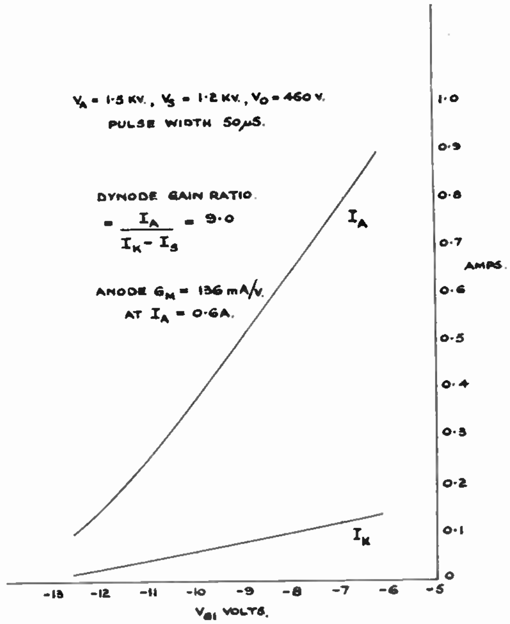


Fig. 3. Pulsed characteristics of EFP60.

mutual conductance of most valves increases with current, and an investigation was carried out with a number of likely valves to find out the extent of the increase under high current conditions. The most remarkable increase was found in the EFP60, a valve which has one stage of secondary emission, and which consequently has a high mutual conductance even under normal conditions. Figure 2 shows the normal curves of a particular EFP60 where the mutual conductance is seen to be as high as 23 mA/V at 11mA. The dynode gain ratio, i.e. the ratio of secondary electrons emitted by the dynode to primary electrons, is 5·5 : 1. Figure 3 shows the curves for the same EFP60 pulsed into conditions of high current and high voltage for a period of 50 microseconds and low repetition rate. The characteristics are much more linear. The mutual conductance is substantially constant over a wide range of current at about 130 mA/V. This remarkable increase has been brought about partly by the higher dynode gain ratio, resulting from the higher dynode potential, and the remainder by an increase in the mutual conductance of the primary system. By using EFP60s pre-pulsed into these conditions, cascade amplifiers may be designed with wide bandwidth and high linear output using few valves and low mean h.t. consumption. It is necessary to switch the amplifiers into operating conditions before the arrival of the signal and since the switching pulse has to be differentiated out. Thereafter they behave as class A amplifiers until the back edge of the switching pulse arrives. The low frequency response is neces-

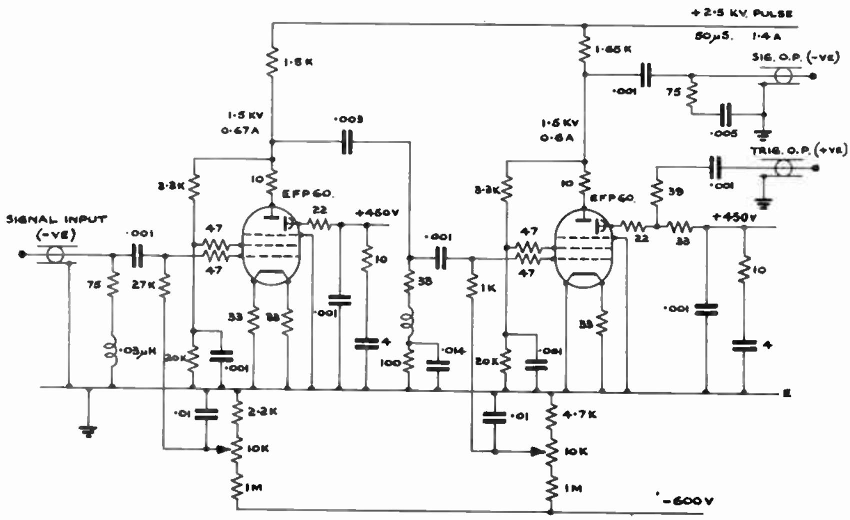


Fig. 4. Pre-pulsed pre-amplifier.

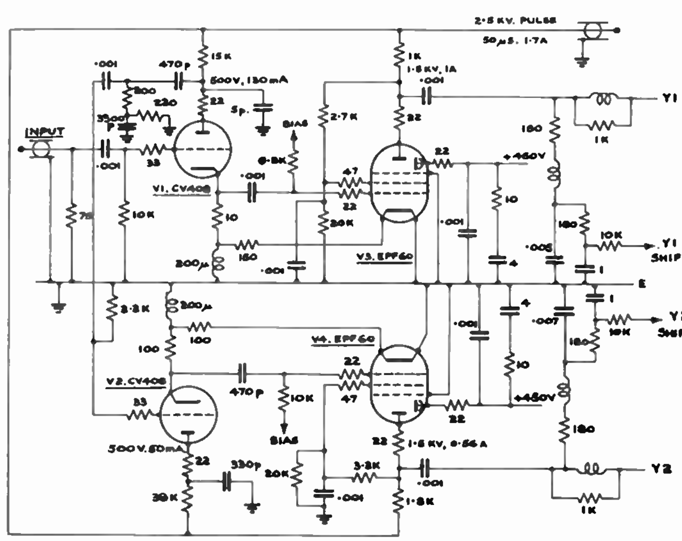


Fig. 5. Pre-pulsed output amplifier.

amplifiers of this type designed by the author require a pre-pulse occurring 50 microsec prior to the signal, with a latitude of the order of $\pm$ 15 μsec. These amplifiers take about 20 microsec to settle down after being switched on, sarily poor but this is not usually serious, since in order to make use of the wide bandwidths the maximum trace duration is restricted to about 100 times the rise time (i.e. to about 0·3 μsec). The pulse repetition frequency is also necessarily low, of the order of a few pulses per second.

Figure 4 shows the circuit of a pre-amplifier employing pre-pulsed EFP60s. This amplifier has a linear output of about 15V into a 75-ohms load. The bandwidth is about 220 Mc/s for −3db, and the gain of the amplifier is 12:1. It is difficult to obtain a bandwidth much in excess of this value with EFP60s, due to the relatively high lead inductances. The cathode lead inductance is about 0·03 microhenries with both leads strapped together. Fortunately in the EFP60 the mutual conductance for cathode current is much lower than that for anode current and the effect of the cathode inductance is thereby reduced.

Figure 5 shows the circuit diagram of the complementary output stage, employing two pre-pulsed EFP60s in push-pull preceded by a phase splitter which is also pre-pulsed. The bandwidth in this case is restricted by the high load which is necessary in order to obtain sufficient output, and is 150 Mc/s for −3db. The linearity curve of the output stage is given in Fig. 6, which shows that the amplifier possesses good linearity for negative pulse input for an output of up to 200V. However, although the amplifier was designed for this polarity, it is able to give a linear output of about 150V for positive input. The gain of the output stage is about 30:1.

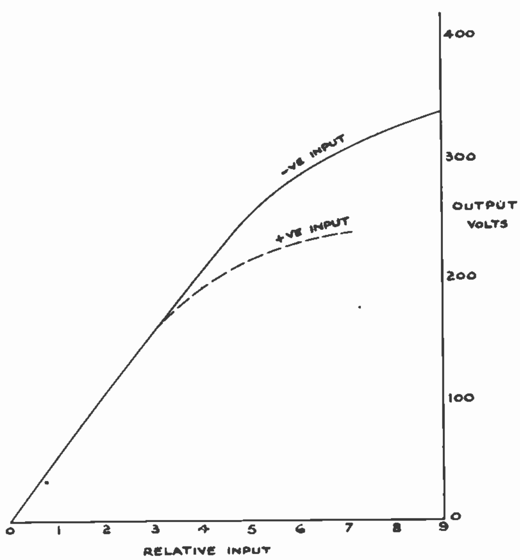


Fig. 6. Linearity curve of pre-pulsed output amplifier under pulse input conditions.

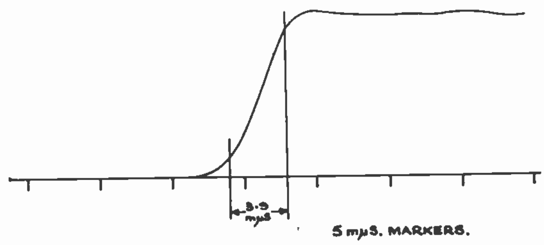


Fig. 7. Step response of complete amplifier.

Figure 7 shows the pulse response of the complete amplifier, including cathode-ray tube. The rise time (10 per cent.-90 per cent.) is 3·9 mμsec, compared with 2·9 mμsec of the output stage alone. Low frequency compensation has been employed as far as possible, and the top of a 250 mμsec pulse is flat within 10 per cent. The gain of the complete amplifier is about 360:1. The bandwidth of the output stage may be increased if a lower linear output is permissible. A bandwidth of 300 Mc/s has been obtained for a single valve with a gain of 4, using dynode and anode to supply a push-pull output.

4. Time-base, Trigger and Beam Brightening

A potential of 400V is required in the case of the VCRX434 to give a deflection of 200 trace widths in the X direction. For time-base durations of 100 mμsec and above, the familiar "bootstrap" circuit may be designed to operate satisfactorily. Under 100 mμsec, however, the d.c. requirements tend to become excessive and transit time effects, etc. cause non-linearity of the trace. A pre-pulsed bootstrap circuit may be used to prevent the use of large steady currents. However, if one is able to make use of pre-pulse techniques, the simple time base shown in Fig. 8 has many advantages. The valve shown is pre-pulsed into the required current and then, when steady conditions have been attained, it is cut off by a trigger pulse. The anode voltage rises at a rate determined by $dV/dt = I/C$, and this rate may be kept substantially constant for the duration of the trace if the inductance L is made large enough. The excursion is in fact the first part of a damped sine wave and a linearity of 1 per cent. may be achieved if the maximum angle θ is equal to $\cos^{-1} 0{\cdot}99$ i.e. 0·141 radians. The inductance L required is equal to τ^2/θ^2C

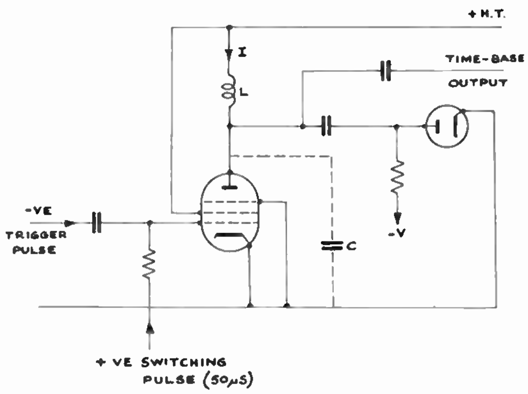


Fig. 8. Pre-pulsed time base.

where τ is the time-base duration, e.g. for

$$\left.\begin{array}{l} \tau = 40 \text{ m}\mu\text{sec} \\ \theta = 0{\cdot}141 \text{ radian} \\ C = 40 \text{ pF} \end{array}\right\} \quad L = 2 \text{ mH}$$

By keeping I, V and the ratio L/C constant, the same linearity is maintained for different time base speeds. A sharp pulse is required to cut off the time-base valve, otherwise non-linearity is introduced at the start of the sweep. Thereafter the sweep is independent of the valve characteristics and this is one of the advantages of the time base. Another advantage lies in the freedom from valve transit time effects, since no feedback system is involved in order to maintain the linearity of the sweep. When the amplitude of the sweep has reached a sufficient level, it is clamped by a diode, thus preventing excessive voltages from being produced at the electrodes. Using EL81s, times down to 5 mμsec have been obtained for a 7 cm trace. EFP60s have been used to produce a symmetrical sweep, with a coil in both anode and dynode. Allowance has to be made for the unbalance of anode and dynode currents and stray capacitances. This valve has the advantage of a short grid base, and it is therefore easier to cut off quickly than most other valves, though it tends to oscillate as soon as it is switched into current, an effect believed to be due to the negative resistance of the dynode during part of the dynode excursion. Pre-pulsed time-bases of the type described above are not suitable for slow sweeps, because the required inductances become so large that the anode current cannot be established in the required time. For a 50 μsec pre-pulse, the limit is about 250 mμsec. However, at this speed conventional time-bases can be designed with ease.

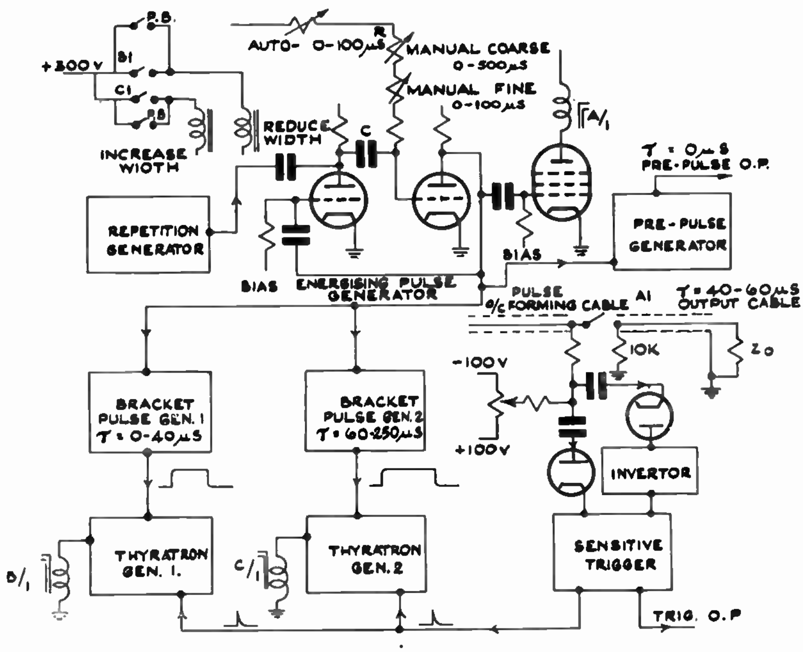


Fig. 9. High-speed pulse generator for pre-pulsed displays.

For most purposes, the pulse required to cut off the time-base valve may be obtained from a sensitive high speed EFP60 trigger circuit, of the type described by Moody, McLusky and Deighton.[3] This valve may be used to supply the beam-brightening pulse to the cathode of the tube, which in the case of the VCRX434 requires to have an amplitude of about 120 volts. It is necessary to ensure that the beam brightening waveform has nearly reached its maximum value before cutting off the time-base valve. To terminate the beam brightening, a beam blanking valve may be used which triggers when the time-base waveform reaches its maximum value. The pulse thus produced is fed to the grid of the tube, and is required to be of greater amplitude and duration than the beam brightening pulse.

5. Pulse Generators

High-speed pulse generators and simulators of many descriptions can be improved with the aid of pre-pulse techniques. The resultant higher mutual conductance and also the higher currents with which to charge stray capacitance lead to much shorter switching times. The rise times thus produced, however, are still longer than that which can be obtained with a relay pulse generator, which is a fraction of a millimicrosecond. In this generator a length of charged cable is discharged into the output cable by means of mercury-wetted relay contacts (described by Lewis and Wells.[4]) The problem in using this type of pulse generator to test pre-pulsed amplifiers is that a pre-pulse is required in the order of 50 microsec before closure of the contacts. Fortunately the mercury relays available have a nearly constant time of closure if the energizing coil is supplied with a high current pulse of constant amplitude. The short term variation is of the order of a microsecond, although the time of closure is about 1½ milliseconds. Figure 9 shows the block diagram of a relay pulse generator which has been designed for use with pre-pulsed displays. The relay energizing pulse is terminated 50 microsec before closure of the contacts, and is used to supply a pre-pulse. To take care of long term drift, if the pulse occurs outside a predetermined time bracket, a servo operates which either reduces or increases the width of the energizing pulse and thus corrects the delay. Coarse and fine manual controls are available for centralizing the range of the servo.

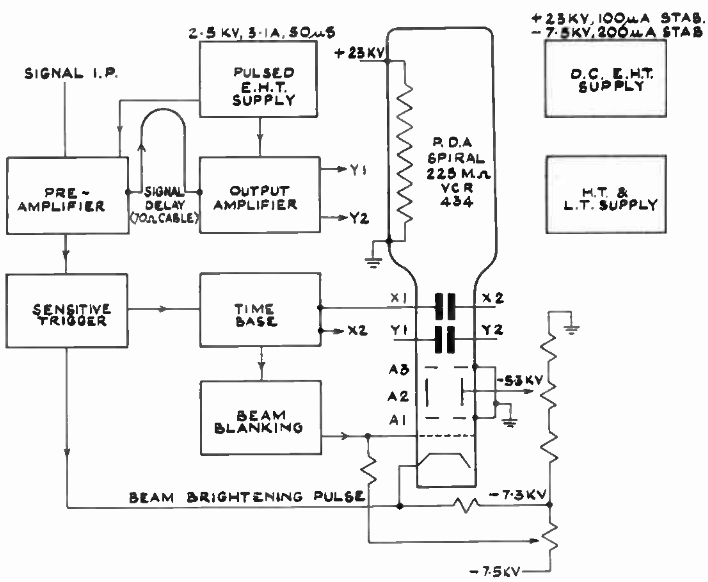


Fig. 10. Block diagram of complete pre-pulsed ultra-high-speed oscilloscope.

6. A Complete Oscilloscope

Figure 10 shows the block diagram of a complete pre-pulsed oscilloscope which has been designed and is in use in small quantities. The amplifiers used are those shown in Figs. 4 and 5. The time-base utilizes a pre-pulsed EFP60, giving a substantially symmetrical output with sweep speeds variable between 40 and 250 mμsec for 7-cm deflection. The sensitive trigger and beam blanking stages also use EFP60s. Figure 11 shows a photograph of the tube unit,

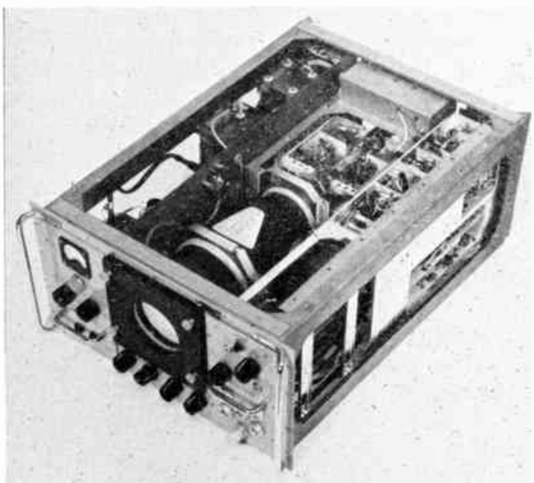


Fig. 11. Photograph of tube unit.

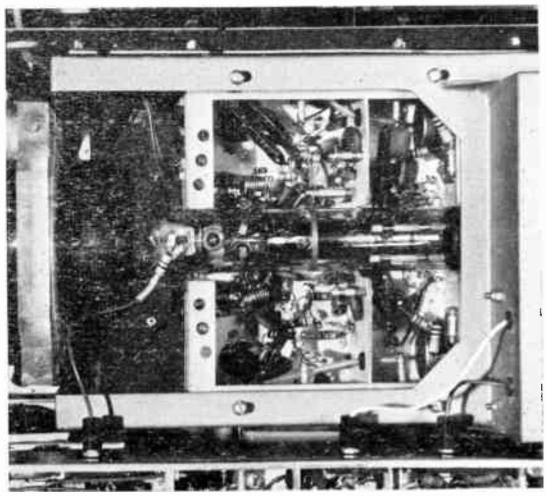


Fig. 12. Photograph of output amplifier.

and Figure 12 shows a close-up photograph of the output amplifier, which is situated under the tube neck to ensure short leads. The pre-amplifier, time-base, trigger and beam brightening stages are built on a removable sub-unit on one side of the tube unit, whilst the e.h.t. supplies for the tube are obtained from a sub-unit on the other side of the tube. The pulsed e.h.t. supply for the amplifiers is obtained from a separate unit. Although the pulse current is high, the mean current is low, due to the low repetition rate of 1 or 2 c/s. The pulsed e.h.t. supply unit is approximately one half the size of the tube unit and it contains an r.f. oscillator e.h.t. supply, giving a d.c. output of about 1 mA and a series pulse modulator valve which is operated by thyratrons. A good part of the unit is taken up with the reservoir capacitor which supply the pulse current.

Since this oscilloscope was designed, a more compact pulsed power supply has been developed. This power supply does not require a separate d.c. e.h.t. supply, and operates from the normal h.t. lines, resulting in a saving of space. The power supply and the complete amplifier and trigger stage are now contained in a sub-unit measuring 10 in. × 4 in. × 22 in. long. The amplifier has also been revised in order to reduce the e.h.t. requirements to a certain extent, but it still has substantially the same characteristics. The output signals are conveyed to the deflector plates through two 120-ohms cables. The circuit of the new power supply is shown in Fig. 13. It operates as follows. The valves V1 and V2 comprise a bootstrap circuit, with the anode potential of V1 near to that of the negative line. A negative input pulse cuts off V1 for a time interval of 80 microsec (controlled by C1, R6). and its anode potential rises at a rate determined by the original anode current and the stray capacitance. As it does so it raises the grid potential of the cathode follower V2, which in turn drives the cathode follower V3. This valve raises the anode potential of V2, and prevents V2 from bottoming. The grid potential of V3 is clamped by X3 when it has risen by 350 volts, thus stabilizing the amplitude of the pulse fed back to the anode of V2. The rise of V1 anode potential, which

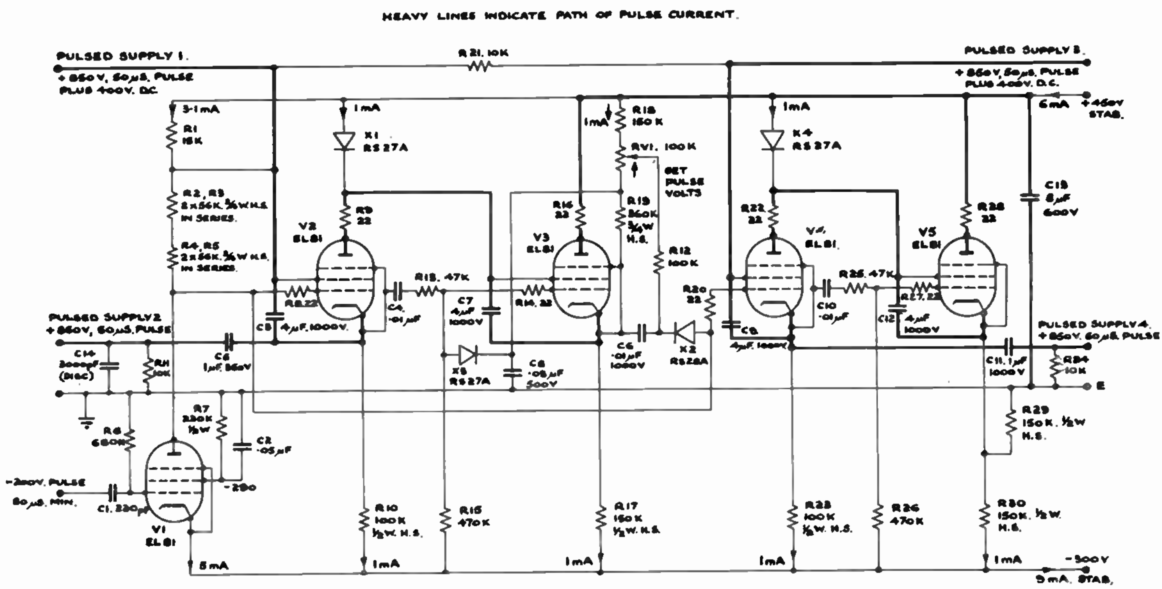


Fig. 13. Pulsed e.h.t. supply.

is maintained at a substantially constant rate by feedback from the cathode of V2, is finally terminated by X2 at a level determined by RV1. The pulse potential on the cathode of V3 is also applied to the crystal, so that the potential at which clamping occurs is about +600 volts. The total excursion of V1 anode is thus about 900 volts, and an output e.h.t. pulse is taken from the cathode of V2. This is supplied into two output leads. The main output is at a d.c. potential of +400 volts resulting in a total potential of approximately +1250V during the pulse. The other is at zero d.c. potential, and is used for feeding the trigger stage and the screens of the amplifying valves. A pulse current of approximately 1·2 amperes is taken from the combined lines. Two more outputs are available from the cathode follower V4, which operates in a similar manner to V2. The anode potential of V4 is raised during the pulse by the cathode follower V5. The heavy lines on the circuit diagram show the path of the pulse currents, which are supplied by C13 through V2 and V3 in series, or V4 and V5 in series.

7. Equipment for Testing Valves under Pulsed Conditions

The pulse duration with which this paper is concerned is in the order of 50 microsec. In order to test a valve under pulse conditions it is necessary to apply a pulse to one or more electrodes. For measurements where the valve is capable of withstanding the full operating potentials under cut-off d.c. conditions, only the control grid need be pulsed. The pulse may be applied through a resistance in the order of 100,000 ohms, limiting the potential on the grid by a clamping diode. The grid characteristics can then be determined for different electrode potentials by varying the bias on the diode, and hence the grid potential during the pulse. The various electrode currents are determined by inserting small series loads and measuring the pulse voltage across them with a general purpose oscilloscope. To speed up the measurements a sawtooth voltage may be applied simultaneously to the grid of the valve and to the X plates of an oscilloscope. This enables the complete grid characteristics to be observed visually or recorded photographically.

A versatile instrument capable of making pulse measurements on a wide range of valves and over a wide range of electrode potentials and currents is necessarily complex. At least one instrument has been produced commercially which will make measurements up to an anode current of 5 amperes at an anode potential of up to 450V. For higher voltage and current measurements it was necessary for a new instrument to be designed. The full details of this instrument are outside the scope of this paper, but the general principle of operation may be followed with reference to Fig. 14. The unit contains a 5-inch diameter cathode-ray tube, on which the anode, screen, control grid or dynode current of the valve under test can be displayed as a function of anode, screen or control grid potential. In addition either of the remaining two electrode potentials not being swept may be varied up to a maximum of ten steps, thus producing a family of curves. The anode and screen grid voltages may be swept or stepped up to a maximum of 3·5kV in six ranges, whilst the total range covered for the control grid is −100V to +100V. The current range of the instrument may be varied over wide limits up to a maximum of 10A anode current, 1A screen current, 5A dynode current and 0·5A grid current. Three series modulators are used to supply the anode, screen and control grid potentials. Each electrode current is supplied during the pulse from a separate reservoir capacitor, and is monitored by amplifying the small voltage pulse developed across a resistor in series with the capacitor. The gain of the amplifier is such that an input pulse of 1 volt amplitude will give rise to full deflection on the c.r.t. face. The anode and screen reservoir condensers are charged by a 5·7kV d.c. e.h.t. supply, whilst the control grid reservoir capacitor obtains its charge from a 300 volt line through a high resistance. The modulators are controlled by sawtooth, stepping or d.c. potentials as required. Each modulator is, in fact, the output stage of a d.c. amplifier, the gain of which is accurately controlled by varying the negative feedback ratio to the input stage (a long tail pair). This varies the voltage range covered by the instrument. The maximum amplitude of the control potentials fed to the modulation amplifiers is 100 volts, and the d.c. and step potentials are monitored by an internal meter. The sawtooth control waveform has a duration of 100 microsec, rising from 0 to 100 volts in this time. It is also applied through an amplifier to the X plates of the c.r.t. The X and Y amplifiers are calibrated by an internal pulse generator, which produces a pulse of precisely 100 volts amplitude. This pulse is attenuated to a level of 1 volt for the calibration of the Y

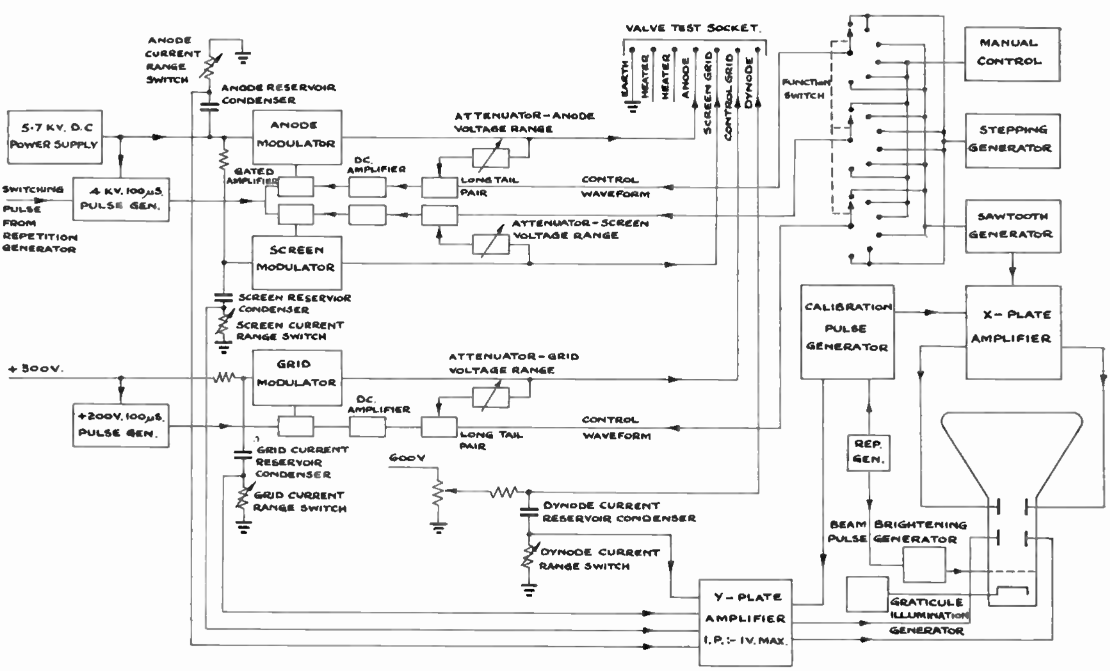


Fig. 14. Pulsed valve tester.

amplifier, and the amplified pulse produces two spots on the tube face, corresponding to the maximum current range of the instrument. The gain of the Y amplifier is adjusted to cover a suitable range on the graticule. In a similar manner the pulse is used to calibrate the X-amplifier. The spots correspond in this case to a change of 100 volts supplied to the modulation amplifiers, and thus to the voltage range of the instrument, set by the negative feedback attenuators. During calibration the instrument is operated at a repetition rate of about 25 c/s, but during valve tests it is reduced to about 1 c/s. It is therefore necessary to take a photographic record, in order to study a family of curves satisfactorily.

8. Future Developments

The paper has mainly described the use of EFP60s in pre-pulsed applications because to date it has proved to be the valve with the highest figure of merit. Microwave valves have also been used, and it will be necessary to rely on development in this field if bandwidths are required greatly in excess of those so far achieved. Since the original work was carried out on pre-pulsed valves, types have been produced with very fine wires and close cathode-grid spacing which have a very high slope under d.c. conditions. These valves may possibly replace the EFP60 for pre-pulsed cascade amplifiers. However, the ultimate solution lies in the design of valves for this application, i.e. for pulse lengths in the order of 50 microsec. Relatively little work appears to have been carried out in this field and it should be stressed that the valves are at present being used under conditions not envisaged by the manufacturer.

The pre-pulsed amplifiers described above show that cascade amplifiers can be designed of comparable bandwidth with that of existing distributed amplifiers. By combining the two techniques it should be possible to design amplifiers with bandwidths in excess of 500 Mc/s and with a high voltage output, using microwave valves.

9. Acknowledgments

The author wishes to acknowledge the contributory work in this field of Messrs. R. A. Fothergill, J. R. Truscott, and Mr. J. W. Birtill, all of the Atomic Weapons Research Establishment. Mr. Birtill was particularly concerned with the design of the pulsed valve tester.

10. References

1. S. D. Abercrombie, "Travelling wave oscilloscopes," *J. Brit.I.R.E.* (To be published).
2. F. E. Whiteway, British Patent Application No. 23402/58.
3. N. F. Moody, G. J. R. McLusky, and M. O. Deighton, "Millimicrosecond pulse techniques Part 1," *Electronic Engineering*, **24**, pp. 214-9, 1952.
4. I. A. D. Lewis, and F. H. Wells, "Millimicrosecond Pulse Techniques," pp. 100-107 (Pergamon Press, London, 1954).

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.

WAVEGUIDE COMPONENTS

Methods for dividing power at high levels, especially with power dividers of the hybrid connection type, are considered in a recent Czech paper. The computation of a hybrid connection is carried out; its properties described and results of experimental measurements presented. A detailed analysis of a power divider consisting of two hybrid connections and a dielectric phase-shifter is then given. The theoretical results are verified by measurement. A further divider consisting of three hybrid connections is also discussed. In conclusion power dividers are evaluated with regard to power handling capacity and use.

"A variable high power microwave divider." K. Varecha. *Slaboproudy Obzor* (*Prague*), **20**, pp. 292-295, May 1959.

MICROWAVE INTERFEROMETRY

A microwave interferometer has recently been described by a professor of telecommunications from Brazil. Detailed analysis is presented for the reflection of microwaves at plane surfaces and the generation of high-order-modes by interference. The basic experimental system comprises a pair of parallel rectangular plates with variable separation, arranged symmetrically with respect to a reference plane and a third parallel rectangular plate whose position is adjusted independently, to effect the auxiliary reflection required to produce interference. Horns with variable angular position are used as signal source and detector.

"The high-order-mode interferometer." J. I. Caicoya. *L'Onde Electrique*, **39**, No. 385, pp. 321-328, April 1959.

D.C. AMPLIFIERS

Two methods of measurement of weak direct currents using high value resistors and integrating capacitors are considered in a recent French paper. In the case of a single stage using an electrometer valve, the linearity is poor and the indicating instrument must be sensitive and therefore fragile. These drawbacks can be overcome by using several stages; negative feed-back applied to such an amplifier reduces the input impedance and makes the reading independent of the gain; it also keeps the grid current error at a constant value and increases the band width whilst reducing the apparent drift. The author studies an amplifier having a very low drift, with a good bandwidth due to direct coupling, the drift being reduced by the use of an a.c. amplifier with a modulator at the input and a demodulator at the output.

"The measurement of weak direct currents by direct coupled amplifiers." G. Friedling. *L'Onde Electrique*, **39**, No. 385, pp. 329-337, April 1959.

TRANSISTOR THEORY

In a paper from the Institut für Elektrotechnik, Zurich, the static current amplification factor is calculated for an example of a transistor with diffused base region. The results are discussed and compared with calculations of transistors with homogeneous base regions. The emitter efficiency of transistors with homogeneous base region is analysed for high injection densities. These results, obtained by calculations, are compared with experiments. Many measurements for *p-n-p* and *n-p-n*, germanium and silicon transistors are shown in order to give some information about the product of emitter conductivity and minority carrier diffusion length in the emitter, for different doping elements.

"The static current amplification factor as a function of emitter current for transistors with diffused and homogeneous base region." Peter Kaufmann. *Archiv der Elektrischen Ubertragung*, **13**, pp. 143-151, April 1959.

TRANSISTOR AMPLIFIER

In a recent German paper the relationship of the "white" noise factor as a function of generator impedance R_G and emitter current I_E in transistors is used to derive values for which the noise factor becomes a minimum. Furthermore, equations for the most favourable combination of both parameters and the appropriate absolutely minimized noise factor are quoted. These values are calculated and measured for a number of transistors within wide variations of the transistor parameters (short circuit current gain α, base impedance R_B, collector saturation current I_S) and as a function of temperature (between $-90°$ C and $+70°$ C). The calculations and the measurements are in good agreement.

"Calculations and measurements for an optimum design of a low noise transistor amplifier." K. Spindler. *Nachrichtentechnische Zeitschrift*, **12**, pp. 250-256, May 1959.

MEASUREMENTS ON VALVE CATHODE MATERIALS

For studying the electrical properties of oxide cathodes a test electron tube has been developed by Siemens & Halske on which both the interface resistance and the bulk resistance can be measured under normal operating conditions. The potential drop developing in the oxide compound can be determined from the final velocity of the emitted electrons with the aid of a retarding field where the electrons are retarded after passing a slit in the anode. This arrangement allows also an investigation of the time behaviour of the electrical parameters of the oxide cathode as a function of temperature and load. Curves of the useful life have so been obtained for a number of active and passive nickel grades. Finally the paper deals with the difference between temperature and current activation.

"Study and assessment of oxide cathodes by means of a test electron tube." G. Landauer and W. Veith. *Archiv der Elektrischen Ubertragung*, **13**, pp. 211-218. May 1959.

MATERIALS FOR ELECTRON TUBES

Unwanted thermionic emission from electrodes other than the cathode may impair the operation of vacuum tubes. An investigation has been made by N. V. Philips into the emissive properties of various materials widely used in the construction of electron tubes. Special tubes were built for this purpose, care being taken to ensure that all parts were scrupulously clean. To simulate practical operating conditions, the cathode evaporation products were allowed to form deposits on the grid. Measurements were also carried out at different operating temperatures. Three types of cathode were used, namely impregnated cathodes, L cathodes and oxide cathodes. Of all materials tested, clean titanium was found to have the lowest emission. It shows a much greater emission, however, if coated with titanium oxide.

"Electron emission of materials for electron tubes." G. A. Espersen and J. W. Rogers. *Philips Technical Review*, **20**, pp. 269-274, No. 9, 1958/59.

VIDICON LAG

Investigations of the problem of the delay effects in vidicon tubes have previously attempted to explain these phenomena mostly by analysing the discharge mechanism of the storage elements in terms of circuit techniques. A contribution by a German engineer reports experimental results that reveal that processes inside the semi-conductor are also responsible in great part for the disturbing phenomena. In the course of these investigations a formation of the layers in the completed tube was found that allows a considerable reduction of the delay effects without impairing the sensitivity.

"Delay effects in camera tubes of the vidicon type" W. Heimann. *Archiv der Elektrischen Ubertragung*, **13**, pp. 221-225, May 1959.

FIELD INTENSITY RECORDING

A recent Czech paper describes a mechanical system of automatic regulation of receiver sensitivity in which a waveguide attenuator controlled by an integrating motor is used as the regulating element. Position of the attenuator piston is a linear function of the signal level expressed in decibels. The system equation is derived and, in spite of the fact that feedback affects the system parameters, is a linear differential equation of the first order valid for any input signal in general. The system described is suitable for field intensity recorders. The paper also points out the possibility of using it in long-term fading research of slowly varying mean (mean-square) value of the received signal.

"A mechanical method for the automatic regulation of sensitivity of field intensity recorders." J. Karpinsky. *Slaboproudy Obzor*, (*Prague*), **20**, pp. 278-284, May 1959.

SINGLE SIDEBAND COMMUNICATION

The advent of single side-band systems with suppressed carrier calls for a high precision variable frequency pilot. The frequency precision required for such a system is at the maximum 50 c/s and existing pilots (stabilidyne, I.G.O. piezo-variateur) cannot be used as their stability is of the order of hundreds of cycles. The advantages of high stability pilots are reliable supervision, speed of connection, conservation of pitch, reliability, elimination of personal errors. A minimum number of frequency changes was required in order to avoid undesirable interference, which must be reduced to 60 db. Thus the synchronizing frequency is obtained after four frequency changes, whereas the synchronized frequency is obtained after two changing-processes for the transmitter and none for the receiver. This high stability pilot comprises three units: a reference oscillator and its dividers giving a precision of 5×10^7, an l.f. loop covering a range of 100 kc/s in kilocycle steps and introducing an error of 0·5 c/s, an h.f. loop covering a range of 1·5 to 24 Mc/s in 100 kc/s steps and introducing no error. The two loops synchronized by output frequencies of the dividers are operated in series and give any number up to 5 figures. (3 for the h.f. and 2 for the l.f. coupler). The locking of each loop is obtained by a phase discriminator which adjusts the oscillator by the aid of a reactance tube. The whole equipment is remotely controlled by means of a special 5 figure counter and the time required to obtain a frequency is between 2 and 20 seconds.

"A high stability pilot: the synchrocompteur." L. Berman. *L'Onde Electrique*, **39**, pp. 392-398, May 1959.

MEASUREMENTS ON MUSIC CIRCUITS

A test unit for the measurement of non-linear distortions and for use mainly in broadcast music circuits is described in a German paper. The single tone method is used for measurements in the lower frequency range while the double tone method is used in the upper frequency range. The equipment is also suitable for testing individual repeaters. A faulty repeater along a line with many repeater stations can quickly be found by means of a combination frequency of the form $2\omega_1 - \omega_2$ and in a better way than with existing equipment.

"A non-linearity test unit for broadcast music circuits." E. A. Pavel and M. Bidlingmaier. *Nachrichtentechnische Zeitschrift*, **12**, pp. 243-249, May 1959.

INTERFERENCE MEASUREMENT

The relative radio interference produced by high voltage line equipment may be determined by coupled measurements in the laboratory. An account has been given of the development of radio noise testing facilities at the National Research Council Laboratories in Ottawa and the methods adopted to minimize interference from the test equipment. Two simple noise measuring circuits which compensate for stray and coupling capacitance are discussed.

"Radio noise testing of high voltage line equipment." A. S. Denholm. *Transactions of the Engineering Institute of Canada*, **3**, No. 1, pp. 13-17, April 1959.

TELEVISION INTERFERENCE

Television pictures may be marred by parasitic signals which fall into three classes, pattern interference and either continuous or impulsive random noise. Statistical tests enable the relationship between the nature of the noise and the subjective disturbance it produces to be established. In the particular case of sinusoidal intereference the annoyance can be presented by a family of curves of equal annoyance. These results permit the calculation of weighting networks, whose use has been standardized by the C.C.I.R. and C.C.I.T.T. It is very important to appreciate the definitions and methods of measurement in order to avoid any error in the interpretation of the results.

"Interference due to parasitic signals on television pictures." L. Goussot. *L'Onde Electrique*, **39**, pp. 352-361, May 1959.

TROPOSPHERIC SCATTER PROPAGATION

The installation of a tropospheric scatter radio system to provide telephone and miscellaneous communication services to Schefferville and Goose Bay and to other locations in Northern Quebec and Labrador was undertaken by the Bell Telephone Company of Canada in association with Quebec Telephone Company. A paper by the engineers concerned reports on the engineering and construction of the project with particular emphasis on the broad planning of the system and on the construction aspects. The first part deals with the communication systems engineering aspects. The equipment operates in the frequency range 755-780 Mc/s and is a broadband f.m. system. There are six stations and the path lengths of the various links are between 90 and 140 miles. The second part of the paper deals with the design and construction of buildings, towers, roads, etc.

"Quebec-Labrador tropospheric scatter radio system." D. J. McDonald and C. E. Frost. *The Engineering Journal of Canada*, **42**, pp. 43-57, April 1959.

TELEVISION TRANSMITTERS

The last step in providing a television service is covering the shadow zones, which, because of their unfavourable geographical situation, fail to receive an adequate signal from the main transmitters. The solution of this problem is in the use of translators which re-radiate the received signal in a different channel without intermediate demodulators. Two papers in *L'Onde Electrique* show how French manufacturers have approached these problems.

The characteristics required of television translators intended for permanent installation and for unattended operation, involve special problems, partly from the point of view of technical features and partly from the point of view of reliability. Further, it is necessary to cater for a large number of combinations of send and receive channels in Bands I and III.

The designs adopted to satisfy the requirements include: direct transposition from receive to sending channel without change to intermediate frequency, the use of miniature resonant cavities, and, when the powers involved require it, the separation of the sound and vision by coaxial systems using hybrids.

"Television re-transmitters (translators)." M. Boxberger. *L'Onde Electrique*, **39**, pp. 362-367.

"Television re-transmitter (translators) of 300 milliwatts and 3 watts." M. Nadeau. *L'Onde Electrique*, **39**, pp. 368-373, May 1959.