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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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THE ARMORIAL BEARINGS OF THE INSTITUTION

OVER the centuries a badge or symbol has been used as a means of identification, and there are numerous early examples of what has been described as "picture writing" to distinguish a man from his fellows. It was not, however, until the 12th century that there was any systematic treatment of heraldic devices, first as a means of identifying families, then of ancient guilds and subsequently universities and corporate bodies.

The incorporation of the Institution required that it should have a seal, but at that time, 1932, the procurement of a coat of arms was hardly attainable. The emblem adopted did little more than convey the Institution's association with radio, or wireless science, and was not in any way a proper heraldic device.

In more recent years the increasing status of the Institution has caused successive Councils to consider petitioning for a Grant of Arms which would indicate the Institution's objects and status. On 3rd March, 1955, it was resolved that a "Memorial" or formal application for a grant of arms be addressed to the Earl Marshal of England reciting the incorporation of the Institution and that one of its objects is the advancement of the science and practice of radio and electronic engineering.

The petition having been approved, the Council of the Institution was required to decide the form or representation of the armorial bearings. As already stated in the 30th Annual Report of the Institution, invaluable advice and

help was given by Mr. Anthony R. Wagner, C.V.O., Richmond Herald. To those thanks, the Council wishes to add its appreciation of the work of Mr. Graham D. Clifford, the General Secretary of the Institution, whose original suggestions ultimately led to the final design of the Armorial Bearings.

Whilst the activities of the Institution could have been represented by some of the conventional devices of heraldry, the Council was unanimous in agreeing that the Armorial Bearings, so far as the shield was concerned, should allude to the pioneers of the science, rather than consist of direct allusions to radio and electronic engineering. In like manner the Supporters pay tribute to the encouragement given to the advancement of radio science by the 7th Duke of Devonshire and Admiral of the Fleet the Earl Mountbatten of Burma.

By Letters Patent under the hands and seals of Garter, Clarenceux, and Norroy and Ulster Kings of Arms dated 10th February, 1956, the Institution was granted the Armorial Bearings and Supporters which are illustrated on the opposite page. The blazon, or official description of the arms, crest and supporters, which together form what is known as an "achievement," is as follows:—

Azure on a Pile Or a Fess Chequy of the first and Argent between in chief a Heart between two Crescents Gules and in base a Boar's head erased Sable; and for the Crest, on a wreath of the Colours, A representation of the head

and shoulders of Mercury proper the Hair Or wearing a Petasus also proper Winged Gold; the Supporters being On the dexter side a Buck proper and on the sinister side a Lion queue fourchee Barry of ten Argent and Gules.

Thus the great contribution of Clerk Maxwell to radio and electronic science made it appropriate to base the shield design on an allusion to his family Arms, these being *Or a Fess Chequy Azure and Argent between two Crescents in chief Gules and a Boar's head couped in base Sable*. From these the fess, crescents and boar's head have been abstracted and appear on the gold wedge shape (pile) on the shield together with a heart. The latter is a canting allusion to the name of Heinrich Hertz, whose work in the field of radio propagation was based on the scientific discoveries of Clerk Maxwell.

The shield itself has a background of blue to represent the sky and the atmosphere through which radio waves are transmitted.

The head of Mercury, the messenger of the Gods, is introduced as the crest in allusion to the concern of the Institution with the promotion of the science of radio communication.

By permission of His Grace the Duke of Devonshire, the dexter supporter, a Buck proper, is directly derived from the Arms of the Devonshire family. In those, the supporters are *A Buck proper wreathed about the neck with a chaplet of roses alternately argent and azure*. In the Institution's Arms, the dexter supporter has been modified by the omission of the chaplet.

In this way the Institution's Arms commemorate the work of William Cavendish, K.G., F.R.S., the 7th Duke of Devonshire. He was Chancellor of the University of Cambridge and provided the endowments which led to the establishment of the Cavendish Laboratory and the Chair of Experimental Physics to which Clerk Maxwell was appointed in 1871.

These notes are based on various reports given to the Council by the General Secretary of the Institution.

In the inaugural Clerk Maxwell Memorial Lecture to the Institution,* Professor G. W. O. Howe referred to the fact that the 7th Duke of Devonshire had been a member of the Royal Commission on scientific education, and that towards the end of his life Maxwell's chief literary work was the editing of the electrical researches of the Hon. Henry Cavendish, F.R.S., a great-uncle of the 7th Duke.

Admiral of the Fleet the Earl Mountbatten of Burma, K.G., also consented to some portion of his own Arms being incorporated in the Institution's Armorial Bearings. This has been expressed in the sinister supporter which is based on the lion, fork-tailed and crowned, which appear in the first quarter of the Arms of the Institution's Vice-Patron.

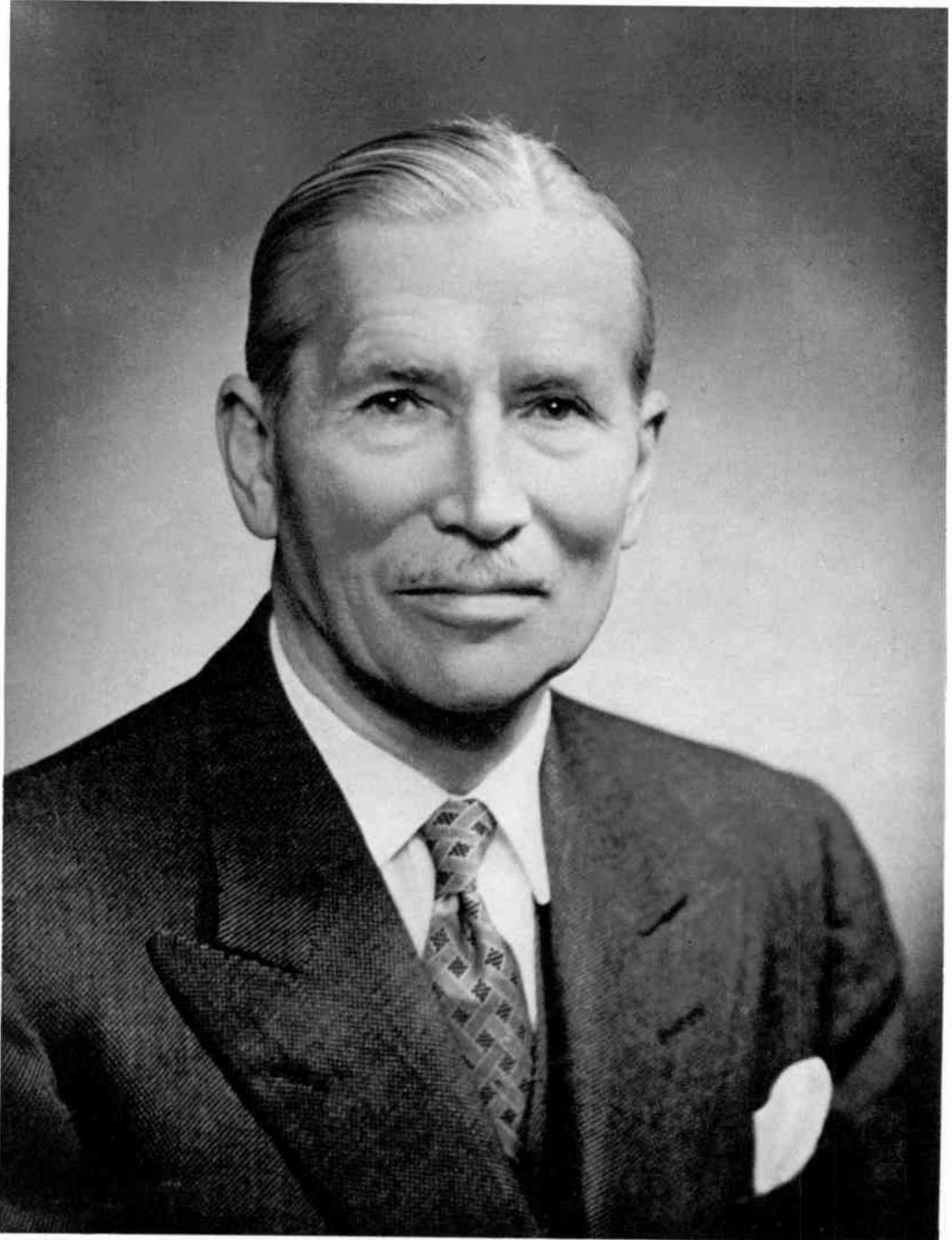
Thus, there is heraldic record of Admiral Mountbatten's interest and assistance in the Institution's work from the early days of its foundation by his membership, his election as a Vice-President, and ultimately his office as ninth President. Subsequently, Lord Mountbatten was appointed Vice-Patron of the Institution in 1950.

The space or mound beneath the shield—termed "the compartment"—is composed of mountain peaks surrounded by clouds in development of the atmosphere theme and the reflection of radio waves therefrom.

When the Institution first adopted a seal which emblematically conveyed the Institution's association with radio science, the motto used was "Pace omnes terras tegetus." As shown in the Armorial Bearings, opportunity has now been taken to adopt the motto *Scientia pro hominibus* which may be interpreted as "Science for the benefit of Mankind." This is the principal aim of all learned societies and the foundation upon which the Institution will base its petition for a Royal Charter.

G. A. TAYLOR.

* *J.Brit.I.R.E.*, II, p. 545, December 1951.



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GEORGE ARMSTRONG MARRIOTT, B.A. (Cantab.)
The Fourteenth President of the Institution.

THE PRESIDENTIAL ADDRESS

of

George A. Marriott, B.A.

*Delivered after the 31st Annual General Meeting of the Institution held in London,
on October 31st, 1956*

THE STATUS which the Institution now enjoys enhances the honour of being elected its President. Whilst expressing my thanks for the honour extended to me, I wish also to pay tribute to my predecessors in this office who have done so much for the Institution. In particular, I must express my appreciation of the encouragement and example which Admiral Sir Philip Clarke has given to us all during his Presidency.

When Sir Philip referred to me as the 14th President of the Institution, my first thought was that whilst there is still much for us to do, we can no longer be considered a junior amongst Institutions. Indeed, the Brit.I.R.E. is now approaching its thirty-second year and has existed almost since the inception of the British Radio Industry. So, within this comparatively young Industry, we are an old established body.

During that time we have witnessed the evolution of an important industry, stretching far beyond the field of communication with which it was at first concerned.

My own professional life started shortly before the Institution was founded and has almost wholly been spent in the Radio Industry. I was fortunate because at the beginning of my career considerable attention was being given to the comparatively new developments then taking place in vacuum physics technique and my company was a pioneer in that field.

Attention was then largely centred on the possibilities of a public broadcasting service. Interest in this possibility had been stimulated by the developments during the 1914-18 war when, as in the second world war, a very intensive research programme hastened the evolution of new scientific and engineering techniques, especially in "wire-less" reception.

I do not intend to review the history of electronic development; that has been done in a number of forms by people much better qualified than me—but I do propose to illustrate my remarks by reference to valve development, which is the section of the industry with which I am best acquainted.

It is significant that 40 years ago the main application of radio was to the needs of the military services; today, in addition to radio and electronic equipment being an indispensable item of defence expenditure, the annual turnover of the British Radio Industry is in the neighbourhood of three hundred million pounds—its export figure alone being in excess of forty million pounds.

This latter figure is twenty times greater than that for the immediate pre-war years, and is but one example of the contribution of the radio engineer to the economic life of the country.

It will be obvious, therefore, that the science of radio and electronics and its engineering applications has, in the lifetime of the Institution, opened up new careers for young men, and presents them with a future which offers unlimited possibilities. I felt, therefore, that I could not do better than devote most of my Address to the future of the young engineer.

For this reason I should like to refer to the circumstances in which original research workers in our field undertook their labours, and to the economic and industrial considerations which make possible the further application and use of radio and electronic engineering.

Much of the development in radio engineering has been due to the invention of the valve and subsequent improvements in valve technology and manufacture. We should remember that

the invention of Ambrose Fleming finally arose out of his association with Guglielmo Marconi at the time of the original wireless or radio transmission across the Atlantic.

Marconi had, of course, been using the coherer as a detector of electromagnetic waves, and Fleming realized that something else was needed. It is of importance to note that Fleming had studied at Cambridge under James Clerk Maxwell and that his early life was influenced by a number of inventions in the communication field, including that of Graham Bell, the inventor of the telephone.

Because of that background, Fleming was especially able to appreciate the significance of the experiments conducted by Heinrich Hertz, which showed the existence of electric waves. In addition, Fleming had spent some time with the Edison Electric Light Company, and this experience, coupled with his knowledge and zest for experimenting, led to the making of the first wireless or radio valve.

In his "Memoirs," Fleming describes in these words how he made use of the special lamps he ordered from the Edison Electric Light Company:

"When I received these bulbs, I made the filament incandescent by the current from six cells of a storage battery or accumulator. I then found that the space between the filament and the cylinder would only allow negative electricity to pass from the filament to the cylinder, but not the other way. Hence the bulb acted like a valve in a pipe, which lets water or air only move in it one way.

"Accordingly, I called this instrument a 'valve'. At first I named it an oscillation valve, but before long it was usual to speak of the small particles of electricity sent out from the hot surface as 'thermions' and accordingly I called it a 'thermionic valve'. This name has been very widely adopted.

"My valve had the great advantage that it could not be injured or set out of action by 'atmospherics', as is the case with the coherer."

Fleming's invention was the subject of British Patent No. 27850 in November, 1904, and the

stages by which Fleming developed what he called the "thermionic detector" was admirably set out in the book he wrote in 1919.*



Fig. 1. An early experimental valve

In his Presidential Address to the Institution in 1944, Mr. Leslie McMichael† suggested that interest in the radio valve languished for some ten years owing to the discovery and use of various kinds of crystals as detectors of wireless signals. Perhaps in the development of the transistor history may be repeated!

The work of Fleming and Lee de Forest was supplemented by the French engineers—Paul Langevin, Peri, and Ferrie. Their combined efforts did much to bring about the first extensive production of radio valves during the first world war, which subsequently enabled radio engineers to solve the problem of using more and more frequencies.

The young engineer might well note that so far improvement in radio technology was largely due to a form of combined operations; from the original theories of Clerk Maxwell and Heinrich Hertz, physicists and engineers developed new techniques principally as a result of studying new ideas, sharing experiences, and

* "The Thermionic Valve" (The Wireless Press, Ltd., London).

† *J. Brit. I.R.E.*, 5, p. 2, January 1945.

not being afraid to experiment. Those qualities are needed just as much today as they were then or at the beginning of the Industrial Revolution.

It seems to have been always fashionable to criticize succeeding generations, but for my part I feel sure that the modern professional engineer is not unaware of his responsibilities in the economic life of the nation. In our own field we know that in the years of growing up our main production centred round a domestic consumer market, and this inevitably led to some instability from one year to the other and from season to season.

New ideas were developed, however, and in less than 20 years the industry ceased to depend entirely on the consumer needs of what was virtually an entertainment product. Instead, the efforts of radio engineers were required to provide electronic equipment for a very wide range of industrial purposes, including atomic energy, radar, and communication at home and with overseas countries.

Valve Production

A significant indication of the growth of the industry is shown by the fact that in 1924 Great Britain produced 2½ million valves. Just prior to the last war this figure had increased to 12½ million. Last year, total production exceeded 64 million valves.

Equally significant is the fact that while valves produced for the broadcast entertainment field in 1955 were approximately five times the 1938 figure, mainly due to television, those produced for the Services were seven times and those produced for industrial purposes thirty times the 1938 figure.

In order to achieve such production in about 30 years, a very considerable capital investment has been required. To set up a modern valve factory would cost in the region of three million pounds—a considerable investment even for a large company. For an outlay of this size to be economic requires that the factory should operate at full capacity and be managed and staffed with the engineers capable of keeping abreast of new ideas and of introducing new manufacturing processes such as are essential to meet competition.

The products of a modern valve factory are highly complex and have used, and continue to require, automation to an extent that is unrivalled by any other industry. Indeed, I might say that automation in the valve industry had to be practised a long time before the word entered our modern language. It is, however, one of the perversities of modern production that the more automatic the process, the greater the need for an expanding market. This is in addition to the constant changes required in production technique because of improved design. Long term foresight and engineering planning are essential because the ultimate cost of valves is largely governed by providing a steady production run which will permit of concentration on mechanization.

I think it is acknowledged that increased production has enabled considerable benefits to be passed on to the consumer, both in terms of price and quality of product.

All these points need study and understanding by the engineer, for however altruistic is our purpose in life, the fact remains that the benefits of science cannot be passed on to the common man unless those products are brought within his financial reach.

The Engineering of New Ideas

I have stressed the need for keeping abreast of new ideas. The first world war proved the possibilities of broadcasting; the second world war hastened the development of radio-location or radar, and affords an outstanding example of the advantages of co-operation between Universities, Government establishments, and Industry.

Here again, experimental work was studied by engineers in order to assess the possibilities of practical use. First of all, we had the fundamental work of Appleton in 1924 on the use of reflected waves in order to calculate the height of the ionosphere; this led to the introduction of pulse modulation, and finally to the Air Ministry's appointment in 1933 of a group of engineers, working under Watson Watt, to examine the technical possibilities of locating aircraft.

The success of that enterprise resulted in the establishment of the first network of radar stations, and at the end of 1935 the first production contracts were placed with the British Radio Industry.

Again we should note that original work led to studying new ideas, sharing experiences, and not being afraid to experiment.

The engineer has to realize that there is always scope for improvement. Directly radar had reached the production stage it was realized that a new form of valve was required capable of transmitting very strong pulses at what were then regarded as high frequencies. Necessity was truly the mother of invention, for by then the war was close upon us and there was a great strengthening of the ties between the Universities, Government departments and Industry.

It was important that the valves used for radar transmission should be convection cooled and it was originally thought that valves in a silica envelope, such as used by the Admiralty, would suit the purpose. Soon it was realized that the quantities required would exceed the country's capacity to produce silica envelopes and a new product of the M.O. Valve Company, the C.A.T.15, developed for B.B.C. mobile equipment, was taken as a prototype, redesigned with a heavy filament for pulse working and engineered for air cooling and large scale production.

At the same time the work of Boot, Randall and Sayers at the University of Birmingham, evolved a further new type of valve, the cavity magnetron, for use at much higher frequencies. The ideas of these scientists were put into practical design by the late Dr. E. C. S. Megaw, at that time a member of the same company as myself.

It is not perhaps sufficiently widely known that when America entered the war, samples of these first radar valves, together with a colour film of the production processes, were sent from this country to the U.S.A. to enable America to provide radar for her own uses.

There is no time to show this film this evening, nor would it be appropriate to do so,

but I have made it available to the Institution and think it worth showing at some time; apart from its historical interest, it is an example of production engineering which might be of interest to anyone not well acquainted with valve manufacture.

Research is an ever growing tree. New branches are constantly sprouting and due to radar itself the valve industry has added to the number of special valves it has to produce, many of them small in number but substantial in importance. An example is the disc-seal triode shown in Fig. 3.



Fig. 2. An early production magnetron

Concurrently with the magnetron we have seen the development of the klystron, visualized in 1941 as an r.f. amplifier, and now extensively used as a local oscillator for radar receivers. In the Institution, we particularly remember the work of R. Kompfner, both on the klystron and on the travelling wave tube.* The latter is now a substantial factor in microwave relay links without which it would not be possible to enjoy the standard of Eurovision as we know it today.

Television has, of course, played an enormous part in literally focusing attention upon our profession and the industry we serve. I often conjecture as to what fraction of the millions who watch television are in any way aware

* *J.Brit.I.R.E.*, 10, pp. 283-9, August, 1950.

of its scientific development and of the production technique required.

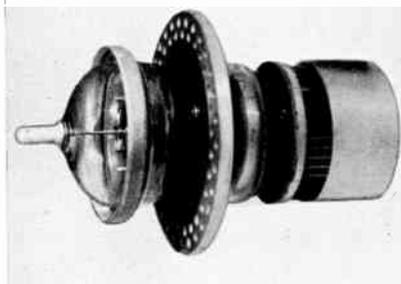


Fig. 3. A u.h.f. disc-seal triode (ACT 25)

To us it is commonplace that a television receiver comprises 400 separate parts, assembled by means of no less than 500 soldered joints, but few of the public realize the production effort required to ensure the satisfactory working of the complete set. On the other hand, of course, we engineers often cannot understand the antics we see performed on the cathode-ray tube!

I have only so far included the cathode-ray tube as one of the parts of a television receiver, but in fact it is itself an excellent example of production technique. The possibilities of the cathode-ray tube were discovered over 70 years ago. The work of Röntgen, Thompson and Braun tells the same story of enthusiastic work in studying new ideas, sharing experiences and not being afraid to experiment.

Figure 4 gives some indication of the early development of the cathode-ray tube before television was ever realized as a commercial possibility. It is interesting to note the similarity in shape of these tubes to present-day devices.

Even before Fleming invented his valve, he originated in 1887 what we now call the "focusing coil," the principles of which continue to be used in the magnetic focusing of cathode rays in television, cathode-ray oscillographs, and electron microscopes.

It was another Professor of London University—Campbell-Swinton—who suggested that the future of television would lie in the development of electronic means for the transmission and reception of the picture. Unfortunately, Campbell-Swinton did not go very far beyond the theoretical stage, and John Logie Baird followed a very different line of thought on the lines pioneered by Paul Nipkow. The very first detailed moving picture was, in fact, transmitted by the Baird system in 1926.

Thus we had a new idea—broadcasting vision as well as sound, and it is to the everlasting credit of Great Britain that our own British Broadcasting Corporation was prepared to be the medium for sharing experiences and not being afraid to experiment. In 1929 the world's first experimental television service was started, and television receivers became available to the public. A great deal of development work was still needed, however, in order to establish a really satisfactory service.

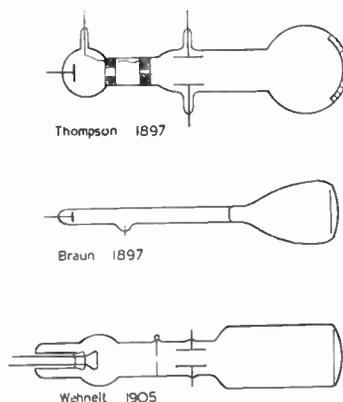


Fig. 4. Early cathode-ray tubes

The Baird system, with its slow mechanical scanning of only 30 lines, was not satisfactory. It soon became obvious that fully electronic television depended upon the further development of the photo-electric cell, and the cathode-ray tube. The photo-electric cell was already well known as a laboratory instrument and used as a controlling device in the transmission of

still pictures by radio before the advent of television.

Concurrently there was a need for much improvement in circuitry. Work in that direction was largely assisted by improvements in valve and tube technique. Overall, the industry was once more in the position of having to encourage the studying of new ideas, sharing experiences and, with the aid of the B.B.C., not being afraid to experiment.

Time does not permit of the tribute that we should all pay to the many engineers and research workers who brought the camera and the television tube to their present stage of development. A great deal of advance has taken place in circuit technique, but we must never forget that coupled with what was largely experimental work, the B.B.C. provided the first public service of its kind in starting high definition television transmission from Alexandra Palace in November 1936.

This new service quickly put a demand on the valve industry, and although it could not be regarded as a very economical run by present day standards, the British valve industry was producing by 1938 ten thousand cathode-ray tubes per annum.

Within two years the industry had to put aside its interest in television as an entertainment factor, but the same equipment, the same tubes and the same valves were playing a vital part in the defence of Great Britain.

I have already shown that improving technique in valve production enabled the products of the radio and electronics engineer to be available to the many instead of to the few. In much the same way, the cathode-ray tube required further engineering development. For example, in the first place the evacuation and sealing of cathode-ray tubes was done on a small scale with rotary pumps, but large scale production evolved the tunnel pump.

The production of cathode-ray tubes has steadily increased, and the factories of this country are now producing well over 2½ million tubes a year. The emphasis must now be on further improved methods of production.

Indeed, the prospects for engineers able to improve production methods, coupled with reliability, are probably as great as, if not greater than, the prospects open to the research or development engineer.

Improved production technique is of paramount importance to industry as a whole, and in our own particular field the increasing use and application of electronics will largely govern our ability to create and meet demands on an economic basis.

For example, cathode-ray tubes are being increasingly required for purposes other than television. In 1947 over 25,000 tubes were bought for industrial and research uses—double the total production of 1938. In 1955, the production, other than for television, had risen to well over 110,000. Obviously the demand will increase with the development of industrial electronics—especially in the field of atomic energy.

These achievements do not, however, warrant complacency in this as in any other industry. History shows that in invention the race between countries is almost always won by a short head only. Great Britain can rightly claim to have been “there or thereabouts” in every race in radio development, and often first in some of the more important races. But the going is getting more difficult and the pace does not slacken. It is vital to this country that we keep our position in the forefront.

Whilst I have, in the main, considered the valve as the essential element in receiving, rectifying, and amplifying the radio signal, the story of the valve as a transmitter is almost equal to the story of the original idea, that is team work in development and production efficiency.

This is, I think, the lesson which should be impressed on every young engineer that, above all, he should take every advantage of learning new ideas and of exchanging opinions with his fellow workers, and that he must understand at least some of the economic implications of bringing those new ideas to practical realization.

It is one of the prime functions of such Institutions as our own to provide the necessary

facilities for the dissemination of post-graduate knowledge. It can only so function if there are members prepared to pool their knowledge, as well as being able and qualified to make use of new information.

In recent years, we have found an increasing number of applications of our art which might generally be referred to as Industrial Electronics. The electronic television camera was one of the most outstanding contributions to an effective television service; it is now being used increasingly in inspection and detection techniques. There is great opportunity for an expansion of such equipment in order to eliminate dangerous work or indeed, to do tasks incapable of human inspection.

The Institution's 1957 Convention will be wholly devoted to the many applications of electronics to various industries and the essential part which electronics does, and will increasingly play, in developing automation. In all those applications the valve is an integral component and the cathode-ray tube will also play a complementary role, not only in industrial processes, but in commercial life.

In the rapid sequence of development, however, we are still far from making the maximum use of what has already been done or instructing others in its usage.

Television on a closed circuit has many applications, but I think that one of the more important in the life of any nation may be the use of the closed circuit for instruction. We know that there is a great shortage of teachers and lecturers, and in addition we have not the accommodation for all our pupils and students.

Some education authorities have experimented with the use of television technique in improving and extending teaching facilities. It took very many years before education authorities took full advantage of the possibilities opened by sound radio in teaching. I think that more could be done in using the application of television for such purposes—in other words, bringing automation into instruction!

I now turn to semiconductors, with specific reference to crystal diodes and transistors.

Reference has been made to the rapidly expanding production of valves, but it should not be forgotten that fully or partially transistorized equipment became available during 1954 and 1955 and this obviously caused a small displacement of valve requirement. I consider, however, that the increasing application and use of the transistor will not materially affect the requirement for valves for many years.

There are many outstanding problems in material construction which affect the economic cost of production of semiconductors; the basic materials of these would seem on first consideration to cost less than the basic materials of a valve. Much work must be done, however, on material production methods to cut costs and eliminate wastage.

Semiconductors also require less fabricating once we are assured of a consistent quality of material. For the present, a high frequency transistor costs more than a valve, but quantity production will reduce this cost when we are assured of a higher yield in mass production.

The immediate demand for semiconductors is to replace valves which at present work under a considerable handicap, such as space limitations, vibratory movement, or difficulty in power supplies. An excellent example, of course, was in the development of centimetric radar or (as I prefer to call it) radiolocation, where thermionic valves were not wholly suitable; this stimulated development of co-axial diode crystals.

Radio engineers are well aware of some of the possibilities of using transistors and crystal diodes. We have, however, to overcome the present temperature and frequency limitations of semi-conductor devices. Improvement of these factors will enable transistors to be more extensively used in all forms of printed and potted circuitry, such as needs to be incorporated in guided missiles. The great problem will then be production, which lends support to the view I have already expressed, that the further development of our art calls for increasing attention, by the engineer, to the problems of large scale quality production.

It seems to me that in current discussion on the work of our universities and other educational institutions, and indeed, in the work of our professional bodies, we tend to become too pre-occupied with the glamour of the words "Research and Development."

Research requires not only a heavy capital outlay, but also men of specialized outlook and inventive minds whose bent for research is usually conditioned by an ability to benefit from experience and what they see. I do not believe that the young graduate is necessarily equipped with the knowledge or outlook to justify him concentrating entirely on research projects.

I also strongly stress the desirability of our examining the whole of our apprenticeship and trainee schemes. Most large companies subscribe to the view that the graduate benefits from works experience, even if his ultimate aim is employment in development and research projects.

I hope that during my Presidency I shall have an opportunity of discussing with our Education Committee the desirability of including in our examination syllabus a paper covering the essential principles of industrial management and production. It is the task of the industrial engineer to turn proved theories into practical products and I visualize that such a paper would require the young engineer to study and understand some of the problems attendant upon converting a research or development project into a marketable commodity.

Such knowledge would also, I think, be invaluable in enabling the engineer to understand management problems and bring closer the understanding between management and labour—a relationship which so often revolves around the engineer who may sometimes have the worst of both worlds!

It would seem that not all people have been convinced either of the importance of electronics engineers, or of the necessity for the Institution itself. I can only comment that the patron of all engineers, Archimedes, claimed that he could move the earth if he had a long enough lever and something to rest it on! Whilst the principles laid down by Archimedes have served

engineers throughout the centuries, there is, unfortunately, no sure mathematical formula for shifting the minds of men!

Though engineers have contributed greatly towards man's comfort in life, there is a danger that they become complacent and lag behind in turning to practical advantage the full benefits of scientific discovery.

There are other considerations which compel us to devote more thought and energy toward improving our production methods. For example, here in Great Britain our population has increased by over 20 million since the beginning of this century and the needs of the population could not possibly have been met by the conditions which prevailed 50 years ago. The deficit has only partly been avoided by the work of the engineer.

I think therefore it is proper that the coming generation, of whom so many will be required in the various branches of engineering, should be reminded of the economic importance, as well as of the dignity of the engineering profession.

It is common knowledge that our profession finds it difficult to recruit enough younger engineers. It is vitally necessary to increase the number of entrants to the profession and in particular young men who can turn to professional advantage the opportunities which exist in our expanding industry.

On the whole I do not think that our industry has been lacking in providing facilities for "sandwich" schemes of training, graduate vacation courses and post-graduate training schemes. It is a matter of conjecture as to whether these schemes are utilized as well as they might be. Indeed, it might be argued as to whether the present conditions of full employment act with as much advantage to the individual as the schemes of apprenticeship which served the engineer so well in the early part of the century.

Whilst the present scheme of introducing technological courses of instruction has much to commend it, it will be very many years before industry feels the full benefit. I suggest, therefore, that young men might reconsider the

advantages of an apprenticeship course.

Notwithstanding our own shortages at the present time, we in Great Britain are very conscious of the compliment paid by the rest of the world to our system of training by the inducements offered to encourage British engineers to emigrate. However nice that compliment, the fact remains that we cannot at present spare engineers for export if we are to have the people at home capable of following up new ideas and developing new techniques.

I should make it clear that we must encourage the interchange of ideas. It is indeed wise to send engineers to other countries to study their methods from which we can learn much, particularly in the engineering of goods for mass production. We reciprocate by the facilities we provide in this country for training and other exchange of information. The fact remains, however, that we must obtain more recruits to our industry before we can afford to send our engineers to permanent jobs overseas.

I think that by focusing attention, as it does, on the need for training young people up to reasonable standards in physics and the various branches of radio and electronic science, this Institution performs a valuable and essential service to the community.

An engineer has been defined as "a man who uses the forces of nature for the welfare of mankind." It is a very simple definition, but it has always been the principal object of all professional engineering bodies and should remain the precept of the individual engineer.

There is, perhaps, a modern tendency to confuse our individual responsibilities. The true engineer is not primarily concerned with economics or politics, though that is not to say that he willingly accepts present conditions as being final. To the contrary, he creates new conditions by providing tunnels, bridges, aeroplanes and atomic energy, all of which bring to the community at large many blessings. To do these things is the engineer's main function, and it is only as a secondary interest that he becomes concerned with the ways in which economics or politics handle his inven-

tions. Certainly history has shown that, as in the Industrial Revolution, economic considerations cannot, and must not, restrict the activities of the engineer.

We should not, therefore, be easily deflected from our main task as engineers—to utilize the gifts of nature for the benefit of man. To do so has been the prime purpose of the engineer ever since man inherited the earth, and we must constantly seek new ways of improving the standards of life and the ways of living. Only in that way can engineers lighten man's labours—a process which has been given many names, the modern version being automation. No one possessed of a logical mind, least of all engineers, can possibly see any logical reason for stopping such evolution. In my opinion, to stop progress would be to invite starvation.

The last 50 years provides countless examples of the benefits of engineering development and production. At the beginning of this century expenditure on motoring was small, and communication in the radio sense was confined to experiment mainly for defence purposes. The radio valve is surely a supreme example of the translation of scientific knowledge and experiment into a commonplace article as a result of the engineer applying his ingenuity to production technique.

It may be doubted whether even Clerk Maxwell could have fully appreciated the ultimate end of the path which was opened up by his enquiring mind, but we have good reason to be proud of such as him. With such example we have even more reason to play our part in fulfilling the declared object of the Institution—"To advance science, especially in the field of radio and electronics, for the benefit of man."

I deeply appreciate the honour of becoming the 14th President of the Institution, more especially because our Institution exists to carry on the tradition of advancing science.

As members of a great engineering profession we should be very proud of the part which we have played in increasing the general standard of living and in providing more opportunities for man to embrace cultural pursuits and broaden his vision.

INSTITUTION NOTICES

Obituary

The Council has learned with regret of the death of the following member and has expressed sympathy with his relatives.

Squadron Leader William Alfred Spacey (Associate) died suddenly from heart failure on October 7th last at the age of 47. Squadron Leader Spacey entered the Royal Air Force in 1928 and was commissioned in 1943. He was a Squadron Commander responsible for instruction in tele-communications equipment maintenance at No. 3 Radio School. He was elected an Associate of the Institution in May 1956.

Students' Essay Competition

The General Council has recently announced the subjects for the Institution's Essay Competition for 1957. The first of these competitions was held in 1956.

Entries are now invited from registered Students of the Institution as well as Graduates under the age of 23 and essays must be submitted by the 30th June. There is a choice of subject which may be:—

- (1) "The Evolution of Radio Communications."
or
- (2) "The Role of Electronics in Industrial Development."

A prize of £10 10s. 0d. will be awarded for the best essay and £5 5s. 0d. for the second best. At Council's discretion, additional prizes may be awarded to essays which are highly commended.

Last year it is possible that the style of many authors was cramped by the limit in length. The Council now asks that essays should be between 3,000 and 5,000 words and should be submitted typed on one side of the paper only.

Visit to Harwell

The Technical Committee is to arrange visits during the Spring and Summer of 1957 to industrial and research establishments in London and the Home Counties. The first visit will be to the Atomic Energy Research Establishment, Harwell, on Wednesday, April 24th; the visit will occupy a full day.

As only a small party can be conducted round Harwell, only Members, Associate Members, Associates and Graduates of the Institution can be included and it is probable that a ballot will have to be held. Members interested should apply immediately to the General Secretary for further details regarding the arrangements.

New Year Honours List

The Council of the Institution has congratulated Wing Commander Albert James Mott on his appointment as an Officer of the Military Division of the Most Excellent Order of the British Empire. Wing Commander Mott is at present Command Radio Officer at Coastal Command Headquarters. He was elected an Associate Member of the Institution in 1949.

Subscriptions from Members Overseas

Members abroad should note that Annual Subscriptions for the year 1957-58 should be received by the Institution on April 1st. The rates of subscription were given on page 477 of the September issue of the *Journal*.

Members resident in India, South Africa or New Zealand may remit either direct to London, or to the Institution's local bankers, specifying that the amount is for the credit of "The British Institution of Radio Engineers," as follows:

New Zealand: Bank of New South Wales, Auckland.

South Africa: National Bank of South Africa, 40 Simmonds St., Johannesburg.

India: State Bank of India, St. Mark's Road, Bangalore.

Separate advice should be sent to the London Office of the Institution immediately payment is made to the local bank.

Graduateship Examination

In 1956, a total of 1,596 candidates entered for the Graduateship Examination. The results of the November 1956 examination, the first to be held under the new regulations, will be published in the February issue of the *Journal*.

Entries from home candidates for the May 1957 examination must be received by 1st April. Overseas candidates should ensure that their entries for the November 1957 examination reach the Institution by 1st May.

Journal

This issue of the *Brit.I.R.E. Journal* is notable in several respects. Firstly, it is the largest single issue since the Institution first published its proceedings thirty-one years ago. The Journal comprises 116 pages, including plates, text and advertising announcements. Secondly, 10,000 copies have been printed for distribution to members and subscribers.

THE CIRCUIT DEVELOPMENT OF A HEAVY-DUTY DYNAMIC BALANCING MACHINE*

by

L. H. Vale (Associate Member)[†]

SUMMARY

The process of dynamic balancing is described in general terms to establish the requirements of the electronic circuitry. The method employed enables direct indication to be given of both the amount and position of existing unbalance. A description is given of a tunable amplifier and its limitations discussed. The operation of the equipment depends, to a large extent, on the performance of a pulse forming circuit. This circuit is required to function at very low frequencies, and a novel method of developing a pulse at the crest of the input cycle, without involving long-time constants, is described.

1. Introduction

Dynamic balancing of rotatable bodies has been standard practice for many years. The increasing use of gas turbines and similar machines has more recently considerably increased the need for equipment capable of dynamically balancing large rotors to very fine limits.

It is not intended in this paper to discuss the mechanical aspects of the subject except insofar as it is necessary to the sense of the text. For more general information on dynamic balancing and its application, papers by Eggleston, Child and Binns^{1,2,3} should be consulted.

2. Types of Balancing

The fundamental difference between static balancing and dynamic balancing may be described in the following manner:

If a rotor, which has a certain radius but a negligible thickness, fixed on a true and balanced shaft, is placed so that the shaft lies across a pair of horizontal knife edges, and is allowed to turn freely, the heavier side of the rotor will assume a vertically downwards position. Addition of the correct amount of mass to the rotor vertically above the shaft (or

subtraction vertically below) will cancel out this unbalance, and the centre of gravity of the rotor will coincide with the centre of the shaft. This is the process of static balancing. Rotating the shaft will not cause a displacement of the centre of gravity and no forces will be generated.

Consider two such disks, fixed some distance apart on the shaft, one of which is balanced and the other unbalanced. Should the combined rotor now be statically balanced by adding or subtracting mass to the previously balanced disk, the centre of gravity of neither disk will coincide with the centre of the shaft. If the shaft is rotated, the rotation will tend to be about the centres of gravity of the individual disks and, if the shaft is constrained by bearings, forces will result from the unbalance, even though the combined rotor is statically balanced. Unbalance in any rigid rotor that has axial length can be resolved into moments in two planes normal to the axis, one on each side of the centre of gravity, and must be suitably corrected in these two planes, individually, in order that the rotor be dynamically balanced.

3. Function of Machine

The function of a dynamic balancing machine is to indicate the angular position and amount of mass required to be added or subtracted in given planes to the body under test in order to correct existing unbalance. Generally

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U.D.C. No. 621.37/8:621-755.

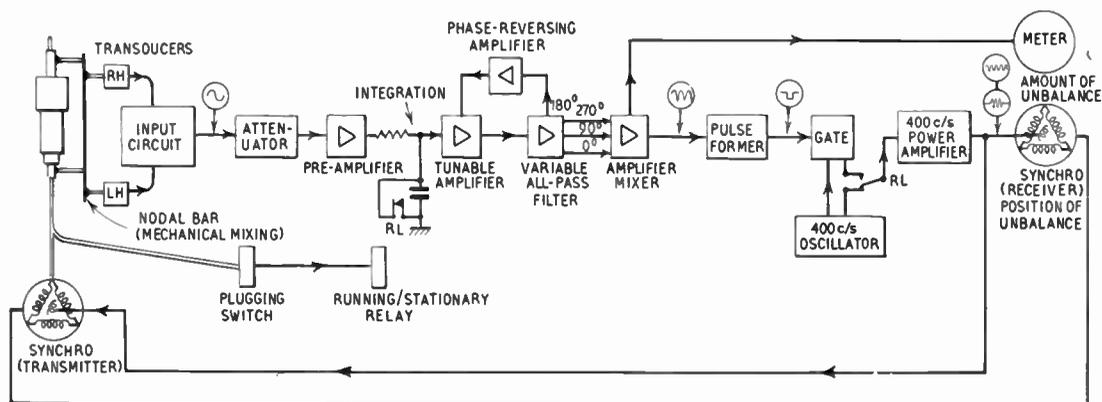


Fig. 1. Block diagram of the electronic circuit of the Balancing Machine.

the correction planes are situated in mechanically convenient positions towards the axial extremities of the rotor. In the simple case of a cylinder rotating on its axis, the correcting planes would normally be the circular end faces.

The rotor to be balanced is placed in bearings—preferably those in which it will finally run—which are fixed to the cradles of the machine. The cradles are suspended in the machine so that they are able to move only in a horizontal direction, normal to the axis of the rotor. If the rotor is perfectly balanced, the cradles will remain stationary when the rotor is turning on its axis. Any unbalance in the rotor, however, will cause the cradles to vibrate sinusoidally. The amplitude of the vibration and its phase angle with respect to the angular position of the unbalance are functions of the unbalance moment, the stiffness of the cradle suspension, the mass of the rotor, the losses in the mechanical system, and the speed of rotation; these are analogous, respectively, to output voltage and phase angle, input voltage, elastance, inductance, resistance, and frequency in an electrical series LCR circuit. The matter is generally more complicated because, while unbalance can be considered to be resolvable into moments in each correction plane, these two moments will not necessarily be equal to each other in either amount or angular position, and yet will be,

to use an electrical term, tightly coupled to each other by the rigidity of the rotor.

It will be seen to be necessary for amplitude of vibration caused by the unbalance forces in each plane to be isolated from the effect of unbalance in the other plane; and that the amount of each unbalance moment and its angular position on the rotor must be accurately indicated.

If we mechanically couple a transducer to each cradle of the machine, movement of a cradle will cause the respective transducer to generate a voltage which can be proportional to either the velocity of movement or displacement of the cradle. In the system to be described, moving coil transducers are used, and the voltage generated is proportional to velocity. The transducers are seismically mounted to reduce the effects of floor vibration.

If the rotational speed is above the natural resonant frequency of the rotor in its cradle assembly, the system behaves analogously to a series LCR circuit operated at a frequency above its resonant frequency. If the speed is sufficiently higher, the stiffness of the suspension ($1/C$), the losses of the system (R), and the small variations in frequency (ΔF) tend to have smaller effect, and the amplitude of vibrations can be considered to be a function of out of balance moment and mass of rotor only. The rotor tends to turn on a new axis through its

centre of gravity and the movement of the cradles is almost 180 deg. out of phase with that of the unbalance. In this machine the period of natural mechanical resonance of the rotor in its cradles is determined by inspection, and the rotational speed is set to above a specified proportion higher. The electronic equipment is required to be capable of handling frequencies corresponding to the specified range of speeds, in this case 4.5 c/s to 25 c/s, corresponding to a speed range of 270 rev/min to 1,500 rev/min.

4. Requirements of Circuitry.

The three general requirements of the electronic circuitry, then, are as follows:

- (1) It must be possible to isolate the effects of unbalance in each plane from the effects of unbalance in the other plane.
- (2) The amount of unbalance in each plane must be measurable.
- (3) It must be possible to indicate on the rotor itself the angular position of the unbalance.

4.1. General Description of Circuit

The block diagram of the electronic circuitry is shown in Fig. 1.

The transducers are connected individually to the input circuit, which provides controls for isolation of the signals from individual planes, separate amplitude controls for each plane, a switch to reverse the polarity of input signals, and an attenuator which allows the overall gain from transducer to amplifier to be adjusted in 20 db steps.

From the input circuit the signals are fed through a pre-amplifier into a tunable amplifier, both of which operate at transducer frequency. It will be appreciated that the limit of sensitivity of the system is the ratio of response to signals at rotational frequency to that at other frequencies. These latter signals, as presented to the measuring portion of the circuit, consist of (a) slow drifting of voltages in the input stages, (b) low frequency signals developed by the transducers due to the rotor swinging at its resonant frequency (this is sometimes greatly amplified by an unfortunate choice of driving

belt length), or (c) higher frequency signals developed by the transducers in responses to noises in bearings, gear box, etc. While every precaution is taken in the mechanical design to limit the last two, it is found impracticable to dispense with some sort of selective circuit and still reach the desired sensitivity.

After the pre-amplifier there is an integration circuit. The purpose of this circuit is to make the velocity proportional signals amplitude proportional, i.e. the response of the system from transducer displacement to the measuring system is independent of frequency over the chosen range, provided that the selective circuit is adjusted to the frequency in use.

From the tunable amplifier the signals are fed through an "Amplifier-Mixer" stage to an a.c. voltmeter, which indicates AMOUNT OF UNBALANCE, and into the pulse forming circuit, which develops a pulse of given angular duration at the peak of each input cycle; these pulses open a gate which allows the output from a 400 c/s oscillator to be fed into a power amplifier and applied to a pair of synchros. The shaft of the transmitter is mechanically connected to the shaft of the rotor under test, the shaft of the receiver actuates the pointer showing POSITION OF UNBALANCE, and the windings are connected stator to stator, rotor to rotor, in the normal synchro link connection.

The general requirement (1) is satisfied in the input circuit, requirement (2) is accomplished by measuring the amplitude of the amplified and integrated transducer signals; (3) is accomplished as follows.

Referring again to the block diagram, Fig. 1, if 400 c/s power is supplied to the rotors of a pair of synchros and their stators are connected in parallel, the shaft of whichever one is not restricted will tend to rotate until it adopts a certain position. If the shaft of the first (transmitter) synchro is turned to a new position, the second (receiver) synchro will tend to turn through a similar angle. If the receiver synchro is one specially designed to have low friction bearings and brushes, and is used, as in this case, to activate only a counterbalanced pointer, it will adopt its correct shaft position

relative to the transmitter synchro shaft to within a degree or better. Therefore, if the transmitter synchro shaft is mechanically coupled to the rotor under test and rotates in fixed angular positions relative to the rotor, then when the rotor is turned the receiver synchro will also turn, with a constant angular relation to the rotor.

Should the power supply to the synchros be applied only when the rotor (and therefore the transmitter synchro shaft) is at a certain angular position, any torque developed in the receiver synchro will tend to return the receiver pointer to a corresponding position and maintain it there. The application of power will occupy some time, however, and during this time the receiver will tend to follow the transmitter, and the receiver pointer will flicker to a degree which is a function of the time that the power is applied during each pulse.

Under the conditions of operation of this machine, mentioned earlier, the movement of the transducers is almost 180 deg. out of phase with the unbalance, i.e. when the angular direction of unbalance in one plane of the rotor is horizontally towards the front of the machine, the corresponding transducer will be at the extremity of its travel towards the rear of the machine (neglecting forces reflected from unbalance in the other plane). Because of the lag in voltage introduced by the integration circuit, this will correspond to the time of peak instantaneous value of the signal applied to the pulse forming circuit is reached.

A pulse is formed symmetrically in time about this peak and power is supplied to the synchros during the pulse period. When the rotor is being driven, therefore, the indicator adopts a position corresponding to the rotor position when the direction of unbalance is horizontally forward in the machine.

It is of interest to observe that, should the pulse formed as above be used to illuminate the rotor by energising a suitable light source in the right position, the direction of the unbalance will be directly illuminated stroboscopically. This method is in general use.

In this machine, a plugging switch on the

drive motor shaft drops out when the drive power is removed and the rotor comes to rest. This disables the pulse forming circuit, and, with 400 c/s power applied continuously to the synchros, the rotor is pulled around slowly until the indicator is again at its previously adopted position. The point from which material must be removed to correct the unbalance is then horizontally forward.

Should it be more convenient to add material to the rotor to attain balance, it is necessary only to reverse the polarity of the voltage from the transducers: the direction of unbalance will then be horizontally towards the rear of the machine.

It was decided, after conducting tests using a rotary switch as a substitute for the transducer and electronic circuitry, that a suitable compromise pulse width (between indicator torque on one hand, and needle flicker on the other) was 30 deg. of transducer frequency, but that pulse width ± 33 per cent. of this were tolerable. In order to smooth out the oscillations of indicator shaft caused by the finite duration of pulse, the indicator synchro is damped with an eddy current disk and permanent magnet.

4.2. Accuracy Requirements

The accuracy of measurement of AMOUNT OF UNBALANCE is ± 5 per cent. Because of the integration the readings are proportional to transducer displacement and so are a function of rotor mass. The normal setting up procedure is to attach a known amount of unbalance at the desired radius on a correction plane on a balanced rotor of the type to be tested (if a balanced rotor is not available sufficient unbalance must be added to render negligible any existing unbalance). With the corresponding variable control in that section of the input circuit, the gain of the system is set so that the AMOUNT OF UNBALANCE meter reads a convenient figure. The steps on the input attenuator are 10:1 and the meter scale is engraved in corresponding colours, so that it will now indicate amount of mass to be added or taken off the rotor at the given radius.

The effects of inaccuracy in indication of angular position of correction point are

illustrated in Fig. 2, which is a vector diagram of forces acting in one correction plane.

- The symbols used in Fig. 2 are: —
- F_u = Unbalance force existing on rotor.
- φ = Angle of error in application of correction force.
- F_{R1} = Minimum resultant force when correction force F_c is applied at an error angle of φ .
- F_{R2} = Resultant force when magnitude of correction force $F_{c2} = F_u$, at an error angle of φ .
- F_{R3} = Resultant force when correction force $F_{c3} > F_u$ at an error angle of φ .

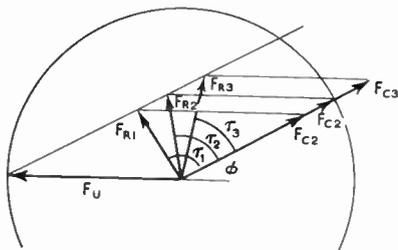


Fig. 2. Vector diagram showing the forces acting in one correction plane.

If a correction force F_c is applied with an error angle φ , by the cosine law,

Resultant Force,

$$F_R = \sqrt{F_u^2 + F_c^2 - 2 F_u F_c \cos \varphi} \dots\dots(1)$$

Having determined the magnitude of F_R , the angle τ between F_R and F_c can be determined by the sine law: —

$$\tau = \sin^{-1} \left(\frac{F_u}{F_R} \cdot \sin \varphi \right) \dots\dots(2)$$

τ is acute if $F_c > F_u \cos \varphi$
 and obtuse if $F_c < F_u \cos \varphi$.

From Fig. 2 it will be seen that the magnitude of the resultant reaches a minimum when its direction is normal to the correction force, as illustrated by F_{R1} .

Therefore the condition of minimum resultant is reached when

$$F_{c1} = F_u \cos \varphi \dots\dots(3)$$

and under this condition,

$$F_{R1} = F_u \sin \varphi \dots\dots(4)$$

$$\text{and } \tau_1 = 90^\circ \dots\dots(5)$$

$$\text{If } F_c = F_u = F_{c2}$$

From eq. (1),

$$F_{R2} = 2 F_u \sin \frac{1}{2} \varphi \dots\dots(6)$$

$$\text{and } \tau_2 = 90^\circ - \frac{1}{2} \varphi \dots\dots(7)$$

It can be determined from eq. (6) that if the exact amount of correction is made but with angular error of 2.8 deg. the resultant out-of-balance force will be 5 per cent. of the original out-of-balance force; however, if we allow the amount of unbalance remaining after correction to be as large as 10 per cent. of the original amount of unbalance, and, owing to tolerances in measurement, the amount of correction made is 5 per cent. greater than the correct amount, from eq. (1) it can be determined that the permissible angular error is 4.8 deg. If the correction is 5 per cent. less than the correct value, the permissible angular error is 5.1 deg.

The effect of this total tolerance of 10 per cent. is that, under the extremes of inaccuracy, the amount of unbalance can be reduced by a factor of 10 after each measurement, which is in accord with normal practice.

The sensitivity of the circuitry is to be such that accurate indication of amount and angular position of unbalance can be obtained when the amplitude of vibration of the rotor due to unbalance is 0.00002 in.

To summarize the requirements of the circuitry, these are: —

- (a) To indicate amount of unbalance, ± 5 per cent., and
- (b) Angular position of unbalance, $\pm 4.8^\circ$.
- (c) These indications are to be obtained with inputs as low as those corresponding to transducer movements of 20×10^{-6} in. amplitude.
- (d) The frequency range of the equipment is 4.5 c/s to 25 c/s.

In addition to these requirements, it was considered necessary to keep the circuitry simple, to use standard components as far as possible, and to reduce the number of valves, i.e. glass envelopes, to a minimum.

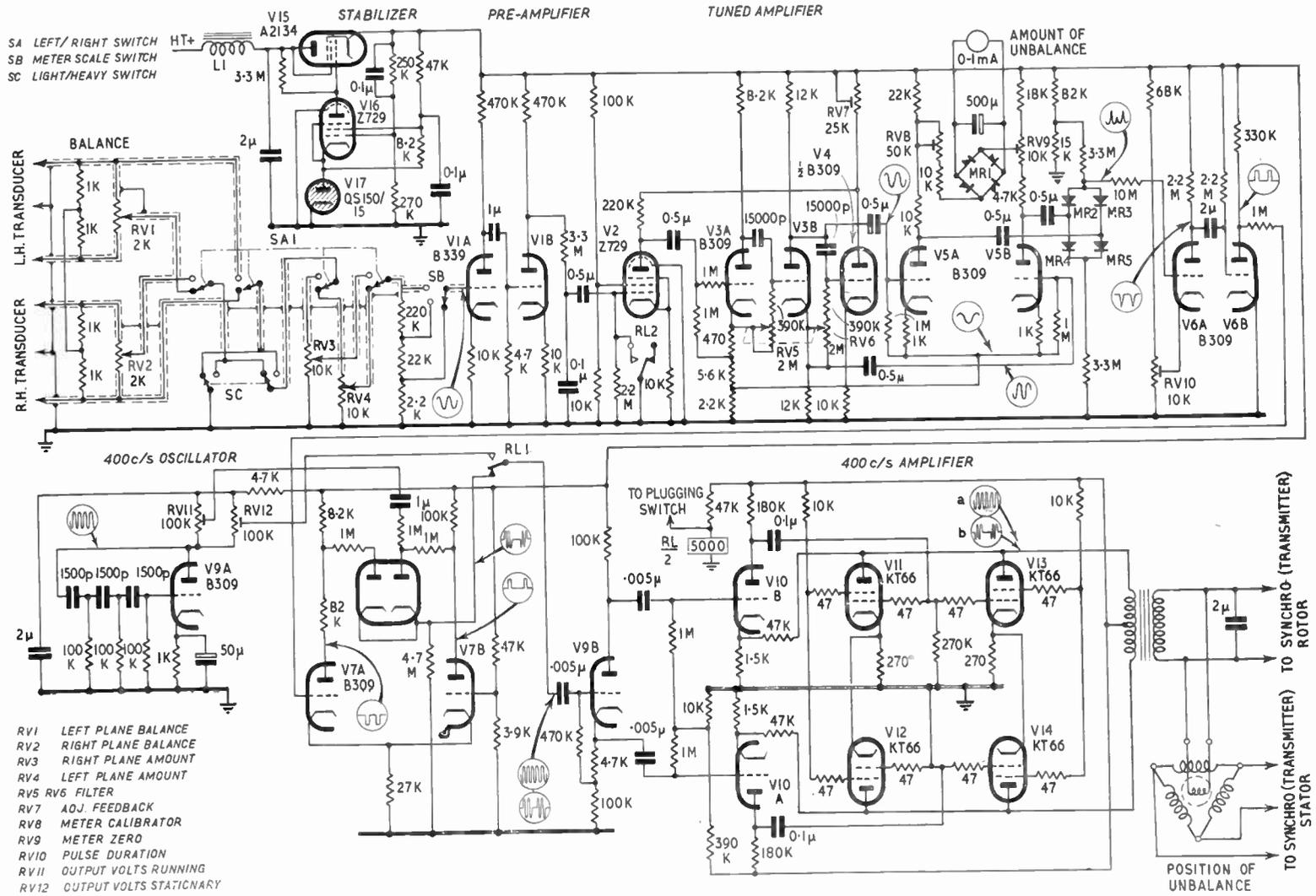


Fig. 3. Circuit diagram of the Balancing Machine.

5. Detailed Description of the Circuit.

The development of the individual sections of the circuit will now be described in more detail. Fig. 3 shows the full circuit.

5.1. The Input Circuit

The requirements of this section have been mentioned in the discussion of the block diagram, and these are met with the circuit shown in Fig. 4. In this circuit, the heavy lined portion of the diagram contains that portion in use when the PLANE switch, SA, is turned to RIGHT-HAND PLANE.

It will be seen that the signal appearing between the arms SA1 and SA2 is the sum of the voltage across R4 and the voltage existing between the R1, R2 junction and the arm of RV1. R1, R2, and the two sections of RV1 form a resistive bridge across the terminals of the l.h. transducer, so that any voltage generated by the l.h. transducer which is in phase with that generated by the r.h. transducer, whether positively or negatively, can, by adjusting RV1 to a suitable value, be subtracted from the voltage across R4, which is half the output from the r.h. transducer.

The voltage generated by the r.h. transducer can be the result of, firstly, the vibration caused by unbalance in the r.h. plane itself, and secondly, the vibration caused by the unbalance in the l.h. plane, which is coupled mechanically, through the rotor swinging about a node, to the r.h. plane. This latter vibration is essentially of the same phase angle at the two correction planes, but, depending on the configuration of the rotor, may be of the same or of opposite polarity. Therefore a sufficient portion of the l.h. transducer voltage may be added or subtracted to the r.h. transducer voltage to balance out the mechanical coupling.

The BALANCE controls RV1 and RV2 are set up by adding artificial unbalance to one plane only of a rotor. By suitably adjusting the opposite BALANCE control it is possible to read the same amount and position of unbalance in that plane whether the artificial unbalance is present or not.

With some configurations of rotor it is not possible, with the limited amount of bucking

voltage obtainable with this circuitry, to obtain complete isolation between the planes. There is, however, a mechanical method of coupling between the cradles and transducers which allows a much wider range of adjustment. The BALANCE controls are, however, always used as a final means of accurate adjustment.

The LIGHT/HEAVY switch, SC, enables the voltage between the arm of SA3 and earth to be reversed in polarity, according to the need to add or subtract mass from the rotor in the correction process.

The AMOUNT controls, RV3 and RV4, afford separate, continuously variable, adjustment of system gain in each setting of the PLANE switch. With SA in the r.h. position as shown in Fig. 3, RV3 is in circuit. These controls are used, as described earlier, to calibrate the AMOUNT meter in terms of actual mass to be applied to the particular rotor being balanced.

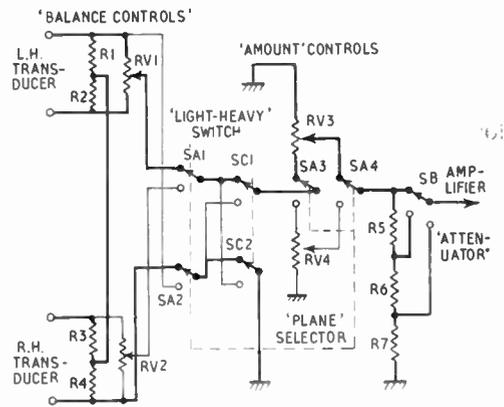


Fig. 4. The input circuit.

SB is a simple voltage divider attenuator. The resistors are such that each step is in decimal relation to the adjacent ones. The scale of SB is labelled COARSE, MEDIUM, and FINE and the colouring of each word corresponds to the colour of the appropriate scale on the AMOUNT OF UNBALANCE meter. Attenuation of the signals before amplification restricts the range of signal amplitudes passed through the amplifier during operation to 10:1 and so tends to facilitate the attainment of constant gain.

The BALANCE and AMOUNT controls, the LIGHT/HEAVY switch, and the FILTER tuning control are intended to be set up by a suitably competent person prior to the balancing operation, and are situated under a cover plate. The machine operator need use only the PLANE selector switch and attenuator switch while rotors of any one type are being balanced.

5.1.1. Gain requirements

The signals from the attenuator are fed to the amplifier sections. The signal appearing at the input of the pulse forming valve V6A (refer to the main circuit diagram, Fig. 3) is required to be at least 2.5 volts peak value in order to form a well defined pulse. It will be shown later that this corresponds to an output from the mixer amplifier of 7 volts r.m.s. The output of the transducers when measured at 4.5 c/s is 4.4 microvolts r.m.s. per microinch of armature travel. So that the output with 20 microinches of travel at this, the lowest frequency to be used, is 88 microvolts. Because of the resistive bridge arrangement in the input section, the transducer signal is halved, so that the smallest signal into the amplifier is 44 microvolts. The integration circuit introduces a loss of 9.7 times at 4.5 c/s so that the total gain between the input of the pre-amplifier and the output of the mixer amplifier is required to be at least

$$\frac{7 \times 10^6 \times 9.7}{44} = 1.54 \times 10^6.$$

The resistance-capacitance couplings between stages, being proportioned to introduce a very small amount of phase shift, will have a negligible effect on gain, so that the figure for the total gain can be considered as equal to actual valve gain.

The gain of the mixer amplifier and tunable amplifier sections is limited by other considerations. It is necessary, therefore, to design the pre-amplifier so as to furnish the additional gain required.

The gains of these sections are as follows:—

Mixer amplifier 11.5,

Tunable amplifier (including phase splitters) 180,

therefore the required gain in the pre-amplifier section is 785.

5.2. Pre-amplifier

In addition to obtaining the required gain it was decided that the pre-amplifier circuit should (a) use some negative feedback to stabilize the gain, and that (b) only one glass envelope should be used if possible.

The gain is somewhat higher than can be obtained in a single stage, except perhaps a pentode operating under "starvation" conditions.⁴ It has been reported that the gain of a pentode under such conditions is not particularly stable⁵, and, in any case, the amount of gain available, even in the most favourable circumstances, would not be sufficient to allow the application of negative feedback.

The cascode circuit with additional anode resistor for the first triode⁵ appeared to be worth investigating. This was found, however, due to the direct coupling between stages, to be rather critical of d.c. values. A re-adjustment of bias on the second triode was required each time the valve was changed.

The circuit finally adopted has two conventional resistance capacitance coupled stages. The cathode resistors are not bypassed and provide 10 db negative feedback in each stage.

The gain of each stage can be found from the formula

$$A_1 = \frac{\mu R_L}{(\mu + 1) R_K + r_a + R_L}$$

With the components values used, and using information supplied by the valve manufacturers, the gain of each stage is slightly over 30.

The total gain of the pre-amplifier is therefore ample, and stability is good. The effective input capacitance of the high μ triodes will have a negligible effect over the range of signal frequencies handled.

5.3. The Integration Network

The integration network, R and C , will introduce a voltage loss of $\sec \theta_1$, where $\theta_1 = \tan^{-1} CR$, the phase angle (lagging) of the circuit. With the values used, 3.3 M Ω and 0.1 μ F, and allowing for an additional 0.14 M Ω source impedance from V1B and the anode load

resistor, the voltage loss is 9.7 times, and the phase angle 84.1° , lagging, at 4.5 c/s. Considering the phase angle as compared to an ideal integrator phase angle of 90° lagging, it is 5.9° leading. The difference reduces as the frequency increases, and so is cumulative with the phase errors due to the intervalve coupling circuits. The voltage across the capacitor decreases proportionally with increasing frequency, and so the circuit compensates for the increasing output with frequency characteristics of the velocity proportional transducers.

Any 50 c/s interference from the grid of VIA and the transducer wiring will be attenuated 20.9 db more than a 4.5 c/s signal, and as the transducer voltages are proportionally greater at higher frequencies, the effect of this interference can be neglected in practice.

5.4. The Tunable Amplifier

The tunable amplifier is of the type using an all-pass filter and positive feedback. It uses no specially manufactured components, and if the required selectivity is not great, is sufficiently stable in gain. It has been in use for some years and has more recently been described in the literature.⁶

A block diagram is shown in Fig. 5 and a circuit diagram in Fig. 6.

It will be seen in Fig. 5 that input voltage is fed through an input amplifier, an all-pass filter, and a phase-reversing amplifier. The output of the phase-reversing amplifier is fed back to the input of the all-pass filter.

The characteristic of a perfect all-pass filter is that it does not vary attenuation with varying frequency but only affects phase. In this circuit the former quality is very closely approached, and the variation of phase angle, or rather the frequency at which a given phase rotation is accomplished, is variable.

The phase angle through the phase-reversing amplifier is 180° , so therefore if the input is of a frequency so as to be also reversed in phase by the all-pass filter, the feedback will be most positive at this frequency. The signal is reversed in phase at only one frequency in the low-frequency range with the components used,

and therefore the positive feedback and gain are greater at that frequency. The circuit then allows adjustment of the component values in the all-pass filter so that the frequency at which phase reversal occurs is variable between the required limits, thus making possible a greater gain at any chosen frequency within the range.

It was felt that a positive feedback circuit might not be desirable in equipment required to have stable gain characteristics. However the effect of the circuit is to present an anode load to the input amplifier that is greatest at the desired frequency; also the input amplifier itself is stabilized by cathode feedback, so that the gain tends to become closely proportional to anode loading; further the gain of the other valves in the circuit is highly stabilized. Finally a stabilized h.t. supply is used. The objections to a positive feedback circuit are therefore minimized, and in addition it was found that other requirements, notably ease of adjustment, limited the selectivity obtainable, rather than gain stability. This subject is examined more fully later.

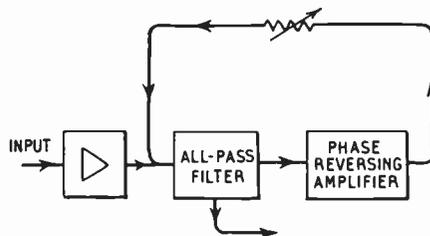


Fig. 5. Block diagram of the tunable amplifier.

Considering the circuit as above, the logical place from which to take the output voltage would seem to be the output of the input amplifier. However, it is possible to take a low impedance output from the first cathode in the all-pass filter. Later, when it was found necessary to obtain output voltages in quadrature for the operation of a pulse forming circuit, these were also available from the all-pass filter circuit.

Referring to Fig. 6, the operation of the circuit is as follows:

VI is the input amplifier. A pentode is used under semi "starvation" conditions with an

unbypassed cathode resistor. V2 and V3 are two halves of a high mutual-conductance double-triode, and V4 is half of another similar valve.

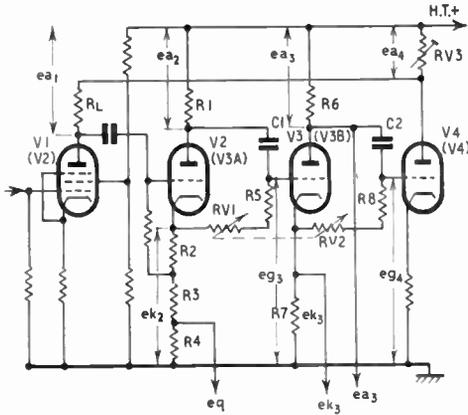


Fig. 6. Circuit of the tunable amplifier.

The output was originally taken from the cathode of V2, but when three voltages in quadrature were later required, the output from V2 (eg), was tapped down the cathode resistor to furnish a voltage in a definite amplitude ratio to ek_3 and ea_3 , the outputs from V3. This relationship is discussed in Section 5.5.

V2 and V3 operate in an identical manner as phase splitters.

$$R1 = R2 + R3 + R4, \text{ and } R6 = R7;$$

thus $ea_2 = -ek_2$, and $ea_3 = ek_3$.

The two resistance capacitance networks are identical in so far as: $C1 = C2$, $R5 = R8$, and $RV1 = RV2$, as they are ganged.

Figure 7 shows the vector relations in the V2 circuit. ea_1 , being the grid voltage of V2, is in phase with ek_2 , but ek_2 is reduced in amplitude. The gain from grid to anode

$$A' = \frac{\mu R_1}{(\mu + 2) R_1 + r_a}$$

With the valve and components used ($R_1 = 8.2 \text{ k}\Omega$, $r_a = 10 \text{ k}\Omega$, $\mu = 65$) this is 0.945. It will be seen that the angle by which eR leads ek_2 , $\gamma = \tan^{-1}(1/\omega CR)$ and that the angle by which eg_3 leads ek_2 is 2γ , so that the

angle of lead of the V2 circuit is $2 \tan^{-1}(1/\omega CR)$ and that when $\omega CR = 1$, $2\gamma = 90^\circ$.

V3, being a similar circuit, will have also a leading phase angle of 90° , when $\omega CR = 1$, and a similar gain, so that the total gain of two stages is approximately 0.9, and phase angle 180° .

At other frequencies, when $\omega CR \neq 1$, the gain is still about 0.9, but the phase angle will be $4 \tan^{-1}(1/\omega CR)$.

The output from the C2, R8, RV2 network is reversed in phase by V4. The anode load of V4 is variable, to provide a convenient means of varying the loop gain, AB , of the three stages.

It will be appreciated that at the frequency when $\omega CR = 1$, which we shall call f_0 , that ea_4 will be in phase with ea_1 , and of amplitude $AB.ea_1$ so that the effective impedance across ea_1 will be $1/(1 - AB)$ times its initial value, and resistive.

At infinitely low or high frequencies the effective impedance will be $1/(1 + AB)$ times the initial value and also resistive.

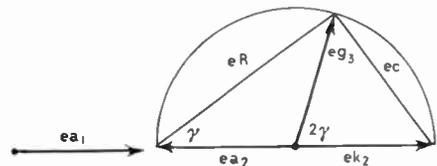


Fig. 7. Vector relationships for the circuit of V2.

At other frequencies the effective impedance across ea_1 and the phase angle can be shown best by a vector diagram of the voltages ea_1 , ea_4 and eR_L , which is shown as Fig. 8. The angle $\eta = 180^\circ + 4 \tan^{-1}(1/\omega CR)$.

By definition, the effective impedance across ea_1

$$Z_{\text{eff}} = R_L \cdot \frac{ea_1}{eR_L}$$

The operating conditions of the pentode V1 are such that the anode impedance is much higher than R_L . The grid resistor of V2 is returned to a tap on the cathode load and so its effective resistance is increased in the ratio:

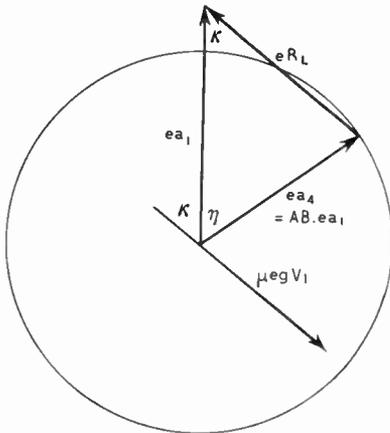


Fig. 8. Vector diagram for V2 circuit of general application.

$$\frac{1}{1 - A'(R3 + R4)(R2 + R3 + R4)}$$

It is a reasonable approximation for simplicity's sake to assume that both V1 anode impedance and V2 input resistance can be considered infinite. Then the gain of V1 will be proportional to its anode load, i.e. Z_{eff} .

Thus $A_1 \propto R_L \cdot \frac{ea_1}{eR_L}$

If we reduce AB to zero, by turning $RV3$ to zero resistance, then

$$ea_4 = 0 \text{ and } eR_L = ea_1,$$

Hence $Z_{eff} = R_L$ at all frequencies.

The component values are such that under these conditions the gain of V1, which we shall term A_N , is 20.

The gain of V1, when $RV3$ is adjusted to give a loop gain of AB (which must be always less than unity, otherwise Z_{eff} becomes a negative resistance at f_0 and the circuit oscillates) can be found by applying the cosine formula to the triangle in Fig. 8.

$$e_{RL} = \sqrt{\{ea_1^2 + ea_4^2 - 2ea_1 \cdot ea_4 \cdot \cos \eta\}}$$

which, since $ea_4 = AB \cdot ea_1$,

$$= \sqrt{\{ea_1^2 + (AB \cdot ea_1)^2 - 2AB \cdot ea_1^2 \cdot \cos \eta\}}$$

but $\frac{A_1}{A_N} = \frac{ea_1}{eR_L} = \sqrt{\frac{1}{1 + AB^2 - 2AB \cdot \cos \eta}}$

In Figure 8 the vector of μeg_{V1} is drawn from the apex of the angle η . If ra_{V1} and rg_{V2} are infinite this vector will be parallel to the vector of e_{RL} ; and phase angle between μeg_{V1} and ea_1 will be equal to the angle z .

Using the sine formula on the triangle of Fig. 8,

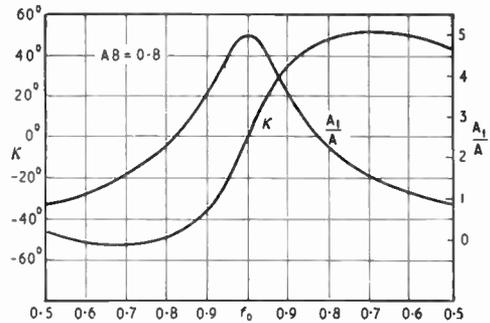
$$\frac{AB \cdot ea_1}{\sin z} = \frac{eR_L}{\sin \eta}$$

but $eR_L = ea_1 \cdot \frac{A_N}{A_1}$

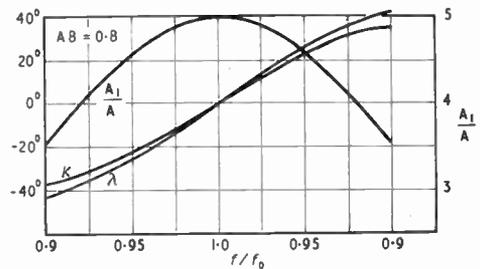
Hence $z = \sin^{-1}(AB \cdot \frac{A_1}{A_N} \sin \eta)$

Figures 9 (a) and (b) are graphs of these two functions against frequency relative to f_0 for $AB = 0.8$.

It will be noticed in Fig. 9 (b) that the phase angle z varies greatly for a comparatively slight change in gain near the f_0 point. This allows



(a)



(b)

Fig. 9. Graphs of the functions for the amplifier gain and phase angle.

the phase angle to be adjusted by the tuning potentiometers without greatly decreasing the filtering ability of the circuit. It also indicates that, with $AB=0.8$, a 1 per cent. change in frequency, i.e. in rotational speed, will cause a phase error of approximately 5 deg.

The method of adjusting the amplifier is to add unbalance to a rotor sufficient to override any unbalance that may be existing. The FILTER control is then adjusted so that the POSITION OF UNBALANCE indicator is free from error. Under these conditions the whole system has no error in angular indication, provided that the rotational speed remains unchanged.

Figure 8 indicates that maximum value of α the phase angle, occurs when eR_L vector is tangential to the locus of $AB.ea_1$ with varying values of η . With these values of η and AB , eR_L vector is perpendicular to $AB.ea_1$ and $\eta = \cos^{-1} AB$.

$$\text{Therefore } \alpha = \sin^{-1} AB$$

$$\text{and } d\alpha = \frac{d(AB)}{\sqrt{\{1 - (AB)^2\}}}$$

Graphically it may be shown that, as AB increases in value, the knee of the α curve (Fig. 9 (a)) approaches the f_0 ordinate and increases in value. Therefore, the change in phase angle with variations of frequency around the f_0 value increases sharply when AB increases. While some phase change with frequency is useful for compensation, as mentioned earlier, an oversteep slope makes tuning of the filter very difficult and the error due to frequency changes, i.e. rotational speed, can become too large to be tolerated.

This being a positive feedback circuit, variations in AB will cause correspondingly greater variation in overall gain in the proportion

$$\frac{d(A_1)}{d(AB)} = \frac{A_N}{(1 - AB)^2} \text{ at } f_0$$

It will be seen, therefore, that the factor AB should not be too close to unity, if the gain is to be stable, and therefore the filtering effect of the circuit is limited by such stability considerations.

Variations in AB will be caused by variations in supply voltages, valve constants, or resistor values. The supply voltage is from a 250V regulated line, and in any case variation in voltage has a very small effect on gain.

The valve constant that will most probably change is the anode impedance. Referring to the formula for phase splitter gain given above, it will be seen that for high mutual conductance values, variations in anode impedance will have only second order effects.

Variations in resistor values can be quite serious. If both cathode and anode resistor of one stage change together the effect can be neglected, but if one changes with respect to the other a change of gain occurs equal to half the change in relative resistance. Given a change of 1 per cent. in one resistor only when $AB=0.8$, the resultant change in A_1 could be ~ 2.5 per cent. With similar change but with $AB=0.9$, the change in A_1 would be 5 per cent. It is indicated, therefore, that the anode and cathode resistors should be high stability types and also, again, that AB should not be too high. A figure of AB of 0.8 was chosen as being a suitable compromise between filtering and these two factors.

The method of adjustment is to turn RV3 to zero ohms and check the gain; turn RV3 to some other value and adjust the filter control (RV1, RV2), to give maximum gain; then adjust RV3 so that the circuit gain is $1/(1 - AB)$ times the gain when RV3 is zero.

The total phase angle of the circuit is slightly greater than α , the phase angle due to the load on V1. It will be noticed that, for reasons given later, three output voltages are taken from the circuit. This is necessary with the final pulse forming circuit used. Two of the voltages, ea_3 and ez_3 are of opposite phase and constitute the main input to the pulse circuit. eq is approximately a quarter of ez_2 and is in quadrature with ea_3 and ek_3 . As these voltages are taken from the circuit after the first RC filter section, an additional phase rotation is added to α , equal to 2γ , namely $2 \tan^{-1}(1/\omega CR)$. This is quite small around $f_0=f$ but is added to α in Fig. 9 (b) to show the complete phase angle λ .

5.5. Pulse Forming Circuits

Three pulse forming circuits (shown in Figs. 10, 12, and the left-hand portion of Fig. 16) were investigated.

As stated in Section 4.2, it is required to form a pulse symmetrically, in time, about the peak of the signal voltage. The first circuit (Fig. 10), as is usually the case, was simplest, but was not practicable because of the long time-constants necessarily employed in an attempt to obtain accuracy in phase. It is possible to obtain greater accuracy with the second circuit, which is only a little more complicated, but the time-constant, although much reduced, remains a problem.

The third circuit performs very satisfactorily in practice. It does not generate a d.c. voltage from the signal, and so the question of time-constants is not introduced. Both the second and third circuits are considered to be novel.

The first two circuits did not require the use of a mixer-amplifier as does the third, and as the pulse is formed on the crest of the input cycle, the input signals to the first two could be in phase with transducer displacement and taken from the outputs of V2 (Fig. 6) through a suitable amplifier. The additional phase angle mentioned above— 2γ —was not introduced and α was the total phase shift of the tunable amplifier.

5.5.1. Circuit of Figure 10

In the circuit shown in Fig. 10 the input is applied to the grid of V1.

If there is no a.c. input applied to the grid of V1, the cathode voltage rises to a value sufficiently high to almost cut off the anode current of V1 through R1. If a.c. is applied to the grid, the cathode voltage will rise until the grid only becomes sufficiently less negative with regard to the cathode to allow the valve to pass anode current on positive peaks of the input cycle for a certain time—during which the pulse is formed across R1. With a constant value of a.c. input, the voltage across C1 maintains a certain d.c. level; if the time constants of charging and discharging circuits are sufficiently long (as they must be in practice) to enable us to consider that the charge and

discharge currents are constant while flowing, the proportion of the time of anode conduction to the time of the remainder of one input cycle is equal to the ratio of the average charging current to C1 (during the conduction period) to the average discharge current, through R2.

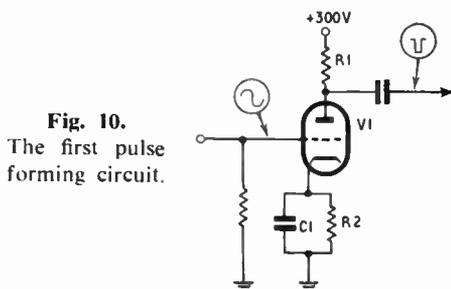


Fig. 10. The first pulse forming circuit.

Referring to Fig. 11, as the charging current will cause the cathode voltage to rise during the pulse period, it follows that the time elapsing between the peak of the a.c. cycle and the cessation of charging current will be smaller than half the total duration of the pulse period. Therefore the pulse will not be symmetrically disposed about the peak of the a.c. cycle.

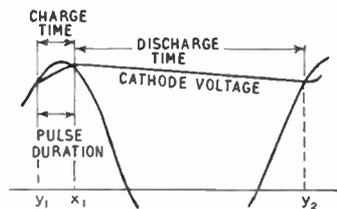


Fig. 11. Charge and discharge characteristics of the circuit of Fig. 10.

If the desired pulse width is 2α radians, and the tolerable error is β radians, then, because of the comparatively long time-constants involved, we can simplify calculations by making the approximations that the charging and discharging rates are linear, and, in this case, if $\alpha \gg \beta$

then cathode voltage at time $x = k \cos(\alpha - \beta)$ and cathode voltage at times $y_1, y_2 = k \cos(\alpha + \beta)$

Therefore the rate of discharge is

$$\left\{ 1 - \frac{\cos(\alpha + \beta)}{\cos(\alpha - \beta)} \right\} \text{ of original charge in}$$

$(2\pi - 2\alpha)\omega$ sec, if the angular velocity of the

signal is ω rad/sec. Hence the time-constant, T_1 , of the discharge circuit can be derived from the expression

$$T_1 = \frac{2\pi - 2\alpha}{\omega \left(1 - \frac{\cos(\alpha - \beta)}{\cos(\alpha + \beta)} \right)}$$

and the time-constant for the charge circuit

$$= \frac{2}{\omega} \cdot T_1$$

For a 30 deg. wide pulse, an input frequency of 4.5 c/s, and an allowable error of 0.25 deg., the necessary time-constant of R2,C1 would be 92.5 seconds.

This would mean that the equipment would not generate pulses for about five minutes after the attenuator switch had reduced the input by a 20 db step, and also, that the cathode capacitor, C1, would be impractically large.

5.5.2. Circuit of Figure 12

This circuit employs both sections of a double triode—one to charge the cathode capacitor to a suitable value, and the other to develop the pulse in its anode circuit.

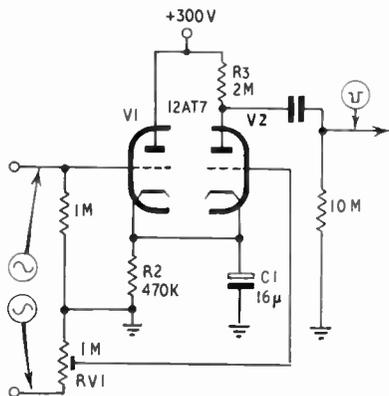


Fig. 12. The second pulse forming circuit.

One advantage of this circuit over that of Fig. 10 is that, by separating the functions of charging the cathode capacitor and developing the pulse, the time constant of the components in the cathode circuit can be reduced to a more practical value.

The major advantage is one of accuracy.

Should the amount of charge to C1, during the V2 conduction period, equal the amount of discharge through R2 during the same period, the cathode voltage will be the same at the end of the pulse period as at the commencement, and the pulse will be symmetrically disposed in time on the crest of the input sinusoid.

There is a practical advantage also; RV1 now offers a convenient method of adjusting pulse duration.

The limiting factor on the reduction of the C1, R2 time constant is the allowable variation of pulse duration.

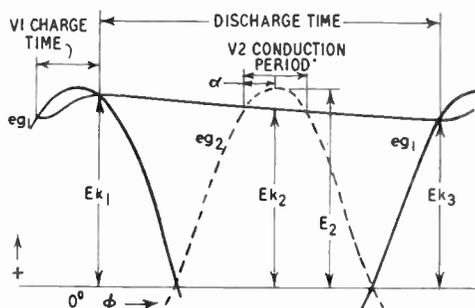


Fig. 13. Charge and discharge characteristics of the circuit of Fig. 12.

It will be seen in Fig. 13 that the rate of decay of cathode voltage E_k , from its maximum value E_{k1} at the end of the charge time, to its value E_{k2} midway through the pulse period, is a function of the time constant of C1,R2, whereas the actual time between the positions of these two values is a function of input frequency. E_{k2} will be greater for higher input frequencies. It is necessary, therefore, to make the time constant of C1,R2 large enough to permit this variation to be tolerable.

Because of the smallness of the charging time constant through V1, E_{k1} can be assumed to occur very closely to $t=0$, and the approximations may be made that the time between E_{k1} and E_{k2} values will equal half the time of one input cycle less half a pulse width, and that the value of E_{k1} is independent of input frequency.

For a pulse of angular duration 2α and considering that E_{k2} maintains a constant value

during the pulse period,

$$E_{k2} = E_2 \cdot \cos \alpha$$

where E_2 is the peak value of the alternating voltage on V2 grid.

If we permit an increase of 2ζ in the pulse width, when the input frequency is reduced over the entire range, then

$$E_{k2} = E_2 \cdot \cos (\alpha + \zeta)$$

at the low-frequency end of the range.

If f_1 = highest input frequency,
and f_2 = lowest input frequency,

then the discharge rate of C1, R2 should be less than

$$\left\{ 1 - \frac{\cos (\alpha + \zeta)}{\cos \alpha} \right\} \text{ of original } \left\{ \frac{1/f_2 - 1/f_1}{2(1 - \alpha/\pi)} \right\} \text{ sec.}$$

Therefore the minimum allowable time constant of C1,R2 is

$$\frac{1/f_2 - 1/f_1}{2 \left(1 - \frac{\alpha}{\pi} \right) \left(1 - \frac{\cos (\alpha + \zeta)}{\cos \alpha} \right)}$$

For a pulse width of 30 deg. (2α) at 25 c/s (f_1), and permitting an increase of 6 deg. (2ζ) at 4.5 c/s (f_2), we can determine from the above formula that the minimum time constant is 6.7 seconds.

The formula for determining the value of the V2 anode resistor ($R_L V_2$), so that amounts of charge and discharge from C2 are equal during the pulse period, is developed as follows:—

The amount of discharge of C1 through R2 during the pulse period, under the same conditions as above, and considering that E_k remains constant throughout the pulse period, is

$$K.E_2 \cdot \frac{2 \sin \alpha \cdot \cos \alpha}{R_2} \text{ coulombs,}$$

and the amount of charge to C1 through V2 during the same period

$$\frac{K.E_2 \cdot \mu V_2 \int_{\frac{1}{2}\pi - \alpha}^{\frac{1}{2}\pi + \alpha} \sin \varphi \cdot d\varphi - 2 \sin \alpha \cdot \cos \alpha}{R_L V_2} \text{ coulombs.}$$

Therefore, to make these amounts of charge and discharge equal,

$$R_L V_2 = \frac{\mu V_2 \cdot R_2 \left[-\cos \varphi \right]_{\frac{1}{2}\pi - \alpha}^{\frac{1}{2}\pi + \alpha} - 2 \sin \alpha \cdot \cos \alpha}{2 \sin \alpha - \cos \alpha}$$

The actual time constant of R2, C1, in Fig. 13 is approximately 8 seconds,

$$R_2 = 0.47 \text{ M}\Omega$$

$$\mu V_2 = 60 \text{ and}$$

$$\alpha = 0.26 \text{ radians}$$

applying these values to the above formula, $R_3 \cong 1.7 \text{ M}\Omega$.

An anode load of 2 MΩ and succeeding grid resistor of 10 MΩ were used.

Although the time constant of the cathode circuit components was much smaller than that of the corresponding components in the circuit of Fig. 9 (b), it was still long enough to disable the equipment for a considerable period when the input was reduced by the attenuator in a 20 db step.

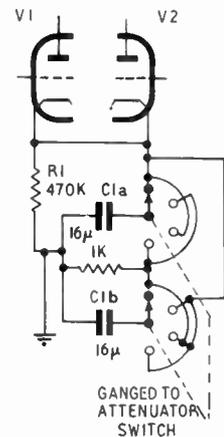


Fig. 14.
Special switching circuit for attenuator used with circuit of Fig. 12.

In order to obviate this effect, a switch was mechanically ganged to the attenuator switch and alternate cathode capacitors are provided (see Fig. 14). The switch was so connected that when one capacitor is connected across the cathode resistor, the other capacitor is discharged through a low value resistor, so that, whenever the attenuator switch is moved to a new position, a discharged capacitor is connected across the cathode resistor.

The pulse forming circuit of Fig. 12 was used in the machine in practical tests. It was found to suffer from the following practical faults:—

(1) Higher than signal frequency interference present in the input signal caused severe interference on the pulse.

(2) Addition to the input signal of other signals of much lower frequency, such as may result from unevenness in the belt drive or by the test rotor swinging at its natural resonant frequency, causes the pulses to be intermittently formed and to vary in phase.

(3) Any maladjustment of pulse duration caused angular errors.

(4) The time constant of the cathode components still presents a major problem. The decrement time of the tuned amplifier is sufficiently large to make the circuit of Fig. 14 ineffective. Input signals due to large mechanical displacement or to electrical transients on starting also disable the circuit for quite long periods.

5.5.3. The Mixer Amplifier

These faults indicated that it would be advisable to develop a pulse forming circuit that:—

(a) Did not require a d.c. reference voltage to be developed from the input a.c. voltage.

(b) Formed the pulse symmetrically about the point where the instantaneous value of the a.c. input voltage was zero rather than at maximum value, and therefore when the rate of change of voltage was greatest. This should give a greater signal-to-noise discrimination, a pulse of greater amplitude, and allow greater tolerance in the factors affecting pulse width.

If two sinusoidal voltages of amplitudes $\hat{E}q$ and $\hat{E}a_3$, 90 deg. out of phase, are added, the resultant will be a voltage displaced in phase by an angle ξ from the larger voltage towards the smaller, where

$$\tan \xi = \frac{\hat{E}q}{\hat{E}a_3}$$

so that if we add eq , to both ea_3 and ek_3 (see Fig. 6), the resultant voltages eg_3 and eg_5 will be $(180^\circ + 2\xi)$ out of phase, as shown in Fig. 15.

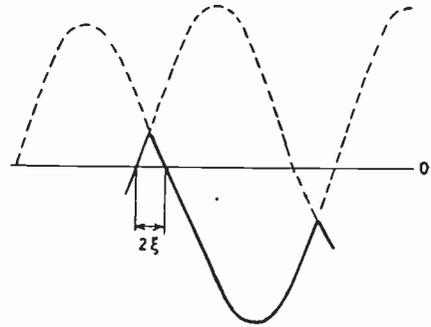


Fig. 15. Resultant voltages obtained in the mixer amplifier.

Should the more negative of the instantaneous values of these two voltages be chosen by a suitable rectifier circuit, the resultant will be a voltage of the form shown by the solid line in Fig. 15. It will be seen that this voltage is more negative than the value of the d.c. component of the a.c. voltages at all times except for a period 2ξ , symmetrically positioned about the point where the original voltages ea_3 and ek_3 were zero. If we connect this resultant voltage to the grid of a valve whose grid cut-off point is at the same d.c. voltage as the mean value of the a.c. voltages, a negative pulse will be developed in the anode circuit during the period required.

For a pulse width of 30° ,

$$\xi = 15^\circ,$$

$$\tan \xi = 0.27,$$

Therefore $\hat{E}q = 0.27 \hat{E}k_3$

Referring to Fig. 16, it will be seen that ea_3 and ek_3 are fed to each grid of V5, and eq to the common cathode connexion. Between the grid and cathode of the one triode unit, the input voltage is the vector sum of ea_3 and eq ; and to the other triode, the vector sum of ek_3 and eq . The anode output signals are ea_4 and ea_5 as shown in Fig. 15.

The amplifiers are used with unbypassed cathode resistors and the stage gain is 11.5. Advantage is taken of the comparatively large output signals in the anode circuit of the mixer amplifier to measure the amount of unbalance. The AMOUNT OF UNBALANCE meter is connected across portion of the anode loads.

5.5.4. The Final Pulse Forming Circuit

The anodes of the mixer amplifier are capacitance-coupled to a rectifier bridge. The negative arm of this bridge, i.e. the junction of MR2 and MR3, is connected through a high resistance to the grid of V6A. The voltage waveform at the MR2 and MR3 junction is of the form shown by the solid line in Fig. 15. The function of MR4 and MR5 is to equalize the d.c. current passed by MR2 and MR3, so that a charge does not begin to accumulate on the coupling capacitors.

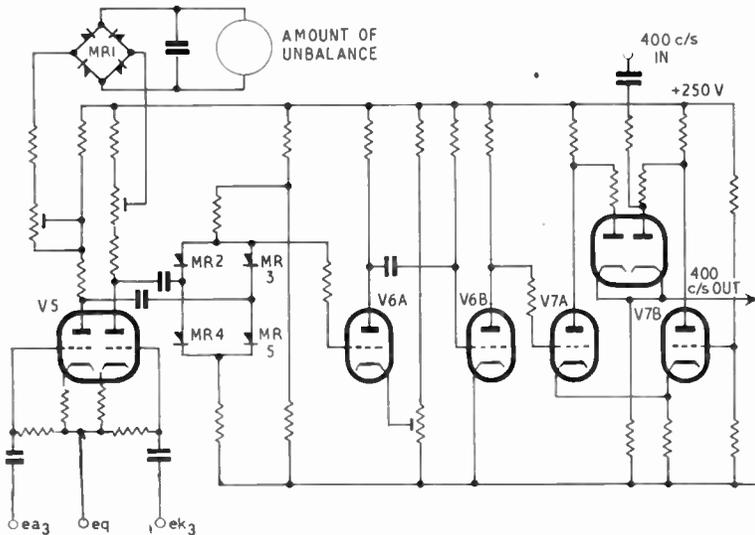


Fig. 16. Circuit diagram of the mixer-amplifier.

The voltage at the MR2, and MR3, junction adopts the value of whichever has the more negative instantaneous value, ea_4 or ea_5 . It will, however, become more positive than the average value of the input voltages during the pulse period, because of the small amount of positive bias on the bridge. The cathode of V6A is set by a preset potentiometer to a value such that V6A is cut off at all times of the input cycle, except the pulse period. During this period V6A will pass anode current and a negative pulse will be formed across its anode resistor.

The pulse is then amplified and squared by V6B and a long tailed pair, V7, gives two

outputs to the gate circuit, which are of opposite polarity. The conditions of operation of V7 are such that either one triode or the other is cut off, i.e. one anode only is at 250V positive, at any given time. During the pulse period V7B anode is at 250V positive, and for the remainder of the cycle V7A is 250V positive.

The AMOUNT OF UNBALANCE meter is an a.c. rectifier type volt-meter connected across a portion of the V5 anode loads. The large capacitor connected across the meter terminals is to reduce the needle flicker on low signal frequencies.

5.6. The Gate Circuit

The gate circuit originally used is shown in Fig. 17. In the absence of a pulse the diodes conduct and effectively short-circuit the 400 c/s input. The diodes stop conducting when a negative pulse is applied to the anodes and the 400 c/s input is then applied to the amplifier. In addition to the 400 c/s input to the amplifier, the pulse is also applied. This was considered as a possible advantage insofar as the output stage could be negatively biased during the period

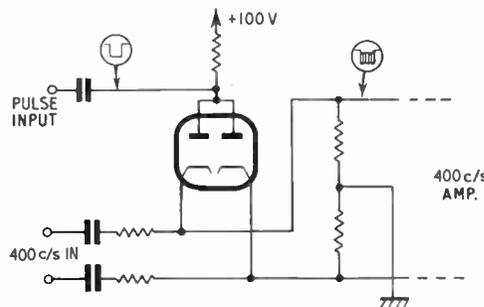


Fig. 17. The original gate circuit.

between pulses, and the average power dissipation of the final valves and power supply would be reduced. However, in practice the anode current variations in the final stage caused large variations in high tension voltage. It was very difficult to prevent inter-action with the input circuits. In order to remove this d.c. component from the gated input to the amplifier the final circuit shown in Fig. 3 was adopted. As either diode will be conducting, the cathodes will remain at a d.c. voltage approximately 200 V positive but only when the left-hand section of the double diode is conducting, i.e. during the pulse period, will the 400 c/s signal be passed to the amplifier.

5.7. The 400 c/s Oscillator

In order to use standard components, a phase-shift type of oscillator is used. The frequency stability is not required to be high and the frequency can be adjusted during the test period by varying the centre resistor in the RC network.

The anode load of the oscillator is composed of two potentiometers fed in parallel; the voltage at the arm of one is fed via the gating circuit to the power amplifier. The other potentiometer arm by-passes the gating circuit and feeds directly into the amplifier. The RUNNING/STATIONARY relay selects the potentiometer used; the gating circuit is used during RUNNING conditions only. The potentiometers allow the 400 c/s input to the synchros to be individually set for each condition. Because of the low mark-space ratio of the pulses it is possible to increase the power to, and hence the torque from, the synchros during the RUNNING period.

5.8. The 400 c/s Power Amplifier

This amplifier follows normal practice. The output stage is capable of 60 VA output, and the 30 db feedback gives a low output impedance. This is necessary because the power requirements of the synchros increase sharply when they are not in corresponding positions relative to each other. They will be so until the POSITION meter assumes correct indication.

The loading on the power amplifier is the impedance of two synchro rotors in parallel.

When the two synchros are in corresponding angular positions there will be no stator current, so that we can consider the amplifier as feeding into the parallel rotors of two synchros that have open circuit stators.

From manufacturer's data⁷:

Rotor impedance of 1½" synchro (indicator),
81 + j560 ohms.

Rotor impedance of 2" synchro (driven),
28 + j166 ohms.

Hence impedance of the rotors in parallel is
21 + j127 ohms.

In order to make the loading more resistive, a 2µF capacitor is connected in parallel with the rotors. With this capacitor the load impedance is 149 + j300 ohms.

A more resistive load would be desirable but the angular error in indication increases with capacitance, because the circuit becomes less able to handle transients. The value of 2 µF was selected as a compromise between angular error and load power factor. The angular error at 20 c/s pulse repetition frequency is 4 deg, using the 2 µF capacitor.

6. Acknowledgments.

The writer wishes to thank E.M.I. Electronics Limited for permission to submit this thesis, and to acknowledge the supervision of Mr. D. P. Franklin, and the assistance of Mr. A. F. Stevens on the project described.

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THE 1957 CONVENTION

“Electronics in Automation”

June 26—July 1

FOR this year's Convention the Institution will be returning to Cambridge. The accommodation and lecture theatre arrangements are similar to those enjoyed at the 1951 Television Engineering Convention.

Following the usual pattern of Brit.I.R.E. Conventions, the programme has been divided into six sessions in order to provide reasonable coverage of the theme of the Convention. Some thirty main papers will be read in the Clerk Maxwell and Green Lecture Theatres in the Cavendish Laboratory; discussion groups will also be organized in the Reading Room and elsewhere in King's College itself.

The Third Clerk Maxwell Memorial Lecture will be delivered on Thursday evening, June 27th, by Professor Sir Lawrence Bragg, F.R.S. The Convention Banquet will take place in King's Hall on Friday evening, June 28th.

The majority of Institution members and other delegates who are attending the whole of the Convention will be accommodated in King's College. Allocation of available rooms will be made through the Institution; members and others who will *not* be staying in College, or who are only attending for *part* of the Convention, will be required to make their own accommodation arrangements.

It is hoped that members will arrive by 7 p.m. on Wednesday, June 26th. To ensure accommodation in King's College completed registration forms must be lodged with the Institution by April 12th next. Early application is advised.

The inclusive charge for members *only*, for the entire Convention, is £12 10s. This includes accommodation charges, meals, Convention Banquet and preprints. Delegates wishing to stay in King's College over Sunday night may do so at a small extra charge.



King's College Chapel from the "Backs"

PROGRAMME

Session 1: Office Machinery and Information Processing.

Session 2: Machine Tool Control.

Session 3: Chemical and Other Processes.

Session 4: Simulators.

Session 5: Automation in the Electronics Industry.

Session 6: Automatic Measurement and Inspection.

PROFESSOR JAMES CLERK MAXWELL

THE role of James Clerk Maxwell as the "father" of radio science is well known. His work is honoured by the Institution in the establishment of the Clerk Maxwell Memorial Lectures—first suggested by Leslie H. Bedford,* the tenth President of the Institution. Since 1945 the senior premium awarded by the Institution has also borne the name of Clerk Maxwell.

A detailed account of Clerk Maxwell's life was given in the inaugural Memorial Lecture by Professor G. W. O. Howe, which contains a reference to the memorial tablet set up in Westminster Abbey in 1931 on the centenary of of Maxwell's birth. Another memorial to Maxwell was unveiled on October 15th, 1956, by Sir George Thomson, Master of Corpus Christi College, University of Cambridge, in the Mitchell Hall, Marischal College, Aberdeen, on the occasion of the centenary of Clerk Maxwell's appointment as Professor of Natural Philosophy at the College.

The memorial panel is shown here and its central feature is the bronze portrait bust by C. d'O. Pilkington Jackson showing Clerk Maxwell as a young man under thirty during his tenure of the Chair from 1856 to 1860. The two devices flanking the inscription are the sign of Saturn, a reference to his paper on the "Stability of Saturn's

Rings," for which he received the Adams Prize at Cambridge in 1857, and the thermodynamics formula, $dp/dt = JCM$, which he used as a pun on his name.

Nearly 10 years earlier, Maxwell had worked on the theory of colour, and speaking of this era in Clerk Maxwell's life, Sir John Cockcroft has said "... the theory of colour is still a live subject and one of great importance in the future of television."†

When the fusion of Marischal College and King's College, Aberdeen, took place, the two Chairs of Natural Philosophy were combined and, as the junior professor, Clerk Maxwell was retired on a pension for life. He was later appointed Professor of Natural Philosophy at King's College, London, where he remained until 1865. He then retired to an estate at Glenlair, in Gallogway, which he had inherited from his father, to prepare his "Treatise on Heat" and "Electricity and Magnetism."

In 1871 he was appointed to the new Chair of Experimental Physics at Cambridge, where he designed and superintended the erection of the Cavendish Laboratory on which some information is given on pages 3 and 4 of this *Journal*.

James Clerk Maxwell died in Cambridge on November 5th, 1879, aged 48.

† The Second Clerk Maxwell Memorial Lecture, by Sir John Cockcroft, K.C.B., C.B.E., F.R.S., *J. Brit. I.R.E.*, 14, p. 389, September 1954.

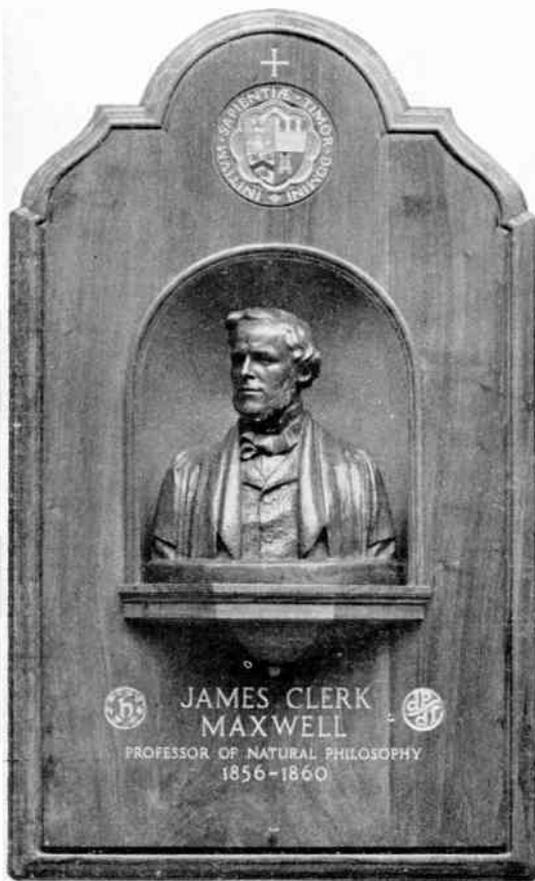


Illustration by courtesy of "Nature"

* Presidential Address, *J. Brit. I.R.E.*, 9, p. 2, January 1949.

MATERIALS USED IN RADIO AND ELECTRONIC ENGINEERING

A Survey by the Technical Committee of the Institution

5. THE ELECTRODEPOSITION OF METALS *

1. Introduction

The study of electrochemistry originated in the latter part of the eighteenth century when Volta developed the electrochemical pile which bears his name; this discovery was followed by many others, the most outstanding contribution being that of Michael Faraday in 1833. By that time many investigators were interested in the practical aspects of electrodeposition, and the first patent dealing with this process was that granted in England to Elkington and Barratt in 1838. This patent dealt with the use of electroplated coatings for decorative purposes, and two years later Elkington was granted a patent covering electrodeposition of silver; it is interesting to note that the process described therein is very similar to that in commercial use at the present time. Almost one hundred and twenty years have elapsed since the granting of Elkington and Barratt's patent, and during this period the electroplating industry has grown continuously and is now extremely large; its importance is best judged by the fact that electroplated coatings are an essential feature of either the products or the productive machinery of practically every branch of industry.

Electrodeposition is defined as the production of a layer of one metal on another, by making them the anode and cathode in an electrolytic cell containing a solution of a salt of the metal to be deposited. The two common methods in practical use are termed vat plating and barrel plating. In both cases the parts to be plated are immersed in a vat containing the electrolyte but whereas in the first main method the parts are individually suspended or held on wires or

racks, in the second method they are collectively tumbled in a rotating barrel. The parts to be plated function as the cathode in both methods. In barrel plating a special contact which forms part of the barrel structure completes the circuit.

The growth of the electroplating industry has been accompanied by a continuous increase in the number of metals and alloys which are commercially available as electroplated coatings and by substantial developments in electroplating technique. As an example of the former trend, the electrodeposition of indium, rhodium and copper-tin alloys may be cited, while the latter trend can best be illustrated by a brief description of the modern practice for bright nickel plating.

Electrodeposited nickel is chiefly used as an undercoating for decorative chromium plating, and in view of the wide popularity of this finish, large scale automatic or semi-automatic plants are frequently employed. The electrolyte—more popularly known as the solution—is continuously agitated and is heated to increase the rate of electrodeposition; heaters and thermostatic controls are therefore required. The solution is continuously circulated through an electrolytic purification tank and then through filters; the electrolytic purification tank is an auxiliary bath where the principal metallic impurities, iron, copper, lead, and zinc, are electrodeposited at suitable current densities. The nickel anodes in the main vat are surrounded by calico bags to prevent insoluble matter from contaminating the solution, and, in addition to the main constituents in the electrolyte, organic addition agents are employed which may be classified as follows:—

- (a) Brighteners, for the production of a bright deposit.

* Report approved by the General Council for publication on 11th December, 1956. (Report No. 12.) Based on a report prepared by E. E. Webster (Member) and R. W. Stobbs, F.I.M., F.R.I.C.
U.D.C. No. 621.357.7.

- (b) Wetting agents, added to minimize pitting.
- (c) Levelling agents, which result in a deposit with improved ability to fill up small scratches or other damage marks.
- (d) Anti-stress agents, the object being to reduce the internal stress which is an undesirable feature of bright deposits.

From this brief description it will be obvious that it is impossible to describe adequately the principles and practice of electroplating in the present report, but full details are available in the books and other publications cited in the Bibliography.

2. The Uses of Electroplated Coatings

Electroplated coatings are utilized for several purposes, and it is convenient to classify these in the following way, although it must be remembered that most types of electrodeposited finish will fulfil several of these purposes:—

- Decoration.
- Oxidation and corrosion resistance.
- The reduction of potential differences at bimetallic joints.
- Electrical and magnetic properties.
- Solderability.
- Wear resistance.
- Bearing properties.
- Electroforming.
- The rectification of components.

2.1. Decoration

The production of a decorative finish was the primary object of electroplating when the industry began over one hundred years ago; for example, silver plating resulted in a bright attractive finish similar to that obtained with solid silver, and yet very much cheaper. The amount of decorative electroplating undertaken to-day is very large and probably exceeds half the total value of the work handled by the industry, but it is obvious that a decorative finish must also possess some resistance to oxidation and corrosion, for were this not the case the coating would fail to fulfil its purpose. The decorative aspect will not receive further consideration in the present report, but it should be remembered that in many engineering applications a pleasing appearance provides a

useful adjunct to the main function of the coating.

2.2. Oxidation and Corrosion Resistance

Both ferrous and non-ferrous metals are electroplated in order to improve their resistance to the particular environment which they are likely to encounter in service, and the atmospheric conditions to which electronic equipment may be exposed are by no means constant. On the one hand the environment may only comprise the interior of private dwellings in temperate climates, conditions which may be classified as mild; conversely, however, the equipment may be exposed to tropical conditions where high temperatures and high humidities, combined with periodic cycling through the dew point, can result in rapid corrosion if the components are inadequately protected. Such conditions are classified as severe, a term which also applies to marine atmospheres where the air is contaminated by salt, and industrial atmospheres.

The protection of iron and steel by electroplating is achieved in one of two ways:—

- (a) By electrodeposition of a coating which possesses an intrinsic resistance to corrosion superior to that of iron, and which, in addition, is anodic to the iron in the presence of an electrolyte. Cadmium and zinc are typical examples of this type of coating which confers a definite protective value at small exposed areas of the basis metal by electrolytic action, i.e., the coating is attacked in preference to the base metal.
- (b) By electrodeposition of a coating which possesses an intrinsic resistance to corrosion superior to that of iron, but which is more noble than (i.e., cathodic to) the iron in the presence of an electrolyte. This is the case with electrodeposits of copper, nickel, chromium, tin, lead, silver, and gold. With this type of electroplating the coating must be free from porosity if complete protection is to be achieved. If any pores are present the attack on the basis metal in the presence of an electrolyte is usually accelerated, i.e. the basis metal is anodic to the coating.

In both the above cases the most satisfactory way of increasing the protective value of the coating is to increase its thickness. With zinc and aluminium, however, electrodeposited metals are more noble than (i.e., cathodic to) the basis metal, and hence the coating must be thick enough to be practically free from porosity if adequate protection is to be achieved.

2.3. *The Reduction of Potential Differences at Bimetallic Joints*

The corrosion of metals at bimetallic joints is dependent upon the presence of an electrolyte and is accelerated by the inherent potential difference between the two metals in that electrolyte, i.e., a galvanic couple is established. Due to condensation and the presence of traces of impurities in the air such as carbon dioxide, sulphur compounds, and especially salt in marine atmospheres, the presence of an electrolyte is by no means infrequent and hence the potential difference between metals in contact should be as small as possible if corrosion is to be avoided. (See Section 6 and Tables.)

2.4. *Electrical and Magnetic Properties*

It is invariably necessary to obtain good electrical conductivity at the mating faces of switches, plugs, sockets, and other connectors. The use of silver plating for this purpose is well known and is very useful, especially where maximum high frequency conductance is required. In cases where wear resistance is also required, the electrodeposition of rhodium applied over the silver results in a very satisfactory contact.

An interesting example of the use of coatings for their magnetic properties is the employment of electrodeposited cobalt-nickel alloys in magnetic storage devices for electronic computers.

2.5. *Solderability*

In the electronics industry it is frequently necessary to provide a surface coating to which soldered connections can readily be made using non-corrosive fluxes, and, in the case of steel, cadmium plating provides an excellent coating

for this purpose, as well as providing adequate protection against corrosion. Tin-zinc and silver may be easily soldered using non-corrosive fluxes provided that the surface of the coating is not tarnished, and, in the case of tin, the migration of alloys from certain basis metals is prevented by the prior deposition of a thin coating of nickel.

2.6. *Wear Resistance*

It is sometimes necessary to provide a wear resisting surface on cams, rollers, levers, and similar mechanisms where loss of dimensional accuracy may lead to faulty performance in service. Both heavy nickel and hard chromium electroplating are suitable for providing a wear resistant surface on components, but in cases where severe wear is likely to be encountered, chromium is undoubtedly superior as, in addition to a high hardness value, it also possesses a low coefficient of friction. The use of rhodium on electrical contacts was briefly referred to in 2.4 above.

2.7. *Bearing Properties*

The properties required in bearings used on engineering equipment are subject to wide variation, depending upon the application involved. In the case of shafts, hard chromium plating often provides an excellent bearing surface on mild steel components, but if heavy loads are involved a stronger basis metal is essential.

Electrodeposited coatings are also used for the production of bearings in highly stressed applications, a very interesting example being the use of silver followed by lead and indium in the production of aero engine bearings. Electrodeposited silver possesses high bearing-fatigue resistance and non-welding properties under conditions of marginal lubrication, and a typical bearing consists of a steel shell which is first given a very thin coating of nickel followed by about 0.030 in. of silver; the silver is then electroplated with lead which facilitates "running-in," and finally with indium, the function of this last coating being to improve the corrosion resistance of the bearing to modern engine lubricants.

2.8. *Electroforming*

Electroforming may be defined as a process for producing components by the electro-deposition of metal on to a base which is an exact but negative replica of the article required. The process is employed for making components which must possess the following features:—

- (a) A high degree of dimensional accuracy.
- (b) Intricate detail.
- (c) A high standard of surface texture.

Important uses of electroforming include the production of metal master patterns for gramophone records, and in recent years the manufacture of fittings for radar, waveguides, and computing cams.

2.9. *The Rectification of Components*

Components which have worn in service to such a degree that they are below the permissible dimensional limits can be frequently reclaimed by electroplating. The two methods in common use in this country are nickel plating and hard chromium plating. Nickel provides good corrosion resistance together with moderate wear resistance, while chromium confers an excellent degree of wear resistance combined with good corrosion resistance. The two coatings are not interchangeable as each possesses certain advantages for specific applications.

3. **Preparation of the Basis Metal**

The pre-treatment of the basis metal prior to electroplating is extremely important, as unless satisfactory adhesion is obtained this coating cannot fulfil its proper function. The objects involved in the pre-treatment of the base metal are clearly stated in the Ministry of Supply Process Specification DTD.901C as follows:—

- (a) They shall be free from all grease, rust, scale, and residue from heat treatment, welding, and allied processes.
- (b) They shall be in a suitable condition for the application of adherent coatings.
- (c) Their mechanical properties shall be substantially unimpaired.
- (d) Their dimensions shall be substantially unaltered.

(e) Their surfaces shall not have been roughened unduly.

(f) Their protective coatings shall be applied without delay.

The DTD specification referred to above only deals with the pre-treatment of steel and copper alloys. In recent years, however, the electroplating of zinc and aluminium base alloys and stainless steel has become of considerable importance, and the preparation of the basis material prior to electroplating is clearly outlined in the specifications of the American Society for Testing Materials.

4. **Electrodeposited Finishes**

In this section the principal features of those electrodeposited finishes which are of interest to the electronics industry are listed in alphabetical order. In addition, the British specifications and related data are summarized in the Appendix for easy reference.

4.1. *Brass*

Brass is a generic title given to alloys of copper and zinc, and can be readily electro-deposited as such. It is occasionally employed for decorative purposes, but its principal engineering use is to provide adhesion in the bonding of rubber to metals, and it is widely utilized for this purpose. The copper content of the electrodeposited metal should be between 72 and 78 per cent. in order to produce the best adhesion.

4.2. *Cadmium*

The primary function of cadmium plating is to provide a corrosion resistant coating, although its good solderability and appearance are also of importance. It is readily soldered with the aid of non-corrosive fluxes, and is the normal protective process used in this country for electronic equipment. The electrodeposition of this metal is covered by specification DTD.904, which stipulates that the minimum local thickness of cadmium shall be 0.0003 in. or alternatively the average thickness shall be 0.0004 in. The minimum of 0.0003 in. of electrodeposited cadmium confers a substantial degree of protection against corrosion, but it

is extremely difficult to obtain this thickness on mating parts, and in consequence the DTD.904 specification permits an average thickness of 0.0002 in. in certain cases. It must be noted, however, that in the latter instance the degree of protection is definitely reduced.

The corrosion resistance of cadmium is improved by a treatment known as chromate passivation, and this treatment is dealt with in Section 5. Steel which has been electroplated with cadmium and then given the chromate passivation treatment is capable of withstanding prolonged exposure in marine and tropical conditions, although in the latter case it should also be sprayed with a fungistatic varnish. It is definitely superior to zinc in such environments. In industrial areas, however, where sulphur compounds are likely to be present in the atmosphere neither cadmium nor zinc are satisfactory, although zinc is sometimes used.

An important use of electrodeposited cadmium consists of the electroplating of springs for marine or tropical service. In such cases the treatment prior to plating should preferably consist of light shotpeening* instead of acid pickling unless the working stress is very low, and a stress relieving heat treatment after plating in order to remove hydrogen embrittlement is essential. Springs treated in this fashion have given excellent performance in a wide variety of applications, but when rapidly fluctuating stresses are involved, a typical example being valve springs in internal combustion engines, superior results to electrodeposition are obtained by shotpeening and applying a zinc coating with a metal spray.

4.3. Chromium

Electrodeposited coatings of chromium may be divided into two distinct types known in general engineering practice as decorative chromium and hard chromium plating respectively. In the case of decorative chromium the deposit is extremely thin, B.S.1224:1953 stipulating at minimum of 0.00001 in., and when applied to steel should always be

preceded by electrodeposited nickel or nickel and copper. The steel is first electroplated with copper which possesses the advantage that it is cheap and can be readily polished, thus permitting the use of steel with a relatively inferior surface finish. The electrodeposited nickel is then applied, the aim being to provide a coating which is substantially free from porosity. Finally, a coating of chromium results in a surface which is highly resistant to tarnish and which retains its bright appearance over long periods. The British Standard specification referred to above permits the use of four different thicknesses of metal as an undercoat for decorative chromium plating, and is intended to allow for service conditions which vary from very mild to relatively severe applications such as outdoor use in coastal areas.

When a chromium-plated article shows premature deterioration in the form of numerous spots of rust, this is probably due to an inadequate thickness of nickel which facilitates the corrosion of the basis metal by electrolytic action of the type referred to in Section 2.2 of this report.

In addition to the use of chromium plating for decorative purposes, chromium is now widely employed in engineering on account of its excellent wear resistance. Its high hardness, low coefficient friction, non-galling and non-wetting properties, have proved of benefit in a wide variety of applications. The life of many tools such as drills, reamers, gauges, and cold forming dies of all types is greatly increased by the judicious application of electrodeposited chromium of 0.0005 in. in thickness. The rectification of worn parts was referred to in Section 2, and a further use of chromium consists of the electrodeposition of "porous" deposits which retain lubricants and thus reduce the wear of cylinder liners and similar components.

4.4. Copper

Electrodeposited copper is principally employed as an initial coating for decorative chromium as mentioned above. It is also used for the electroforming of components, as an

* Shotpeening is a surface treatment employing bombardment under carefully controlled air pressure.

undercoating on zinc base alloys, and for the prevention of carburizing in the selective case-hardening of steel components.

4.5. *Gold*

Gold is employed as an electrodeposited coating for purposes where its excellent resistance to tarnish over a wide range of temperatures and service conditions counter-balances its high cost. It is employed on electrical contacts, and has recently been used on reflectors for industrial infra-red radiation heating.

4.6. *Nickel*

The use of nickel as an undercoating for decorative chromium plating has already been referred to in the section dealing with the latter metal. Nickel plating is also employed because it possesses good corrosion resistance in a wide variety of service conditions. It also confers satisfactory oxidation resistance for mild steel components which are required to function for relatively short periods at temperatures up to 650° C. Electrodeposited nickel has proved extremely useful in the building up of worn components, and in cases where both a heavy deposit and a high degree of wear resistance is involved the nickel coating can be followed by electrodeposited chromium.

4.7. *Rhodium*

Electrodeposited rhodium possesses several admirable characteristics from a radio and electronics engineer's point of view. Its initial hardness as deposited is of the order of 800 V.P.N. (Vickers pyramid hardness test number), and it has a high degree of resistance to both wear and tarnish. It has been found to give excellent service when directly electroplated on to beryllium, copper, or brass; a deposit of 0.0002 in. is adequate, and lubricants can be applied in some cases without detriment to the conductivity of contacts. It may also be employed over electrodeposited silver for similar reasons. In passing, it may be mentioned that while both platinum and palladium can be successfully electroplated, rhodium possesses superior properties in most cases.

4.8. *Silver*

This metal is widely used in the electronics industry for providing surface conductivity on chassis, contacts, and similar components, especially when high frequency currents are involved or low contact resistance required. Although the deposited metal exhibits a relatively low hardness it work-hardens with use and has been found to remain on the surfaces of pins and sockets even after extended periods of use. The deposit only confers a limited degree of corrosion resistance to the base metal and readily tarnishes, particularly in the presence of sulphur compounds. It is interesting to note, however, that practical experiments have shown that the tarnish has no adverse effect on the electrical conductivity. A further advantage of silver plating is its excellent solderability, provided that it has not been stored for long periods.

As was mentioned earlier, electrodeposited silver has been found to possess an extremely high resistance to bearing fatigue and is widely used for heavily loaded bearings. In such cases the silver is subsequently coated with electrodeposited lead and indium, or a lead-tin alloy.

4.9. *Tin*

Electrodeposited tin is normally used as a coating for brass terminals which are subsequently soldered, but cannot be recommended for tags which are stored prior to soldering. Once the surface becomes tarnished the solderability is greatly impaired. Due to this undesirable feature, a superior method of achieving solderability is the use of hot tinning applied by dipping the components in molten tin or tin-lead alloy. In all cases, however, an initial electrodeposit of 0.0001 in. of nickel is recommended as it facilitates solderability. When electro-tinned brass is stored, the migration of tin into the basis metal impairs solderability, while in the case of hot tinning the nickel prevents contamination of the molten tin with zinc.

4.10. *Zinc*

Zinc occupies an important place among electrodeposited finishes as its corrosion resistance, especially after passivation (Section

5), is only slightly inferior to that exhibited by cadmium, except in marine and tropical environments where cadmium is to be preferred.

An important point in connection with the use of zinc plating is that its electrode potential (Section 6 and Table 1) is appreciably higher than that of cadmium, and very much greater than that of silver, and zinc must therefore be used with caution. A further drawback is that when non-corrosive fluxes are employed the solderability of zinc is low. As a result, although many attempts have been made to replace cadmium by zinc due to the lower cost of the latter metal, serious limitations are involved. A more satisfactory alternative is the tin-zinc alloy plating referred to in the following section.

4.11. Alloy Deposition

Several alloys may be electrodeposited from solution in a similar manner to the substantially pure metals dealt with above, but it should be mentioned that a closer control of the electroplating process is essential in such cases. In addition to the electrodeposition of brass, the alloys of interest to the electronics industry are:—

- Cobalt-nickel alloys.
- Copper-tin alloys.
- Tin-nickel alloys.
- Tin-zinc alloys.

4.11.1. Cobalt-nickel alloys

In recent years the electrodeposition of cobalt-nickel alloys has proved of importance in the manufacture of magnetic storage devices where the high coercivity and substantial remanence have facilitated high computation speeds.

4.11.2. Copper-tin alloys

Two basic compositions in this alloy system have been found to possess a useful combination of properties. An electrodeposit of tin-bronze containing 10–12 per cent. of tin provides a reasonable substitute for nickel as an undercoating for decorative chromium, except where a high degree of corrosion resistance is required.

A second copper-tin alloy which has proved of interest is speculum metal, an alloy con-

taining 42 per cent. of tin, remainder copper. This electrodeposit will take an extremely high polish and is suitable for indoor use, although its corrosion resistance is inadequate for outdoor applications.

4.11.3. Tin-nickel alloys

The electrodeposition of tin-nickel alloys has been the subject of considerable investigation, the object being to minimize the quantity of nickel used in the production of decorative chromium plating, but at the present time it has not been adopted to any appreciable extent.

4.11.4. Tin-zinc alloys

This type of alloy plating is growing in importance and provides a satisfactory substitute for cadmium in many cases, although cadmium is superior when severe conditions of exposure are involved. The alloy possesses good solderability and is now being used for the electroplating of radio chassis, loudspeaker frames, and similar components.

5. Treatment after Electroplating

Two production processes are widely employed for the treatment of electrodeposited metals. The first of these is a heat treatment designed to minimize the deleterious effects due to absorption of hydrogen during the electroplating process. All ferrous springs, components made from steels of over 45 tons ultimate tensile strength, and all case-hardened steels, must be given a heat treatment after plating in order to remove hydrogen embrittlement. A suitable treatment consists of heating the components for one hour at 150° C. While this is effective in the majority of cases, it does not completely nullify the harmful effects of hydrogen absorbed as the result of excessive acid-pickling.

The second process which is in general use is the chromate passivation of electrodeposited cadmium and zinc. By immersing the electroplated components in an aqueous solution of sodium dichromate and sulphuric acid, a thin film of basic chromium chromate is formed which, although being only of the order of 0.00002 in. in thickness, greatly retards the formation of corrosion products on cadmium and zinc. The film produced by this

method should be a light iridescent gold or green in colour. Heavy khaki shades indicate over-treatment which is undesirable and does not help corrosion resistance. The process is covered by Specification DTD.923A, and while this refers specifically to zinc, it is equally applicable to electrodeposited cadmium.

6. The Selection of Electrodeposited Finishes

From a study of Section 2 of this report, it is obvious that many factors govern the

selection of electrodeposited metal finishes, and cases where only one requirement must be satisfied are extremely rare; in most cases a combination of properties is required.

The selection of suitable finishes for electronic equipment is clearly dealt with in a specification issued by the Radio Industry Council, R.I.C./1000/B, and definite recommendations are made for specific applications. In using the data provided by that specification, the question of potential differences at bimetallic joints must

Table 1

The Electrode Potentials of Metals against a Saturated Calomel Electrode in Sea Water at 25° C.

Material	Volts	Material	Volts
Magnesium and its Alloys	-1.60	Terne Plate	-0.50
Zinc and its Alloys:—		Tin and its Alloys (see also Tin-Zinc above):—	
Zinc die-casting alloy, B.S.1004 ...	-1.10	Tin-Lead Solders, B.S.219A and B ...	-0.50
Zinc plating on steel to D.T.D.903 ...	-1.10	Tinned Steel to B.S.S.20	-0.50
Zinc plating as above, chromate passivated to D.T.D.923	-1.05	Tin Plating on Steel to D.T.D.924 ...	-0.45
Galvanized Iron B.S.729	-1.05	Chromium:—	
Tin-Zinc Alloy Plating on Steel ...	-0.05	Chromium Plating, 0.0005" on steel ...	-0.50
Cadmium Plating on Steel to D.T.D.904	-0.80	Chromium Plating, 0.00003" on nickel plated steel	-0.45
Aluminium and its Alloys:—		Chromium Metal (99%)	-0.45
Alloy sheet clad with 1% Tin Alloy D.T.D.687	-0.90	Copper and its Alloys:—	
Casting Alloys LM4, LM6, LM9 ...	-0.75	Copper, Brasses, Bronzes, Beryllium-Copper, Cupro-Nickel (70/30) ...	-0.25
Wrought Aluminium B.S.1470-1477, IC, 3L.16	-0.75	Nickel and its Alloys:—	
Alum.-Manganese Alloy B.S.1470-1477, N4	-0.75	Nickel-Copper Alloys B.S.1526, 1529, 1532, 1535	-0.25
Alum.-Manganese Alloy B.S.1470-19477, N4	-0.75	Nickel Plating on Steel, B.S.1224, Ni8S	-0.15
Aluminium coated Alloys B.S.S.L.72	-0.75	Titanium	-0.15
Duralumin type Alloys B.S.1470-1477, H.14 and H.15, B.S.L.64 ...	-0.60	Silver and its Alloys:—	
Irons and Steels:—		Silver Solder, B.S.1845, Type 3 ...	-0.20
Not corrosion resisting—mild steel, low alloy steel, Gleyr Cast Iron	-0.70	Silver	0
Corrosion resisting—12% Chromium, B.S.S61, En.56, B.S.1630 ...	-0.45	Silver Plating on Copper	0
18% Chromium, B.S.S80, En.57	-0.35	Silver-Gold Alloy (not less than 10% gold)	+0.05
Austenitic 18/8, B.S.1631, 1632, En.58, B.S.980—CDS.20 ...	-0.20	Electrical Contact Metals:—	
Lead and its Alloys:—		Rhodium Plating on Silver Plated Copper	+0.05
Lead, B.S.602, 1178	-0.55	Gold	+0.15
Lead-Silver Solder (2.5% silver) ...	-0.50	Platinum	+0.15
		Carbon (colloidal graphite in acetone, evaporated to dryness)	+0.10

be borne in mind, and a table of the electrode potentials of metals against a saturated calomel electrode in sea water is provided therein; this is reproduced as Table 1 of the present report. The Radio Industry Council recommends that the maximum potential difference at bimetallic joints be restricted to the values given in Table 2. Further data on this problem is provided in item 6 in the Bibliography.

7. Methods of Testing

It is necessary to establish the thickness of the deposit on an average basis or of a local area to establish minimum thickness. In addition to assessing thickness, it is often necessary to establish freedom from porosity as discussed in section 2.2, and the adhesion of electroplated coating to the basis metal also requires measurement.

Of the above assessments, the most important one is the determination of thickness, and this is normally carried out in one of the following ways, these being listed in order of their reliability:—

(1) *Strip and weigh*

This determines the average thickness of the coating.

(2) *Microscopic examination on a cross-section*

This determines the local thickness at any point across the section.

(3) *B.N.F. jet test*

This determines the local thickness at coating on a particular spot which is chosen for the purpose.

(4) *Gassing test*

This gives a quick assessment of the minimum and maximum amount of coating.

(5) *Magnetic and thermo-electric methods*

These tests give information relative to the maximum and minimum thickness of the coating.

The above methods exclude a simple fundamental method of determining thickness of coatings, this being the mechanical measurement of parts before and after plating;

Table 2

Maximum Potential Difference (in sea water)	Condition of Service
0.3 volt	Liable to wetting with salt water or normally exposed to the weather.
0.5 volt	Interior parts which may be exposed to condensation but not to contamination with salt.
No restriction	Interior parts of hermetically sealed components and equipment.

providing the parts selected for measurement can be accurately assessed by comparative means, the method has many advantages over and above those stated above.

Porosity checks are simple in nature and normally consist of immersing the plated part in a solution which will demonstrate porosity by the evolution of gas through pin holes, although this is only applicable to coatings which are anodic to the basis metal.

Adhesion is determined by flexing or bending the plated part, or rubbing it heavily with a suitably smooth metal tool to see whether the coating will be detached by this treatment.

All the tests described, with the exception of (5), are of course destructive by nature and the parts submitted are therefore scrap. It is important, therefore, to remember when carrying out any of the tests specified, that the sample tested should be representative of the bulk which is going to be used, as otherwise the tests are invalid and give no information whatever as to the quality of work under consideration.

8. Bibliography.

1. "Modern Electroplating." Edited by A. G. Gray. (The Electrochemical Society, U.S.A.; Chapman and Hall, London, 1953.)
2. "Handbook of Industrial Electroplating." Ollard and Smith. (Iliffe and Sons, London, 1954.)

3. "The Electroplating Engineering Handbook." Graham. (Reinhold, New York, 1955.)
4. "Chromium Plating." Morisset, Oswald et al. (Robert Draper Ltd., New York, 1954.)
5. "The Corrosion Handbook." Edited by H. H. Uhlig. (The Electrochemical Society, U.S.A.; Chapman and Hall, London, 1948.)
6. "Corrosion and its Prevention at Bimetallic Contacts." The Interservice Metallurgical Research Council. (H.M.S.O., 1956.)
7. "Handbook on Electroplating." Canning. 18th Edition. (Birmingham, 1953.)

9. Specifications

B.S.1224:1953 Electroplated coatings of nickel and chromium. 16 pp. 3s.

Covers coatings of nickel and chromium on steel, copper alloy, zinc alloy. It gives limits of finish, thickness and adhesion and specifies a salt spray test. Appendices give methods of carrying out the tests.

B.S.1706:1951 Electroplated coatings of cadmium and zinc on iron and steel. 18 pp. 2s. 6d. Amendment PD 1383.

Provides for essential qualities of electroplated coatings of cadmium and zinc on iron and steel articles. It covers three classes of coating according to the conditions to which the article may be subjected. It also covers details of the requirements for passivated zinc coatings. The specification deals with the purity of the materials used for coating and with the finish and appearance of the plated surface, thickness of coating and adhesion. It includes appendices giving methods of test.

B.S.1872:1952 Electroplated coatings of tin. 12 pp. 2s. 6d.

Provides for the essential qualities of electroplated coatings of tin on fabricated articles of iron, steel, copper and copper alloys. It specifies requirements

for five classes of coatings according to the use to which the coating is to be put. It provides details of sampling, finish, solderability, thickness, adhesion and heat treatment, together with the necessary methods of test.

B.S.—:1957 Electroplated coatings of silver for engineering purposes.

(In preparation)

(British Standards are obtainable from the British Standards Institution, Park Street, London, W.1.)

DTD.903C Zinc Plating.

DTD.904 Cadmium Plating.

DTD.923A Chromate Passivation of Zinc Surfaces.

DTD.919A Electroplating of Aluminium, Steel and Copper with Silver and Nickel.

DTD.924 Electrodeposited Tin Coatings.

Interservices Specification RCS/1000 (Now DEF.5000) "General Requirements for Service Telecommunication Equipment." (Radio Components Standardization Committee.)

(D.T.D. and Interservice Specifications are obtainable through H.M. Stationery Office.)

R.I.C.1000B. "Choice of Finishes for Radio and other Electronic Equipment and for Components therein." (Radio Industry Council, London, 1950.)

(R.I.C. Specifications are obtainable from the Radio Industry Council, 59 Russell Square, London, W.C.1.)

10. Acknowledgments

Thanks are due to manufacturers and others who have given assistance in the preparation of this review.

11. Appendix:

Summary of British Standard and Ministry of Supply Specifications

Finish	British Standard Specifications		Ministry of Supply Process Specifications		Characteristics
	Number and Description	Local Thickness (minimum)	Number and Description	Local Thickness (minimum)	
CADMIUM (On iron & steel)	B.S.1706 Class A—Cd 4 (for exposure to marine or tropical conditions)	0.0004"			Excellent corrosion resistance. Good solderability (impaired by passivation)

ELECTRODEPOSITION OF METALS

Finish	British Standard Specifications		Ministry of Supply Process Specifications		Characteristics
	Number and Description	Local Thickness (minimum)	Number and Description	Local Thickness (minimum)	
CADMIUM (contd.)	Class B—Cd 3 Class C—Cd 2 (For use in dry interiors)	0.0003" 0.0002"	DTD.904. 0.0002" minimum average thickness is acceptable for barrel plated work.	0.0003"	Coating may be passivated if required according to DTD.923.
CHROMIUM (Protective and decorative)	B.S.1224 0.00001" minimum chromium plated over nickel as quoted below.		No DTD Specification.		A lustrous finish highly resistant to tarnish.
HARD CHROMIUM	No B.S. Specification.		DTD.916A, A.R.D. Electrodeposition memoranda Nos. 4 and 6. Thickness according to requirements. (No. 4 on steel; No. 6 special steels, cast iron and non-ferrous metals).		Excellent wear resistance; low coefficient of friction.
COPPER	No B.S. Specification.		No DTD Specification.		
(1) Protection for Heat Treatment (2) Oxidised finishes (3) Heavy deposits for Electro-forming	0.0005" min (Steel) 0.0003" min (Steel, Copper, Aluminium, Zinc). Thickness to be specified				Good initial coating under resistance; low decorative chromium. Seldom used as a finish.
GOLD	No. B.S. or other specifications				High resistance to tarnish. Used in very thin films for decoration and contact purposes.
NICKEL (including Nickel prior to Chromium plating) On Steel	B.S.1224 Ni 12 S for outdoor applications	0.0012" Nickel (or 0.0003" Cu and 0.0010" Ni)	DTD.919 (Nickel finish for Telecommunications Equipment)	Total thickness 0.0008" (of which not less than 0.0004" shall be nickel)	

TECHNICAL COMMITTEE REPORT

Finish	British Standard Specifications		Ministry of Supply Process Specifications		Characteristics
	Number and Description	Local Thickness minimum	Number and Description	Local Thickness (minimum)	
NICKEL (contd).	Ni 8 S Indoor and mild outdoor Ni 4 S Mild indoor	0.0008" Nickel (or 0.0003" Cu and 0.0006" Ni) 0.0004" Nickel		or 0.0004" total average with 0.0002" min. Nickel in the case of barrel plating	Undercoating for decorative Chromium. Used as a finish for decorative purposes and for the rectification of components.
	Ni 2 S Very mild indoor	0.0002" Nickel			
On copper or alloys containing over 50% copper	Ni 5 C Outdoor applications	0.0005" Ni		0.0003" (or 0.0003" average in the case of barrel plating)	
	Ni 3 C Indoor applications Ni 1 C Mild indoor	0.0003" Ni 0.0001" Ni			
On zinc alloy containing over 50% zinc	Ni 6 Z Outdoor applications Ni 3 Z Indoor applications	0.0003" Copper plus 0.0006" Ni 0.0003" Copper plus 0.0003" Ni	Not covered by DTD.919		
On Aluminium and aluminium alloys	Ni-A	Insufficient data to specify a figure		Total thickness 0.0015" of which .0004" shall be nickel	
HEAVY NICKEL (Coatings on ferrous, copper rich and aluminium rich materials)	No B.S. Specification		DTD.905 (See also A.R.D. Electro- deposition Memorandum No. 2)	Thickness according to requirements	
RHODIUM	No B.S. Specification				Hard non- tarnishable deposit. Excellent for achieving low contact resistance.
SILVER	B.S.S. Electroplated Coatings of Silver for Engineering Purposes (in preparation)		DTD.919 On Copper and Copper alloys On Steel	0.0003" average min. in case of barrel plating 0.0008" total coating of which 0.0003" shall be silver (or 0.0004" total with 0.0003" silver average in case of barrel plating	High conductivity excellent solderability when not tarnished.

ELECTRODEPOSITION OF METALS

Finish	British Standard Specifications		Ministry of Supply Process Specifications		Characteristics
	Number and Description	Local Thickness (minimum)	Number and Description	Local Thickness (minimum)	
SILVER (contd.)			On Aluminium and Aluminium rich alloys	0.0015" total coating of which 0.0003" shall be silver	
TIN On Ferrous Metals	B.S.1872 Class A. Sn12S Particularly for contact with foodstuffs Class B. Sn8S Protection against atmospheric corrosion Class C. Sn2S To facilitate soldering Class D. Sn5S Stop-off for nitriding Class E. Sn2AS for screws etc.	0.0012" 0.0008" 0.0002" 0.0005"-0.0001" (average thickness) 0.0002" (average thickness min.)	DTD.924 For protection of Steel To facilitate soldering Masking before nitriding For bolts and threaded parts less than 1/4" dia.	 0.0008" 0.0003" 0.0005" 0.0002"	Good solderability when not tarnished. Only suitable for mild conditions of exposure.
TIN On Non-Ferrous Metals	B.S. 1872 Class A. Sn12S Particularly with foodstuffs Class B. Sn3C Protection against atmospheric corrosion Class C. Sn3C To facilitate soldering Class E. Sn2AC For screws, etc.	0.0012" 0.0003" 0.0003" (or 0.0002" for copper) 0.0002" (av. thickness min.)	Protection of Copper and its alloys To facilitate soldering For bolts and threaded parts less than 1/4" dia.	0.0003" 0.0003" 0.0002" (av. thickness min.)	
ZINC On Iron and Steel	B.S. 1706 Class A. For marine and tropical atmospheres Class B. Class C. For dry interiors Coatings may be passivated. For barrel plating an average thickness equal to the appropriate local thickness is acceptable.	0.0010" 0.0003" 0.0002"	DTD.903	 0.0003" Coatings should be passivated according to DTD.923 if required.	Good corrosion resistance. Poor solderability

A special meeting will be held in London on 10th April next when Mr. R. W. Stobbs, who was jointly concerned in the preparation of this report, will present a paper on the electrodeposition of metals.

APPLICANTS FOR ELECTION AND TRANSFER

As a result of its January meeting the Membership Committee recommended the following elections and transfers to the Council. This list also includes elections and transfers to Graduateship which were considered at the meeting held in November, 1956.

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

Direct Election to Member

CLEMENT, Lewis Mason, B.Sc.(E.E.). *Washington, U.S.A.*

Direct Election to Associate Member

BALDWIN, William Edwin Bernard. *Burgess Hill.*
BATES, Samuel, B.Sc.(Eng.). *Stoke-on-Trent.*
BUTTINGER, Philip, B.Sc.(Eng.). *London, S.W.12.*
CAREY, Peter Herbison. *London, S.W.2.*
HEYES, William, B.Sc.(Eng.). *Sketmersdale.*
JOHNSTON, Capt. Walter, B.Sc., Canadian Army. *Minnedosa, Manitoba.*
PITMAN, Com. Stanley Robert James, R.N. *Lee on Solent.*
WEAVER, Albert Dudley. *Jersey.*

Transfer from Associate to Associate Member

VALE, Lindsay Harold. *Wells, Somerset.*

Transfer from Graduate to Associate Member

FLANNERY, Flt. Lt. Thomas James, R.A.F. *Penrith.*
JARVIS, Cyril Vernon, B.Sc. *Southend-on-Sea.*
TUDDENHAM, Thomas Douglas. *Longtown, Cumberland.*

Transfer from Student to Associate Member

HEMERY, Norman Valentine. *London, S.E.6.*

Direct Election to Associate

BASTIN, Andre Marie Jacques. *Harrow.*
JAMIESON, Lieut. Com. Alexander Hugh, R.N. *Singapore.*
WELLER, William Frank Edward. *Kenton.*

Transfer from Student to Associate

DAVIES, Brian Talbott. *London, N.8.*
FORD, David. *Faversham.*

Direct Election to Graduate

(Elected 10th January)

AUERBACH, Michael. *Nuneaton.*
GORE, William Edward Charles, B.Sc. *London, N.8.*
HEARN, Peter John. *Stevenage.*
MESSENGER, Michael Charles, B.Sc.(Hons.). *Accra.*
RABE, Arthur Guenter. *Welwyn Garden City.*
RICHARDSON, Colin Charles. *St. Albans.*
ROUGHTON, Derek John, D.L.C. *Richmond, Yorks.*
RÜTHERFORD, Norman Sydney. *Nottingham.*
SHARMA, Flt. Lt. Dharm Sheel, B.Sc., I.A.F. *Secunderabad.**
SINHA, Jayanti Kumar, B.Sc., M.Sc. *London, W.11.*
SUCKSMITH, Alan Vaughan. *Newton-Le-Willows.*
TRAWINSKI, Bohdan Stanislaw. *London, N.W.3.*
VENKATESWARAN, Flt. Lt. Gopalakrishnayyar, M.A., B.E., I.A.F. *Cranfield.**
YATES, John. *Manchester.*

(Elected 15th November)

BAKER, Ernest Norman, B.Sc.(Hons.). *Basing, Hants.*
BARNES, Derek Albert. *Mitcham, Surrey.*
BAUCH, Horace Douglas. *Plymouth.*
CAREY, Peter Majendie. *Englefield Green.*
CHAPMAN, Norman John. *London, S.W.16.*
CREESEY, Derek Robert. *Ashford, Middlesex.*
FOWLER, Clement William. *London, N.20.*
GRAY, Norman Frank. *London, S.E.14.*
HAZELL, Grahame Frederick, B.Sc.(Hons.). *Richmond, Surrey.*
HORAN, Ean Michael. *Brentwood.*
INGLETON, John. *Sanderstead.*
JAGER, William Frederick. *Kingston-on-Thames.*
JENNINGS, David Michael, B.Sc.(Eng.). *Cardiff.*
JOHNSON, Dalma Joseph Harvey. *Bletchley.*
JONES, David Lloyd. *Newark.*
KIRKHOPE, James. *Glasgow.*
LEVY, Allan. *Sutton, Surrey.*
LINGLEY, Peter James. *Morden.*
MACKENZIE, George White. *Evesham.*
NG EK POH. *Brighton.*
PASMORE, Tom. *Southall.*
PUTTICK, Arthur David John. *Godalming.*
SLACK, James Austin, D.L.C. *Brentwood.*
VENEIK, Amar Nath, B.Sc.(Eng.). *London, N.20.*

Transfer from Student to Graduate

(Transferred 10th January)

ALLEN, George. *Ilmminster.*
GILL, Ujarar Sing. *Singapore.*
PRAHARAJ, Arakhita. *Bombay.*
RANGARAJAN, K. S., B.Sc. *Salem, Madras State.*
SUBRAMONYAN, S. Harihari. *B.Sc. Kharapur.*
WALKER, James Martin. *Northwich.*

(Transferred 15th November)

BEBBINGTON, Roy Edward. *Yeovil.*
CERESA, Anthony. *Hayer, Middlesex.*
COLLINS, Vernon John William. *Richmond, Surrey.*
ELLERINGTON, William. *London, S.W.4.*
KANWAR, Randhir Singh. *Jodhpur.*
KOUREAS, Varnavas Demetri. *London, N.W.5.*
LAWSON, Bernard John. *Beckenham.*
LYSONS, John Michael Charles. *Chelmsford.*
MAHAJAN, Raj Kumar. *Bangalore.*
SHARMA, Kailash Prasad, M.Sc. *London, W.11.*
YOUNG, William Thomas. *Whitehouse, Co. Antrim.*

* Reinstatement.

Studentship Registrations

In addition to the above elections to Graduateship and higher grades of membership, the Council approved the registration of a further twenty-eight students.

THE DESIGN AND APPLICATIONS OF A GENERAL-PURPOSE ANALOGUE COMPUTER*

by

R. J. A. Paul, B.Sc.[†] and E. Lloyd Thomas, B.Sc.[‡]

SUMMARY

The various factors influencing the design and construction of a general purpose analogue computer are considered in detail, with particular attention to the effects of finite amplifier gain, bandwidth and phase shift. It is shown that one of the most important design criteria is the gain-bandwidth product, and that the bandwidth must be much greater than is generally supposed if the computer is required to deal with lightly damped systems. A description is given of a computer designed for industrial and research applications, based on this analysis and embodying various novel features. The instrument is capable of both repetitive and "single-shot" operation, and can also deal with non-linear problems. A selection of problems chosen to illustrate the wide range of a computer of this kind is then presented. A number of little-known applications, such as the analysis of mathematical series and the construction of economic "models," are included as well as examples of more familiar uses for aircraft simulation and servomechanism design. An indication is given of the lines along which analogue computers may develop in the future, and of other possible applications.

TABLE OF CONTENTS

1. Introduction	4.3.4. Continuous Function Unit
1.1. Choice of Computing Method	4.3.5. Discontinuous Function Unit
1.2. Principle of Operation of the Differential Analyser	4.3.6. Other units
2. General Design Considerations	5. Applications
2.1. Size and Scope of Computer	5.1. Solution of Differential Equations
2.2. Computing Elements	5.1.1. Mathematical analysis
2.3. Ancillary Equipment	5.1.2. Aircraft dynamic response
3. Detailed Design Considerations	5.1.3. Aircraft flutter
3.1. Amplifier Characteristics	5.1.4. Non-linear electric circuits
3.1.1. Gain	5.2. Simulation
3.1.2. Bandwidth	5.2.1. Auto-pilot response
3.1.3. Drift	5.2.2. Coupled mechanical systems
3.1.4. Grid current	5.2.3. Economic models
3.1.5. General comments on amplifiers	5.2.4. Flight simulation
3.2. Network Characteristics	5.2.5. Hysteresis loop
3.2.1. Resistors	6. Conclusions
3.2.2. Capacitors	7. Acknowledgments
3.2.3. Phase shift	8. References
3.2.4. General comments on networks	9. Appendices
4. Description of Computer	9.1. Effects of Finite Amplifier Gain
4.1. General	9.1.1. Adder
4.2. Presentation	9.1.2. Integrator
4.3. Computing Units	9.2. Influence of Finite Bandwidth
4.3.1. Linear Function Unit	9.3. Effects of Phase Shift
4.3.2. Scaling Unit	9.4. Effects of Drift (Offset) Voltage and Input Grid Current
4.3.3. Multiplier	

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[†] College of Aeronautics, Cranfield; formerly with Short Brothers and Harland Ltd., Belfast.

[‡] Short Brothers and Harland Ltd., Belfast. U.D.C. No. 681.142.

LIST OF SYMBOLS

V_1, V_2 etc.	— Input voltage to a feedback stage, relative to a common line.	i_p	— Grid current of amplifier input valve.
V_0	— Output voltage to a feedback stage, relative to a common line.	p	— Laplace operator.
v	— Amplifier input voltage of a feedback stage, relative to a common line.	$L^{-1} \{ \quad \}$	— Inverse operational form of $\{ \quad \}$
	<i>Note.</i> —The addition of a dash, v' denotes the operational form of v .	$-m$	— Open loop gain of computing amplifier.
Z'	— Laplace operational form of generalized impedance Z .	φ	— Phase error introduced by computing unit.
Z_1	— Generalized input impedance.	$\tilde{\omega}$	— Phase shift round computing loop.
Z_2	— Generalized feedback impedance.	ω	— Angular frequency.
V_d	— Equivalent input drift voltage of amplifier.	t	— Computing period.
		T	— Periodic time of waveform.
		τ	— Time lag in computing unit.

1. Introduction

During the last decade the volume and complexity of the calculations involved in the solution of many research and engineering problems have increased to the point where manual computation becomes impossible or impracticable. Thus, unless the problem can be treated empirically, or by approximation, recourse must be had to more elaborate computing equipment than the slide-rule and the desk calculating machine.

In the aircraft industry, for example, manual computation of various conditions governing stability, in particular the phenomenon known as "flutter," may be so protracted that the prototype may be "grounded" until the necessary calculations have been carried out. In this field the problems are usually formulated as sets of simultaneous differential equations, and the demand is therefore for a computer which will solve such equations quickly and present the results in a form convenient for observing the effects of changing the parameters of the system.

In other fields, particularly servo-mechanisms, a similar need exists, but it is not always possible to write down exact equations representing the behaviour of the system. The result is that equipment is often designed by trial and error methods; sometimes, as in the case of automatic pilots, under costly operating conditions. With this kind of problem the requirement is for a computer which will

simulate the transfer functions of the various parts of the equipment to enable the performance of the whole system to be studied. In addition, it is an advantage if actual physical components can be coupled up to the computer for simulation purposes.

Another important reason for the increasing use of computers is their ability to deal with problems involving non-linearities. Such problems are seldom amenable to calculation by straightforward methods and, in the past, have often been avoided or simplified unjustifiably by assuming linearity over a small range of conditions.

1.1. Choice of Computing Method

Computers capable of dealing with problems of the kinds mentioned fall mainly into one of two classes, according to whether they operate on an analogue or a digital basis.

The digital computer is particularly suitable for the solution of large-scale problems to a high degree of accuracy. The scope and accuracy of this kind of computer have been described in numerous reports during the last few years, and will not be discussed further in this paper.

The analogue computer, on the other hand, has a wide field of application, both to the solution of differential equations and for the investigation of dynamic system response, but with an accuracy usually of a few per cent. This limitation is not as serious as might be

supposed, since in many problems the input information is not known to any higher degree of accuracy. The analogue computer has, moreover, the ability to give a very good insight into the physical behaviour of the system. It can generally be constructed at a relatively low cost for the solution of every-day

problems. However some large installations have been constructed, of which TRIDAC¹³ is a typical example.

Another advantage of the analogue computer is the facility with which the time scale of events in a system may be altered, so that a time cycle may be speeded up or slowed down for more convenient observation. In a computer designed for repetitive operation this facility enables the effects of variations of the parameters of the system to be observed immediately.

Yet another feature of the analogue computer is that when operating on a "true" time scale it can be linked up with actual physical components of the system being investigated. This facility eases the task of the equipment designer in that the design can proceed in stages.

Although many forms of analogue computer have been constructed, based on mechanical and other techniques, attention in this paper will be confined to electronic instruments of the differential analyser type.

1.2. Principle of Operation of the Differential Analyser

The principle of the differential analyser type of computer may be understood by considering its application to a simple equation such as that which determines the current in the series resonant circuit of Fig. 1(a) when the initially charged capacitor is discharged by the closing of a switch:

$$L \frac{di}{dt} + R_1 i + \frac{1}{C_1} \int i dt = 0$$

If all the terms of this equation are represented by time-variable voltages, then it can be satisfied by a feedback system arranged as in Fig. 1(b).

This network is derived by assuming that di/dt exists at the point "b", producing i and $\int i dt$ by successive integration, and then adding these terms in the proportions indicated by the appropriate coefficients and with the correct signs. The various adding, integrating and sign reversing stages are based on high-gain negative-feedback amplifiers, coefficient values being controlled by the resistive scaling elements. (It is the *conductance* of the scaling

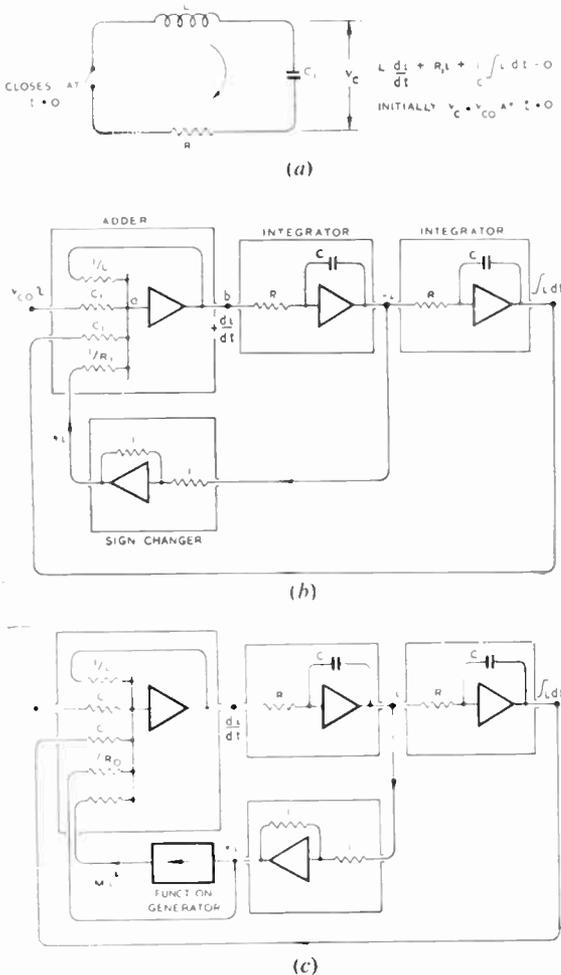


Fig. 1. Principle of operation of differential analyser type of computer.
 (a) Simple electrical circuit and corresponding equation.
 (b) Computing set-up for circuit with linear components.
 (c) Computing set-up for circuit in which the resistance varies according to some function of current i.e. $R_1 = (R_0 + M_i^k)$.

resistor which must be made proportional to the appropriate coefficient.)

That this network does, in fact, represent the equation may be demonstrated by assuming the adding amplifier to have infinite gain. Then for any finite value of the output of this amplifier the signal at its input (point "a") will always be zero. Since this input signal is proportional to the sum of the currents in the various scaling resistors, the required condition is satisfied.

The desired solution will be the variation with time of the dependent variable i , or one of its derivatives, subsequent to the closing of the switch. This is obtained by injecting a suitable voltage transient into the adder and observing the response at the appropriate point. In this case the "forcing function" is a voltage step which changes from a value representing the initial potential difference across the capacitor (v_{c0}) to zero at the start of the solution.

In the example given, both the integrator coefficients ($1/RC$) are assumed to be unity, but in practice it is often convenient to choose other values to suit the problem.

This technique may readily be extended to cover non-linear systems. For example, if the circuit illustrated represented a surge suppressor, the resistor might be a device whose resistance varied according to some function of the current passing through it, such as $R_1 = (R_0 + M_i^k)$. This condition could be satisfied by providing a function generator to supply the additional non-linear term to the adder, as shown in Fig. 1(c).

When solving simultaneous differential equations a computing channel is required for each variable in the set of equations. Feedback connections are then necessary, not only between the stages of each channel, but also between the different channels.

2. General Design Considerations

The approach to the design of an electronic computer depends very much on whether it is intended as a special-purpose instrument or for general application.

Although a general-purpose computer may

sometimes be less convenient than one designed for a specific problem, nevertheless its much wider scope makes it a far more useful tool. This versatility should also, with intelligent design, enable the cost of the non-specialized instrument to be kept below that of an equivalent special-purpose computer. The reason for this is that, in a limited market, only a general-purpose computer can hope to sell in the quantity necessary to enable the design costs to be spread over a sufficient number of units to make the price attractive.

The question of cost is, of course, one of the main considerations in the design of a commercial computer, and usually involves a compromise between performance and complexity. On this point it is considered that complication in mechanical or electrical design is justified only if it leads to easier handling, particularly by non-technical operators, a worth-while increase in accuracy, or greater reliability. Otherwise every effort should be made to keep the design and construction as simple as possible.

2.1. Size and Scope of Computer

Assuming that the computer is intended for general application, the first stage of development is obviously to decide on its size and capacity. A survey of the large number of problems which have been considered by the authors for treatment by analogue computation indicates that a computing installation may be called on to deal with anything from a single linear second-order equation to systems involving fifteen simultaneous non-linear equations of up to third order. Since the simpler problems are, naturally, more common, there is a strong case for making the basic equipment of comparatively limited capacity, say three second-order equations. Then, if the equipment is suitably designed, the more elaborate problems can be dealt with by linking a number of computers together.

The advantage of this arrangement is that, for a given outlay on computing equipment, a considerably greater volume of ordinary day-to-day work can be handled, while still providing for the occasional more complicated problem.

2.2. Computing Elements

For work on linear systems only a small number of different types of computing element is needed. The essential elements are those which perform the operations of integration, addition, and adjustment of scale. Diagrammatically an integrator may be represented as in Fig. 1 by a feedback system comprising a resistance, a capacitance and a high-gain amplifier. Similarly, addition is usually performed with a number of "summing" resistors in conjunction with an amplifier. Scale, or "coefficient," adjustment, involving either amplification or attenuation of a single quantity, can also be effected by an amplifier with resistive feedback, but it is usually more convenient to set the coefficients of an equation by adjusting the relative values of the summing resistors in the adder.

In designing a practical computer these facilities may either be provided as a number of single-purpose integrators, adders and scaling units, or as an equivalent number of general-purpose amplifiers which may be associated with various passive networks to perform the same operations.

Although, at first sight, the former arrangement may seem more convenient, it can easily lead to an uneconomic design where so many different kinds of computing unit have to be provided with each installation that on most problems there will always be a considerable number of idle units.

In addition to these basic elements certain other types are needed to deal with non-linear problems. It appears that these fall into three categories:

- (i) Multipliers, giving the instantaneous product of two time-variable inputs;
 - (ii) Function Generators, capable of giving an output that is any arbitrary single-valued function of the input;
- and
- (iii) Delay Elements, capable of shifting a time variable input along the time axis, without changing its waveform. This category also includes devices capable of altering the parameters of a problem at a pre-determined instant.

2.3. Ancillary Equipment

In addition to the computing units, various other facilities must be provided before a problem can be solved. Apart from power supplies, the most important of these are:

- (i) Means for interconnecting the computing units. In a general purpose computer it should be possible to make these connections easily and neatly to simplify the setting-up and checking procedures.
- (ii) Provision for starting and stopping the computation, and for introducing the appropriate initial conditions.
- (iii) Means for monitoring the performance of the computing units.
- (iv) Arrangements for displaying or recording the solution for comparison or measurement, including timing.

Although the provision of most of these facilities is usually a straightforward matter, the accurate recording of solutions often presents considerable difficulty. For most problems a well-designed computer may be expected to generate solutions which are accurate to within two or three per cent., but measurements on a wide variety of pen recorders show that it is not possible to transfer these solutions to paper without introducing further errors of from two to ten per cent., unless the frequencies concerned are less than one or two cycles per second. This limits the use of such recorders to solutions lasting several seconds.

Of the alternative methods that have been employed for high-speed solutions the "stroboscopic" or sampling method is only suitable for repetitive operation, and involves a large number of valves, while recording on magnetic tape is far from straightforward when dealing with this degree of accuracy.

For repetitive solutions, where the duration of each solution is less than, say, one second, photographic recording from a cathode-ray tube is probably the most convenient, since, with care, the errors introduced can be held within 1 per cent. The authors have found that a camera employing so-called "dry-working" material is a great convenience for this purpose, as positive prints are available within a minute of making the exposure.

3. Detailed Design Considerations

3.1. Amplifier Characteristics

If the computer is intended to deal with solutions lasting for several minutes in real time then the amplifiers must be capable of transmitting frequencies down to zero at the lower end of the spectrum. Although there are various carrier systems which will do this, a properly designed d.c. amplifier still has considerable advantages, and will be the only type considered in the following analysis.

Besides bandwidth there are a number of other important factors which influence the behaviour of a computing amplifier and, therefore, the performance of the computer as a whole. These will now be considered, both individually and in combination.

3.1.1. Gain

Ideally a computing amplifier should have infinite gain. In order to assess the effects of finite gain it is interesting to consider two basic computing networks: the adder and the integrator.

If the adding unit consists of an amplifier of open loop gain $-m$ with voltages v_1, v_2, \dots, v_n applied to input resistors R_1, R_2, \dots, R_n with a feedback resistor R_f , then it is shown in Section 9.1.1. (Fig. 17(a)) that the percentage error due to finite gain is:

$$\frac{100}{m} \left(1 + \frac{R_f}{R_1} + \frac{R_f}{R_2} + \dots \right)$$

As an example, the error in a unity-gain adder ($R_1=R_2=R_3 \dots =R_n=R_f$) with nine input resistors will not exceed 0.1 per cent. if m is greater than 10^4 .

In the case of multiplication by a constant the percentage error becomes

$$\frac{100}{m} \left(1 + \frac{R_f}{R_1} \right)$$

where R_1 is the input resistor.

If the amplifier is connected as an integrator, Section 9.1.2 (Fig. 17(b)), then the percentage error due to finite gain is given by the expression

$$100. L^{-1} \left\{ V'_0 \left[\frac{1}{m} + \frac{1}{pmRC} \right] \right\} / V_0$$

The exact nature of the errors due to

inadequate gain will, therefore, depend on the waveform applied to the integrator. Two particular inputs are considered in Section 9.1.2: the step function and the sine wave.

Figure 18 shows the relationship between percentage error and gain for various ratios of t/RC , i.e. the ratio of computing period to the physical time constant, for a step input. From these curves it can be seen, for example, that the error is 0.5 per cent. for a gain of 10^4 and a computing period of 100 seconds if the time constant is 1 second.

In the case of a sinusoidal input, lack of gain introduces three forms of error:

(i) Reduction of amplitude by a constant factor

$$\frac{\omega mk}{(1+m)\sqrt{(1+\omega^2 k^2)}} \text{ where } k=RC(1+m)$$

(ii) A further exponential change in amplitude by the time-dependent factor

$$\left[\frac{\omega k}{(1+\omega^2 k^2)} \right] e^{-t/k}$$

and

(iii) A (leading) phase shift of $\tan^{-1}1/\omega k$.

If m is again taken as 10^4 and the time constant as one second, then if error (i) is not to exceed 0.5 per cent. the periodic time of the signal must be less than 6.6×10^3 seconds. For the same conditions however, error (ii) limits the duration of the computing period to less than 50 seconds. Similarly, if the phase error is not to exceed 0.01 deg., then the periodic time, T , of the signal must not be greater than about 10 seconds (Fig. 19).

From these considerations it will be apparent that quite a moderate amplifier gain will enable a computer to deal with problems lasting up to about a minute in real time.

3.1.2. Bandwidth

The open-loop bandwidth of a computing amplifier is of significance because it affects the stability and phase shift under feedback conditions.¹

Provided there are not too many inherent phase shifts in the amplifier it is comparatively easy to introduce phase correcting networks which will ensure stability with any particular

feedback network, e.g. as an integrator or as an adder, and with any values of feedback components. It becomes more difficult, though, if the amplifier must remain stable in any kind of feedback circuit with all possible component values.

It is not generally appreciated, however, just how high the cut-off frequency must be for accurate computation, particularly with lightly damped oscillatory solutions such as occur in aircraft flutter problems. In Section 9.2 it is shown that the set-up for a single second-order equation (Fig. 20) will introduce a spurious logarithmic increment of $(\omega_n^2/2\omega_r)t$, where ω_n is the solution pulsance and ω_r is the cut-off pulsance of the adder; assuming the integrating amplifiers to be perfect and that, over the computing periods considered, inaccuracies due to finite open-loop amplifier gain are negligible, the error being due solely to the time constant of the adder. This means, for example, that a computer generating a sinusoidal oscillation with a real frequency of only 1 c/s would need to have amplifiers with a cut-off frequency above 15 kc/s under operating feedback conditions if the solution had to be within 1 per cent. of correct amplitude after 50 seconds. Alternatively, only 7 cycles of an oscillation could be displayed to this accuracy with repetitive operation and a sweep duration of 1 second. It will be appreciated that errors due to finite amplifier gain, discussed in the previous sub-section, may, in fact, be significant and should be taken into account when assessing the overall error.

3.1.3. Drift

Drift in a d.c. amplifier may be defined as a spurious signal causing an output to appear when no intentional signal is applied to the input terminals. It is therefore convenient to regard the drift as arising from a fictitious generator in series with the input of the amplifier. Drift arises from a variety of sources, and can be reduced both by careful design and by various drift-correcting systems.

As an example of the relative importance of this effect it may easily be shown that, if the error superimposed on a 20-volt signal at the

output of an integrator is to be less than 0.5 per cent., then for $t/RC=100$, the input drift must not be greater than 1 mV.

3.1.4. Grid current

Although the effect of the residual drift in an amplifier may be reduced considerably by the use of a suitable drift correction system, a satisfactory method has not yet been devised for the reduction of drift due to the grid current flowing in the amplifier input circuit.

A detailed analysis of the combined effects of grid current and drift in an integrator is given in Section 9.4. This shows that at the beginning of the computing period a grid current i_g produces a spurious output which increases at a rate given by i_g/C . In order to satisfy the same conditions as in 3.1.3, i_g must be less than 10^{-9} amperes if R is 1 megohm.

3.1.5. General comments on amplifiers

It is difficult to assess the relative importance of the various errors introduced by the amplifiers when these occur in combination. If computation must be done on a real time scale, then solutions occupying anything from a few seconds to several minutes, or even hours, may be involved. Emphasis will then be on amplifier gain, at the expense of bandwidth, and on reducing drift and grid current to a minimum. Alternatively, if there is no restriction on the time scale, then it is often an advantage to work repetitively with solution times between 10 and 1,000 milliseconds. In this case the limit will probably be set by amplifier bandwidth.

The real criterion of amplifier performance appears therefore to be the gain-bandwidth product.

3.2. Network Characteristics

3.2.1. Resistors

The choice of type of resistor for use in adding and feedback networks depends partly on whether the computer is to be used for "single-shot" or repetitive solutions. If the former, then the computing period is unlikely to be less than one second, and the significant frequencies in the solution will be comparatively low. For this kind of operation the helical, wire-wound, type of potentiometer is very

convenient. But it must be remembered that, although a potentiometer of this sort may be set to give ratios correct to one part in several hundred, the absolute value of resistance may vary by several per cent. between one unit and another for the same dial setting.

Other disadvantages of potentiometer scaling are the loading imposed on the computing amplifiers and the variations of impedance introduced into the scaling circuits. Also, for repetitive solutions the stray reactance of helical potentiometers is likely to cause too much phase shift to be acceptable, and the only practical alternative seems to be the use of high-stability fixed-value resistors in switched decades.

3.2.2. Capacitors

For satisfactory operation the capacitors in the computing networks must be both stable and low-loss. Of the various alternative types, paper capacitors suffer from excessive dielectric absorption, while mica capacitors are expensive and bulky. The most suitable type of component is probably the polystyrene capacitor, which combines adequate accuracy and stability with low losses.

Unfortunately, it is not possible to obtain polystyrene capacitors commercially to a closer tolerance than ± 1 per cent. However, if economic considerations may be ignored then closer values of capacitance may be achieved by selection or "padding."

3.2.3. Phase shift

Apart from the characteristics of the components themselves, their actual disposition in the computing networks requires very careful consideration if the stray reactances are not to introduce phase shifts into the feedback loops.

In Section 9.3 it is shown that the number of cycles, N , of an oscillatory solution that may be observed before the amplitude error exceeds 1 per cent. is given by

$$N = 0.181 / \zeta$$

for a simple second-order system where ζ is the phase shift in degrees round the loop. As an indication of the severity of this requirement, if ζ is only *one minute of arc* then no more than 11 cycles of the solution may be displayed for this accuracy.

3.2.4. General comments on networks

From Section 3.2.2 it follows that the limit to computing accuracy with most problems is set by the network capacitors, particularly in the integrators. Unless tolerances of better than ± 1 per cent. can be achieved with these components there is little point in trying to improve resistor accuracy and stability. However, if it were practicable to achieve, say, ± 0.2 per cent. for the capacitors, then it would appear to be possible with correspondingly accurate resistors, such as metallic film types, and good amplifiers, to obtain accuracies for individual computing stages considerably better than the nominal 1 per cent. which is often specified for commercial instruments.

Alternatively, and at the expense of versatility, it is possible to accept capacitors of nominal value and "pad" the associated resistors in order to obtain specified time constants.

4. Description of Computer

The design considerations outlined in the preceding Sections were investigated during the development of a commercial general-purpose analogue computer, and before giving some examples of the applications of such an instrument a brief description of the equipment may be helpful.

4.1. General

The appearance of the computer is shown in Fig. 2. It is built in the form of a double-pedestal desk surmounted by a rack into which various computing units may be inserted and secured by quick-release fasteners. Electrical connections are made automatically, and all units of the same type are interchangeable.

For linear problems only two kinds of computing unit are required—described as Linear Function Units and Scaling Units—and space is available for 18 of each of these. For non-linear problems several additional kinds of computing unit are also available, the most important being the Multiplier, the Continuous Function Unit and the Discontinuous Function Unit. Thus a console can be filled with whatever combination of computing units may be needed for a particular problem, and if one

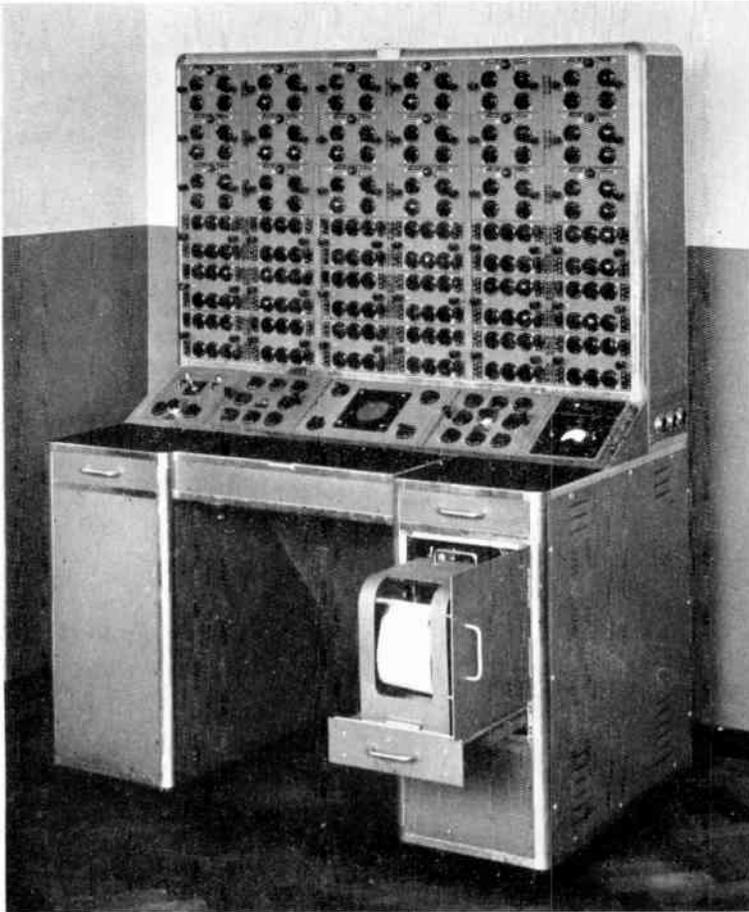


Fig. 2. General-purpose Computer.

console will not carry enough units then two or more consoles may be coupled together without any loss of performance.

The inclined panel at the base of the rack carries a double-beam cathode-ray tube, together with its associated deflection amplifiers and timing unit, various calibrating and metering devices, and the main power controls.

The pedestals of the console contain the stabilized h.t. and l.t. power units, a pen recorder and racks for spare computing units.

Computing units are interconnected for each problem by inserting small bridging plugs into a Set-up Panel which is recessed into the top of the desk and covered by hinged flaps when access is not required. This device avoids the

use of plugs and cords, and makes the process of checking connections a very simple matter. It also enables a card to be punched with holes corresponding to the plugs needed for any particular problem. This card may then be filed for use whenever the same kind of problem has to be set up again.

4.2. Presentation

The computer may be used either repetitively or for single-shot solutions. In the former case the computing period may be set to one of a number of values between 1/30th second and 1 second, and the solution displayed on the screen of the cathode-ray tube. In the latter case computation is generally started and stopped manually, and the solution obtained on the pen recorder.

The normal forcing function is a voltage step whose amplitude may be adjusted on the control panel.

Alternatively, other functions may be generated internally or fed in from an external source.

For repetitive solutions the lower spot on the screen is usually arranged to display self-generated timing marks. However, it may also be used to display any other quantity occurring in the system under investigation, or two different quantities may be applied to the x and y axes of the cathode-ray tube for phase measurement.

At the end of each computing cycle the system is restored to the quiescent condition by means of relays.

4.3. Computing Units

4.3.1. Linear Function Unit

This is the basic computing element. It

consists of a drift-corrected d.c. amplifier associated with a resistance-capacitance network which can be set to perform a variety of operations, including addition and integration,

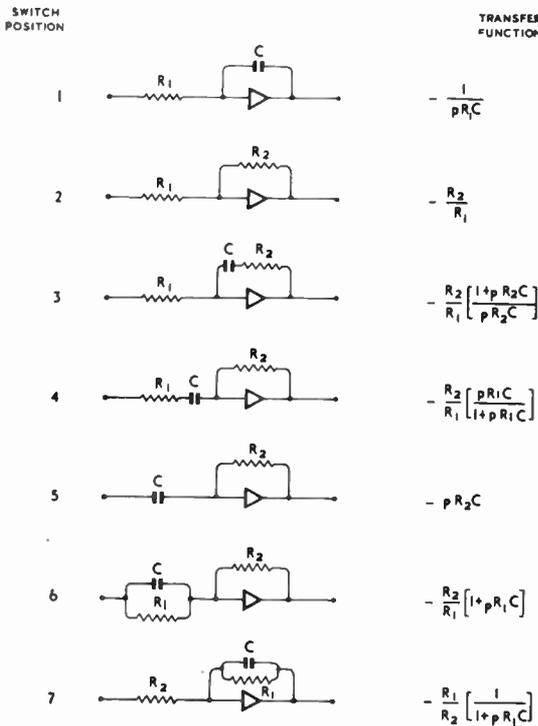


Fig. 3. Alternative arrangements of Linear Function Unit.

as shown in Fig. 3. It also provides facilities for introducing "initial conditions" into a problem.

The amplifier has an open-loop voltage gain of about 12×10^3 and is phase corrected to give a cut-off frequency higher than 15 kc/s under all normal conditions of resistive feedback. The effective input grid current is less than 5×10^{-11} amperes, while the residual input drift voltage is of the order of 1mV without drift correction.

4.3.2. Scaling Unit

This unit is employed to effect scale changing and summation in the interconnections between the other units. It consists of two identical sections assembled on one panel for convenience and standardization, each section

consisting of a 3-decade switched resistance combined with a decade attenuator. The switches of each section may be set to any number between 1.00 and 10.99, and the figure obtained may then be multiplied by a factor of 0.1, 1.0, 10 or 100 with the attenuator.

4.3.3. Multiplier

The Multiplier operates on the principle of "quarter squares," and if signals of magnitudes x and y are applied respectively to the two input terminals, then the output will be proportional to the instantaneous product xy .

Each unit consists of a number of drift-corrected d.c. amplifiers associated with "squaring" circuits of selenium diodes and precision resistors, and the overall accuracy at maximum signal level is within ± 1 per cent. A block diagram of the Multiplier is given in Fig. 4.

4.3.4. Continuous Function Unit

This unit, sometimes referred to as a Function Generator,^{2,3,4} is a device which enables any smooth single-valued function, whether expressed mathematically or as a graph, to be approximated by a series of 21 straight lines. The slopes and intersections of these lines are adjusted manually by a series of potentiometers when the function is set up, the operation being simplified by the provision of an accurate calibrating circuit. The accuracy obtained depends on the shape of the curve to be generated. As an example, a parabola or one cycle of a sine wave can be reproduced to within 1 per cent. It is not necessary for the curve to pass through the origin, but the accuracy of reproduction is likely to fall off if the slope of the curve exceeds about 85 deg. The principle of operation of this unit may be followed from the block diagram of Fig. 5. Each section of the curve is produced by an element consisting of a pair of selenium diodes associated with a biasing potentiometer and an attenuator, and the whole group of elements is isolated from the input and output terminals by buffer amplifiers.

Both the Continuous Function Unit and the Multiplier have cut-off frequencies comparable with that of the Linear Function Unit, and may therefore be used for repetitive operation.

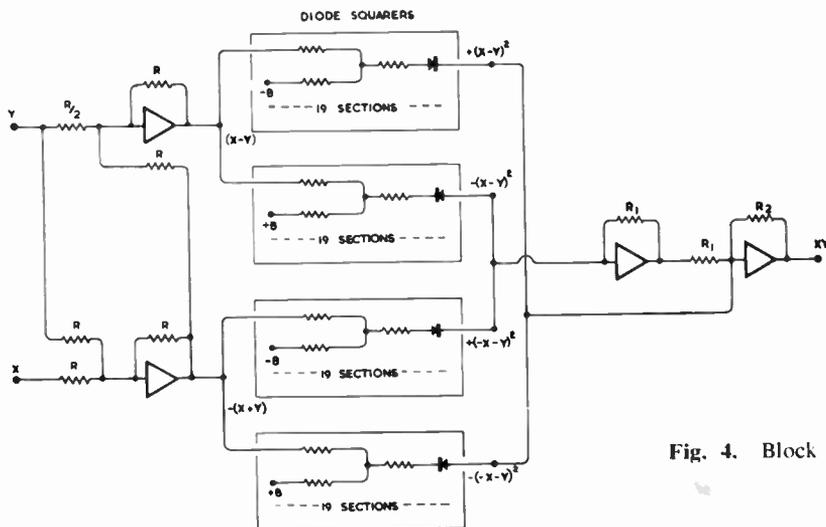


Fig. 4. Block diagram of Multiplier.

4.3.5. Discontinuous Function Unit

The purpose of this unit is to supplement the Continuous Function Unit in cases involving abrupt discontinuities. Its transfer characteristic can be switched to represent either "limiting" or "dead zone" conditions, the positions of the discontinuities and the slope being independently adjustable (Fig. 6). In conjunction with other computing units the Discontinuous Function Unit can also be used to represent hysteresis or backlash (Fig. 16).

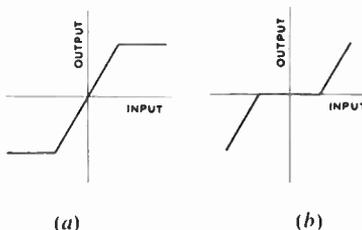


Fig. 6. Transfer characteristic of Discontinuous Function Unit set for: (a) Limiting; (b) "Dead Zone."

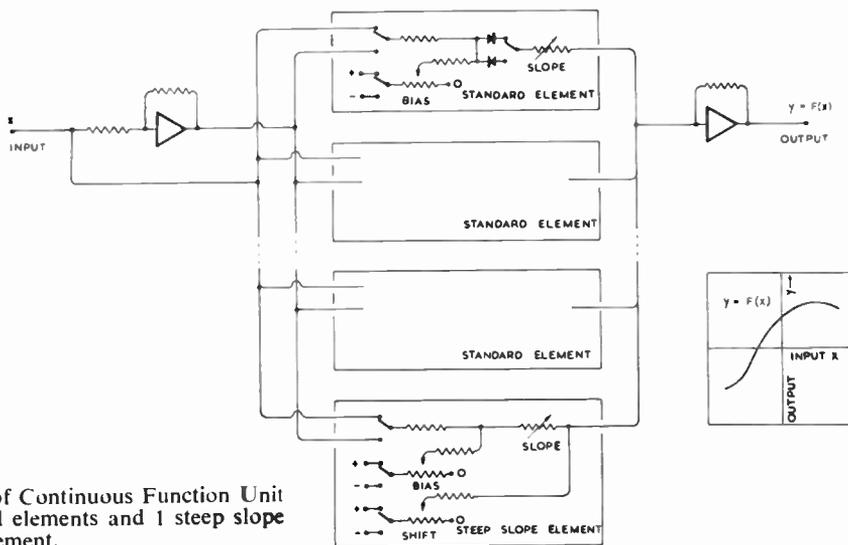


Fig. 5. Block diagram of Continuous Function Unit consisting of 20 standard elements and 1 steep slope element.

4.3.6. Other units

The computing units described above are those in common use. In addition, a number of other units have been developed for certain special types of problem. These include programme units, capable of altering the constants of the system at either a pre-determined instant of time or when one of the variables reaches a certain value, and time delay devices.

5. Applications

An analogue computer of the type described in the previous Section has a very wide range of application, as may be appreciated by reference to the extensive literature on the subject.⁵ In this paper, however, it is not intended to present an exhaustive catalogue of problems and their solutions, but simply to discuss some of the more interesting applications encountered by the authors during the last few years.

In presenting these examples it is convenient to group them according to the way the computer is used. Early instruments of this kind were usually described as differential analysers and, as the description implies, were intended primarily for the solution of differential equations. Many of the problems presented to a general-purpose computer still originate in the form of equations, particularly in the field of aircraft design.

However, many others concern dynamic systems whose behaviour and constants may be known but are not expressed in equation form. These may be classed generally as simulation, a term usually understood to imply the direct representation of the transfer functions of the elements of a system. Such simulation is often carried out in association with some of the actual physical components of the system, and usually in real time. In some cases of simulation a great saving in effort, and often computing facilities, may be achieved by what has been called the "Direct Analogue" approach, in which the equivalent network is established by a logical analysis of the components of the system without writing down any equations. This approach is particularly valuable in bi-directional systems such as coupled electric circuits.

5.2. Solution of Differential Equations

5.2.1. Mathematical analysis

In the realm of pure mathematical analysis an analogue computer can often save considerable time by indicating the general nature of the solutions to non-linear equations. Since a computer is most useful for problems which cannot be solved easily or accurately on paper, it is difficult to find a suitable example for which both computed and calculated results are available for comparison. However, as a simple example the equation:

$$dy/dt = -K(y - y^2) \sin Kt$$

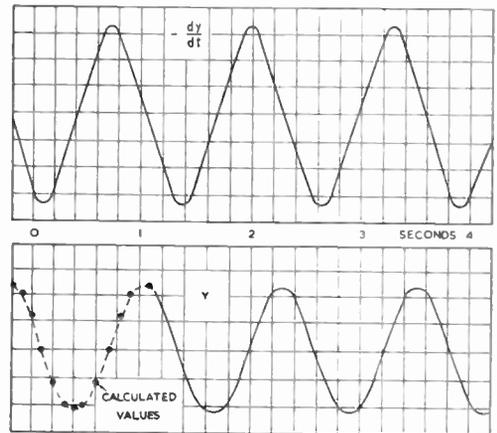
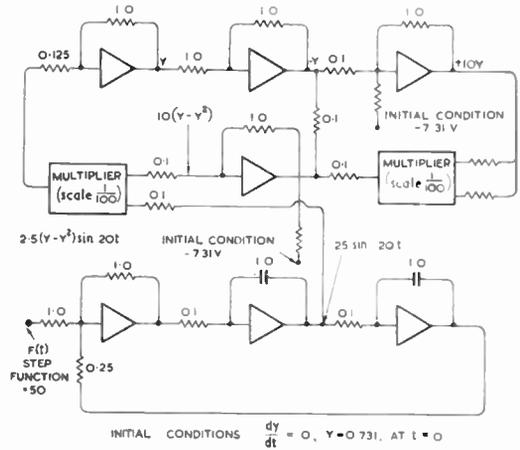
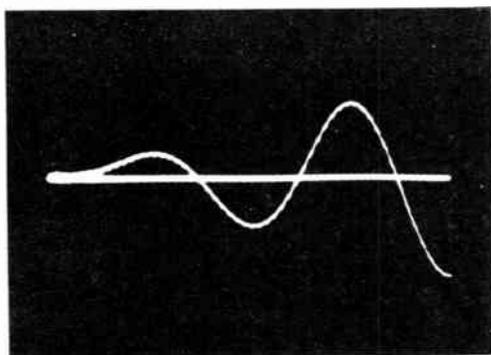
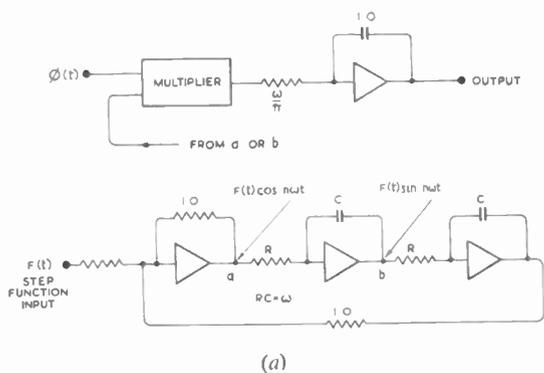
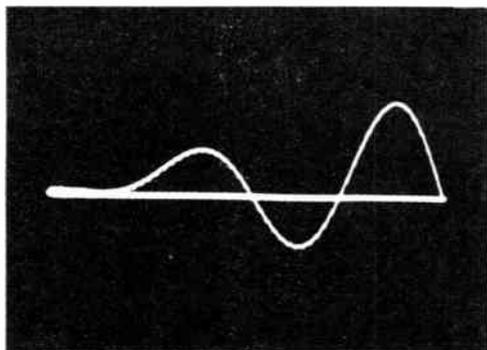


Fig. 7. Solution of equation $dy/dt = -K(y - y^2) \sin Kt$ Showing computer arrangements and recordings of y and dy/dt .



(b)



(c)

Fig. 8. Harmonic analysis of saw-tooth waveform.
 (a) Computer arrangement.
 (b) Photograph of $\omega/\pi \int_0^T \varphi(t) \cdot \sin 2\omega t \cdot dt$ display.
 (c) Photograph of $\omega/\pi \int_0^T \varphi(t) \cdot \cos 2\omega t \cdot dt$ display.

can be shown by analysis to have the solution :

$$y = 1/[e^{-\cos kt} + 1]$$

This equation was solved with a computer arranged as in Fig. 7 which also shows recordings of the variations of y and dy/dt with time. The calculated response for y , shown for comparison, agrees with the computed solution to within a few per cent.

Another, little known, application is to harmonic analysis. Since the Fourier expansion of a periodic function of time $\varphi(t)$ is the series:

$$a_0 + a_1 \sin \omega t + a_2 \sin 2\omega t + \dots + a_n \sin n\omega t + b_1 \cos \omega t + b_2 \cos 2\omega t + \dots + b_n \cos n\omega t$$

where

$$a_0 = \frac{1}{2\pi} \int_0^{2\pi} \varphi(t) \cdot d\omega t$$

$$a_n = \frac{1}{\pi} \int_0^{2\pi} \varphi(t) \sin n\omega t \cdot d\omega t$$

and

$$b_n = \frac{1}{\pi} \int_0^{2\pi} \varphi(t) \cos n\omega t \cdot d\omega t$$

the amplitudes of the various harmonics may be obtained by evaluating the corresponding integrals.

To enable the integration to be carried out with respect to t instead of ωt , a_n and b_n may be re-written:

$$a_n = \frac{\omega}{\pi} \int_0^T \varphi(t) \sin n\omega t \cdot dt$$

$$b_n = \frac{\omega}{\pi} \int_0^T \varphi(t) \cos n\omega t \cdot dt$$

where T is the repetition period of $\varphi(t)$. If T is made equal to the period of computation, then the required coefficient is the integral of the product of $\varphi(t)$ and either $\sin n\omega t$ or $\cos n\omega t$ over one cycle.

Since $\varphi(t)$ is often available in electrical form, it is usually only necessary to set up a second-order computing network to generate $\sin n\omega t$ and $\cos n\omega t$ and then arrange the computer to form the product and integral.

Figure 8 shows the computer set-up used, together with photographs of two waveforms, for a saw-tooth wave. The following table compares the measured values of the coefficients with the calculated ones.

Table 1

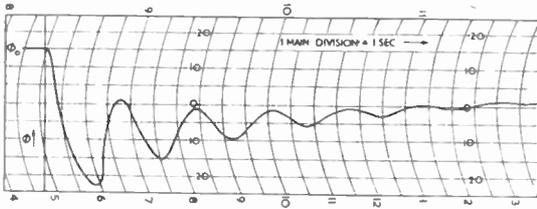
Comparison between Calculated Coefficients of Fourier Expansion and Values obtained from Computer, for a Saw-tooth Waveform

Order of Harmonic <i>n</i>	From Computer		By Calculation	
	<i>a_n</i>	<i>b_n</i>	<i>a_n</i>	<i>b_n</i>
1	2.57	<0.001	2.54	0
2	1.28	..	1.27	0
3	0.875	..	0.846	0
4	0.665	..	0.635	0
5	0.535	..	0.508	0

Although the accuracy of this method tends to diminish as the order of the harmonic increases, it is quite satisfactory up to *n*=5, at least.

5.2.2. Aircraft dynamic response

During the design of an aircraft it is necessary to estimate its dynamic stability under various conditions of flight. A typical example of this



$$\frac{dx}{dt} - y_r x + \mu_2 z = \mu_2 k q = 0$$

$$-\frac{l_y}{i_A} x + \frac{dy}{dt} - \frac{l_p}{i_A} y - \frac{E}{A} \frac{dz}{dt} - \frac{l_r}{i_A} z = 0$$

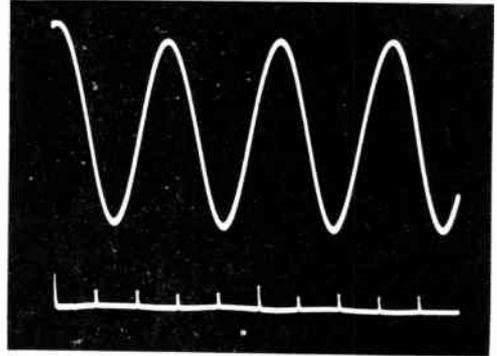
$$-\frac{n_y}{i_C} x - \frac{E}{C} \frac{dy}{dt} - \frac{n_p}{i_C} y + \frac{dz}{dt} - \frac{n_r}{i_C} z = 0$$

$$-y + \frac{dq}{dt} = 0$$

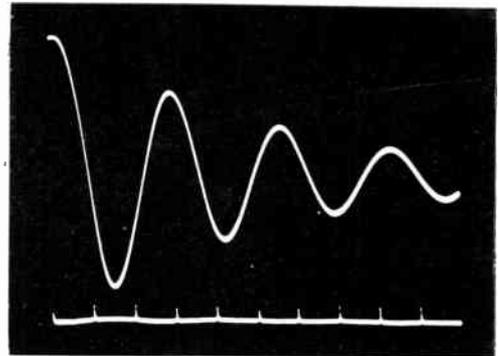
where $x \equiv \mu_2 v / V$
 $y \equiv \mu_2 p b / 2V$
 $z \equiv \mu_2 r b / 2V$

and v = sideslip velocity.
 q = angle of bank.
 p = rate of roll.
 r = rate of yaw.

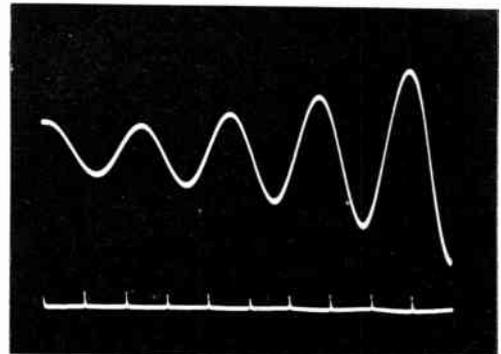
Fig. 9. Aircraft lateral stability problem, showing variation of bank angle after sudden (step) disturbance.



(a) System neutrally stable.



(b) System stable.



(c) System unstable.

Fig. 10. Appearance of trace on computer screen for aircraft ternary flutter problem.

type of problem is illustrated in Fig. 9 and concerns the response of the aircraft to a disturbance about the roll (bank) axis. From the

equations it will be seen that three independent variables are involved, and it is assumed in this case that the coefficients are linear.

The solution of this problem is of interest in that it contains a rapid, fairly heavily damped, oscillation superimposed on a more slowly changing component which is divergent. This, incidentally, makes it of particular use as a test problem, as the low-frequency component is very sensitive to the effects of drift and grid current in the computer.

As an indication of accuracy, the frequency of the damped oscillation was calculated for one set of parameters to be 0.62 c/s with a damping factor of 0.36, while figures of 0.61 c/s and 0.33 were measured from the recording.

5.2.3. Aircraft flutter.⁶

“Flutter” is the term used to describe an unstable oscillation induced by the interaction between inertia and aerodynamic forces in an aircraft—the aim of the designer is to ensure that this dangerous condition cannot occur over the normal range of conditions. Usually the problem is to determine the position of the boundary between the regions of stability and instability in terms of the critical air speeds corresponding to given values of mass balance of a control surface. The solution is therefore a curve plotted from the dial settings of the computer instead of a response on the screen or pen recorder.

A flutter problem consists essentially of several simultaneous second-order differential equations, according to the number of degrees of freedom involved, and the computing set-up therefore contains a similar number of cross-coupled channels. A ternary (three degrees of freedom) problem is quite common, while the most elaborate investigated so far has been a senary system involving three computer consoles linked together.

A typical ternary problem is illustrated in Figs. 10 and 11 and it will be seen that in this particular example the largest discrepancies occurred in the region of the “nose” of the curve, and amounted to some 3 per cent. of the corresponding value of mass balance. As a matter of interest, the frequency of oscillation of the system for 20 per cent. mass balance was measured as 9.6 c/s, and subsequently calculated to be 9.8 c/s.

Many of the terms in these equations contain powers of the forward velocity, and it

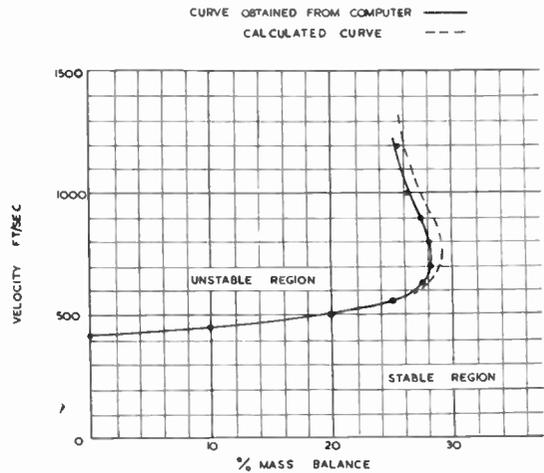


Fig. 11. Equations for aircraft ternary flutter, with typical solution.

$$\begin{aligned}
 a_{11} \cdot \frac{d^2x_1}{dt^2} + b_{11} \cdot \frac{dx_1}{dt} + c_{11}x_1 + a_{12} \cdot \frac{d^2x_2}{dt^2} + b_{12} \cdot \frac{dx_2}{dt} + c_{12}x_2 + a_{13} \cdot \frac{d^2x_3}{dt^2} + b_{13} \cdot \frac{dx_3}{dt} + c_{13}x_3 &= 0 \\
 a_{21} \cdot \frac{d^2x_1}{dt^2} + b_{21} \cdot \frac{dx_1}{dt} + c_{21}x_1 + a_{22} \cdot \frac{d^2x_2}{dt^2} + b_{22} \cdot \frac{dx_2}{dt} + c_{22}x_2 + a_{23} \cdot \frac{d^2x_3}{dt^2} + b_{23} \cdot \frac{dx_3}{dt} + c_{23}x_3 &= 0 \\
 a_{31} \cdot \frac{d^2x_1}{dt^2} + b_{31} \cdot \frac{dx_1}{dt} + c_{31}x_1 + a_{32} \cdot \frac{d^2x_2}{dt^2} + b_{32} \cdot \frac{dx_2}{dt} + c_{32}x_2 + a_{33} \cdot \frac{d^2x_3}{dt^2} + b_{33} \cdot \frac{dx_3}{dt} + c_{33}x_3 &= 0
 \end{aligned}$$

is thus rather laborious to make the necessary manual adjustments to the computer dials when exploring a range of velocities on a big problem. In order to reduce the computing time still further on this kind of problem an alternative version of the Linear Function Unit has been designed, with the switches driven by rotary solenoids. This enables all the velocity controls to follow automatically the setting of whichever unit is chosen as the master.

5.2.4. Non-linear circuits

The circuit chosen in Section 1.3 to illustrate the application of an analogue computer to non-linear electrical networks was actually encountered as a surge suppressor. In this circuit, Fig. 1(c), the total resistance was made up partly by a fixed resistor and partly by a component having a voltage-current relationship $v_m = 430 i^{0.24}$ where i is the instantaneous value of current in the circuit and v_m the voltage across the component.

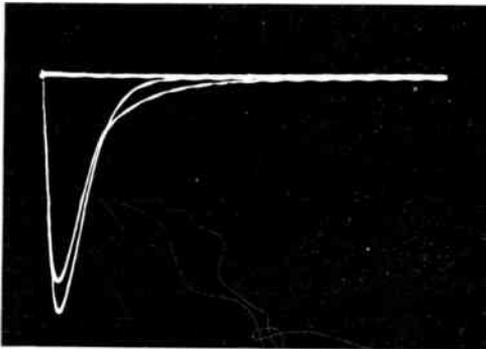


Fig. 12. Variation of current with time for circuit shown in Fig. 1, showing reduction of peak value by addition of non-linear resistance.

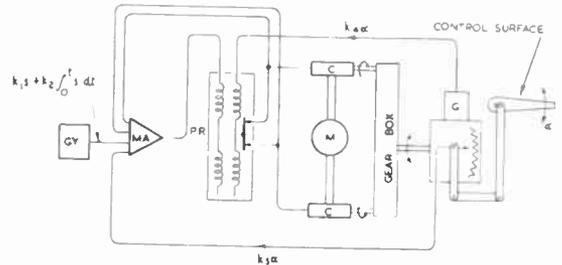
The information required in this case was the variation in current after the switch was closed, both with and without the non-linear element in circuit, and both curves are shown in Fig. 12.

5.3. Simulation

5.3.1. Auto-pilot response

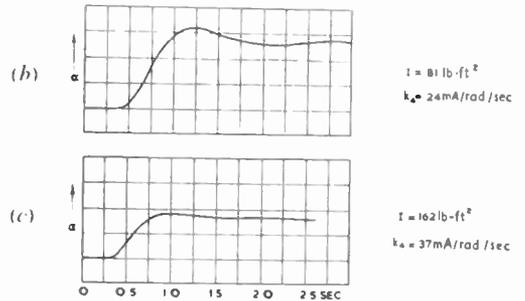
One of the biggest applications of the analogue computer is in the field of servo-mechanisms.^{7,8} Aircraft automatic pilots, in

particular, are a good example of equipment on which a great deal of risk and development time can be avoided by simulation instead of trial and error under actual operating conditions.



MA	Magnetic Amplifier	M	Motor
GY	Gyroscope	C	Clutch
PR	Polarized Relay	G	Tacho-generator

(a) Block diagram.



(b) Response with load inertia of 81 lb/ft.²
 (c) Response with load inertia of 162 lb/ft.²

Fig. 13. Simulation of aircraft autopilot,

The example illustrated concerns the application of an existing electric automatic pilot to a different type of aircraft. In this case two points were in doubt: (i) whether the system would be stable with the reversing clutch used to couple the servo-motor to the output gearbox, and (ii) whether its response, if stable, would be satisfactory under the various inertia loads due to the aircraft control surfaces.

Figure 13(a) shows a block diagram of the auto-pilot. It consists essentially of an integrating rate gyroscope feeding a position-control servo. The servo amplifier operates a relay which, according to the polarity of the

output, energizes electromagnetic clutches controlling the sense of the torque transmitted from a continuously running motor to the output gearbox. The positional feedback is provided by a potentiometer on the output shaft of the gearbox, while velocity feedback is also supplied by a tacho-generator on its input shaft. In addition an arrangement is provided for feeding an opposing signal into the amplifier when a relay closes, so producing a continuous slight hunting of the system.

In the simulation the actual relays of the system were used to ensure correctness of representation, but the remainder of the autopilot was set up in analogue form.

When the original values of the constants k_1, k_2, k_3 and k_4 were set in, the response to a step voltage applied at the output of the gyro was very sluggish, with a rise time of nearly 2 seconds. In order to achieve a sufficiently fast response with adequate damping it was found necessary to reduce the velocity feedback by a factor of about 5. Figs. 13(b) and (c) show the final responses with load inertias of 81 and 162 lb/ft² respectively. There was no tendency to instability, apart from the intentional "dither" which was too small to show on the recordings.

5.3.2. Coupled mechanical systems

Problems involving discontinuities in mechanical systems are encountered in fields as varied as those of marine power transmission, valve design for internal combustion engines and machine tools. Fig. 14(a) shows the basic system involved in a typical problem of this kind. The machine part of mass m_1 is subjected to a reciprocating force $F_1 = a \sin \omega t$ and is initially just in contact with another block of mass m_2 which is restrained by a spring of stiffness s . Information is required on the motion of the two parts, for a given set of constants, when the mass m_1 is suddenly separated from m_2 . The motion is assumed to take place in the horizontal plane, in order to neglect the effect of gravity.

Let V_1 and V_2 be the velocities of the two parts; and F_2 and F_3 the forces, respectively, between them and on the spring. Then while

they are in contact with one another V_1 must equal V_2 , or

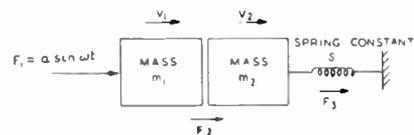
$$\frac{1}{m_1} \int_0^t (F_1 - F_2) dt = \frac{1}{m_2} \int_0^t (F_2 - F_3) dt$$

which leads to the condition

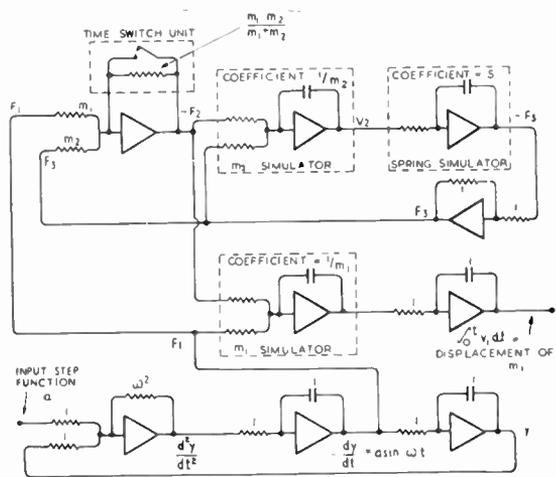
$$F_2 = \frac{m_2}{m_1 + m_2} F_1 + \frac{m_1}{m_1 + m_2} F_3$$

before separation. After separation, of course, $F_2 = 0$.

Having established these initial conditions it is unnecessary to write down the equations of motion, since the computer arrangement may be arrived at by the direct analogue approach mentioned in Section 5.1.



(a) Basic system.



(b) Computer set-up.

Fig. 14. Coupled mechanical system.

The actual set-up is shown in Fig. 14(b) and is almost self-explanatory. It is only necessary to mention that the separation of the two masses is simulated by a programme unit which reduces F_2 to zero at some instant during the display period, while the applied force $a \sin \omega t$

is obtained by applying a step of magnitude a to a separate chain arranged to solve the equation

$$\frac{d^2y}{dt^2} + \omega^2y = -\omega^2a$$

from which $y = a(\cos \omega t - 1)$
 or $-dy/dt = a \sin \omega t$

The results of this simulation for $m_1 = 3m_2$ and $\omega = 10$ rad/sec are shown in Fig. 14(c).

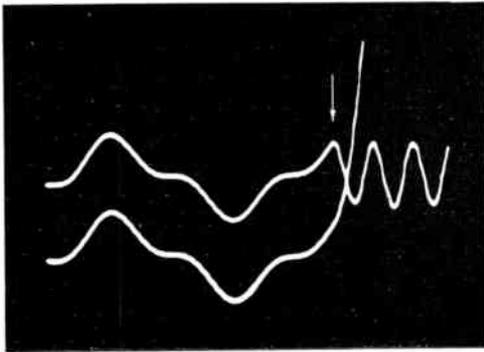


Fig. 14. (c) Recordings of displacements of m_1 , upper trace) and m_2 , showing instant of separation (arrowed).

5.3.3. Economic models

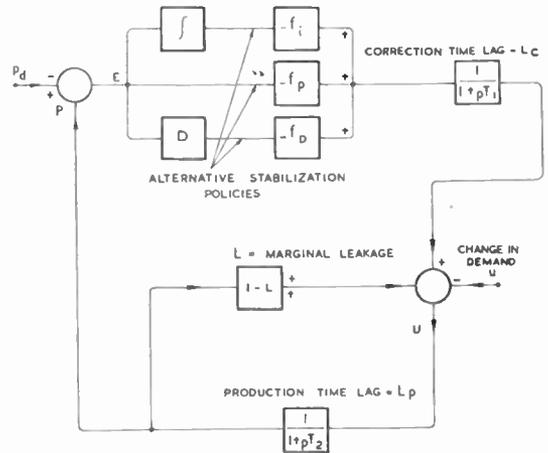
A comparatively new application of the analogue computer is its use as a “model” for the investigation of economic problems. Such a model, consisting of various combinations of standard computing elements, can be constructed to represent the hypothetical structure of an actual economic system.^{9,10} The response of the model to certain disturbances can then be compared with that of the real system, both to check the validity of the analogue and to predict the future behaviour of the system. The model can then be used to investigate the effects of various stabilization policies on such factors as production or prices.

Figure 15(a) shows a model which indicates in simplified form the interaction between production and demand for various stabilization policies. In this diagram production P is supposed to follow demand U , after a production time lag L_p , and in turn to influence

demand through the “marginal propensity to spend” $(1 - L)$, where L is the “marginal leakage.” This main loop is supplemented by one representing stabilization policy, which attempts to maintain a constant level of production by taking appropriate correcting action, such as altering taxation or credit. According to the policy adopted, this correction may be proportional alternatively to production, its rate of change, its integral or some combination of these factors, and is applied through a lag L_c to modify the total demand for production.

The proportionality coefficients are denoted by f_p , f_d and f_i respectively.

For this problem the computer was set up as in Fig. 15(b) to display a period of 10 years on a time scale of 1 second = 1 year, and the response of the system recorded for a step function change in demand for production of magnitude



- P_d Desired production
- P Actual production
- E Error in production
- \int Integration with respect to time
- D Differentiation with respect to time
- f_i Integral correction factor
- f_p Proportional correction factor
- f_d Derivative correction factor
- u Change in demand
- U Demanded production

Fig. 15. Economic system basic “model.”

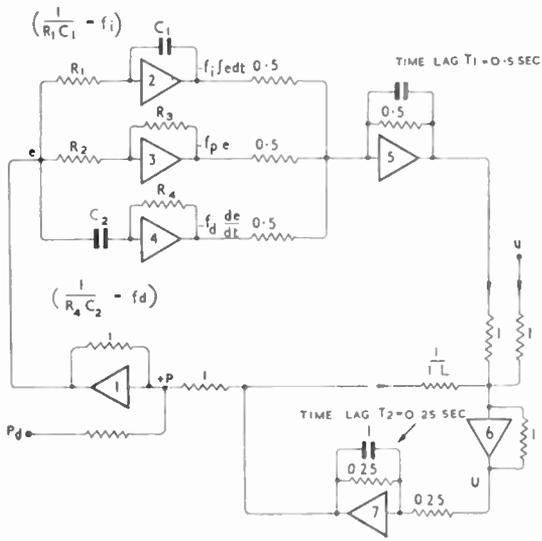


Fig. 15. (b) (above) Computer set-up.

(c)-(f) (right) Fluctuation in production for various stabilization policies. Computing time scale: 1 second=1 year.

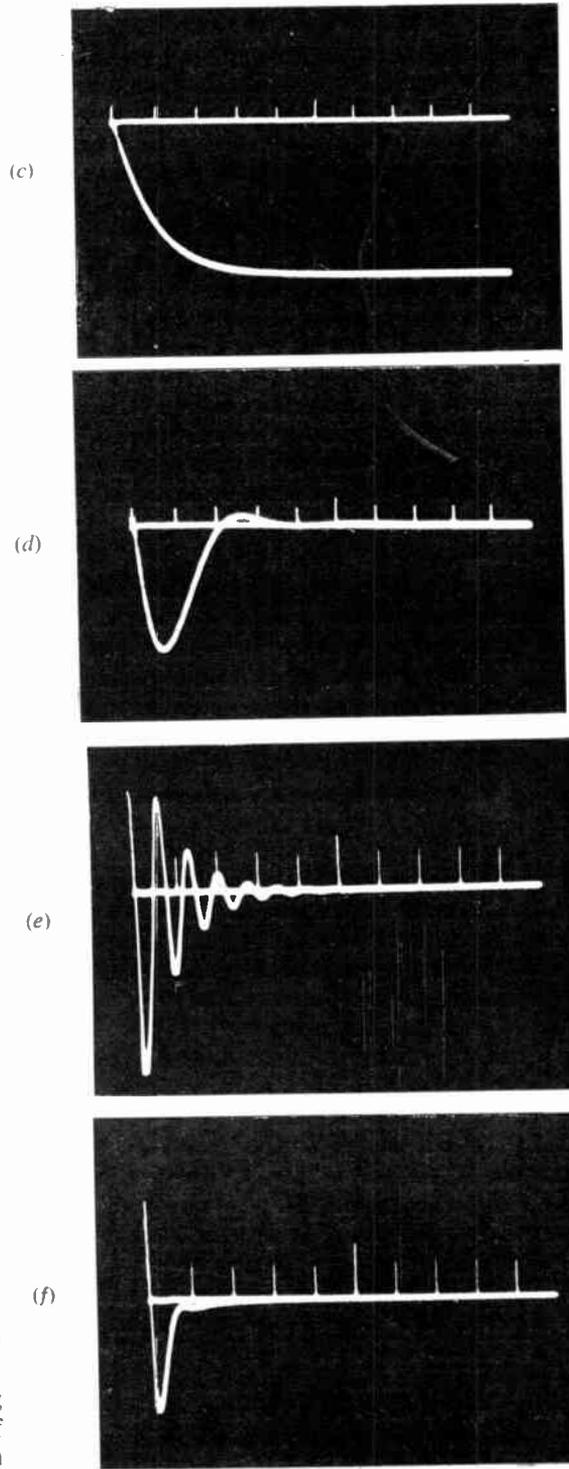
- (c) No stabilizing policy.
- (d) Mixed proportional and integral correction.
- (e) As (d), but stronger policy.
- (f) As (e), but with addition of derivative correction.

u units. In all cases the production time lag was taken as three months, and the correction lag as six months, with a marginal leakage of 0.25.

Figure 15(c) shows the production response in the absence of a stabilizing policy ($f_p=f_i=f_d=0$), production reaching a new steady value in about four years.

Figure 15(d) shows how this response is modified by a policy having a proportional correction factor, f_p of 0.5 plus an integral correction factor, f_i of the same value. The peak change in production has been reduced from $4u$ to $1.5u$ and production level is back to normal after two years. The proportional element in the policy helps to speed up the stabilization and reduce the fluctuation, while the integral term ensures complete long-term correction.

If a stronger policy is adopted, by increasing both f_p and f_i , the maximum variation of production is reduced to $0.5u$ but the system



becomes less stable, with a marked tendency to go into oscillation, Fig. 15(e). However, by introducing a derivative correction factor f_d of value 1.0, the oscillation is eliminated while the residual variation in production is reduced to a disturbance of no more than 0.25 u peak value lasting only 6 months (Fig. 15(f)).

The striking similarity of behaviour between this economic analogue and the feedback systems of communications and servo engineers suggests that it may be possible in the future to control national or industrial economies on scientific principles to avoid the traditional alternation between boom and slump. Unfortunately, it appears at present to be difficult both to arrive at sufficiently simplified models and also to obtain precise information on the system constants.

5.3.4. Flight simulation

Most aircraft flight simulators are large installations intended for crew familiarization and provided with specially designed computing equipment.¹⁵ However, an interest has recently developed in the use of relatively simple "handling simulators" to enable the combination of pilot and machine to be studied as a closed-loop system before the prototype aircraft is built.

In a typical installation the simulator consists of a dummy cockpit coupled to one or more computers which are set up to represent the aerodynamic behaviour of the aircraft under investigation. The cockpit is equipped with the normal pilot's controls and flight instruments, such as compass, artificial horizon, etc. Although these instruments have standard appearance and presentation, they are actually fitted with electrical movements operated by the appropriate signals generated in the computer. Similarly the control column and rudder bar are provided with potentiometers which feed voltages into the computer proportional to the elevator, aileron and rudder movements of the aircraft being simulated. Synthetic reaction forces are also applied to these controls to give correct "feel."

With this equipment an experienced pilot can "fly" a projected design of aircraft "on instruments" and judge whether it will handle

satisfactorily. If it does not, then the characteristics can be modified by adjusting the settings of the computer until the desired behaviour is obtained.

5.3.5. Hysteresis loop

In order to carry out a realistic simulation of certain systems it is necessary to represent true backlash or hysteresis. A device is then needed which will generate a two-valued function of a signal, as distinct from the single-valued output provided by a simple "dead-zone," or limiting, circuit.

Figure 16 shows how a true hysteresis loop may be simulated by the combination of a dead-zone device with a function generator and an integrator. The procedure adopted in adjusting this set-up is as follows:

- (i) The discontinuous function unit is set to give a symmetrical characteristic which determines the width of the loop.
- (ii) The continuous function unit is adjusted to represent a curve which has the shape of the sides of the hysteresis loop.
- (iii) The time constant (RC) of the integrator is made low enough for the unit to act as a high-gain amplifier with negligible integration during the time that the output is rising or falling rapidly (i.e. on the sides of the loop), but high enough to hold the output over the flat portions at the top and bottom of the curve.

Final adjustments to the shape of the loop may be made by displaying it repetitively on the screen of the cathode-ray tube.

6. Conclusions

It has been shown that a general-purpose analogue computer forms an extremely useful tool in almost every sphere of engineering and scientific endeavour. Although it is not generally realized, many instruments now in service have phase or frequency responses that make them incapable of giving even moderately accurate solutions on certain types of problem. However by attention to these, and other, design aspects an overall accuracy of the order of a few per cent. is usually obtainable, and the analogue computer may therefore be regarded as the modern technician's slide-rule.

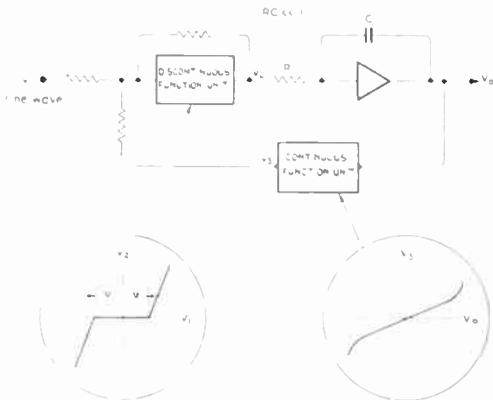
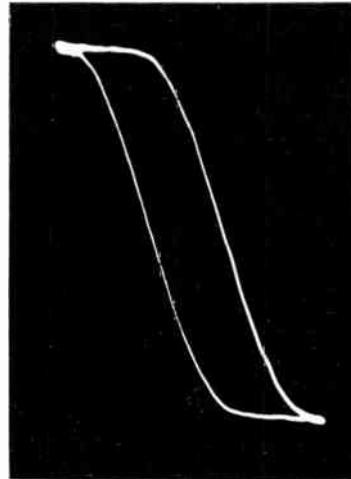


Fig. 16. Generation of hysteresis loop. (above) Arrangement of computing units. (right) Photograph of loop displayed on computer screen.



Despite its limitations in accuracy the analogue computer has a definite field of application alongside the digital computer. One of its main advantages is its ability to perform computation in real time, so that actual components of a system can be included in a simulation. Another important advantage is the physical insight or "feel" that is obtained for the problem.

In the future much wider use will probably be made of the analogue computer for the solution of non-linear problems, particularly as the advantages of this method of treatment become more generally appreciated.

On the future of computer design, apart from improved components, there is a possibility that the use of digital techniques for individual computing units within a continuous overall analogue system may have advantages from the point of view of function generation and integration with respect to variables other than time. Another development may be the use of multi-dimensional displays for high speed computers, where rapid presentation of information is of importance.^{11,12}

7. Acknowledgments

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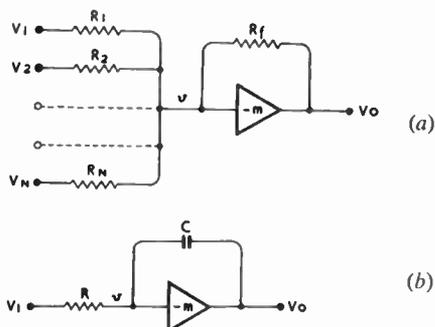


Fig. 17. Basic computing circuits for—
(a) Addition.
(b) Integration.

9. Appendices

9.1. Effects of Finite Amplifier Gain

9.1.1. Adder

Consider the adding unit shown in Fig. 17(a), in which R_1, R_2, \dots, R_N are the input resistors, with corresponding input voltages V_1, V_2, \dots, V_N relative to earth, and R_f is the feedback resistor.

Then, assuming zero input current to the amplifier

$$\frac{V_1 - v}{R_1} + \frac{V_2 - v}{R_2} + \dots + \frac{V_N - v}{R_N} = \frac{v - V_0}{R_f} \dots(1)$$

also $V_0 = -mv$ (2)

Substituting (2) in (1) and re-arranging terms:

$$\begin{aligned} & V_1 \frac{R_f}{R_1} + V_2 \frac{R_f}{R_2} + \dots + V_N \frac{R_f}{R_N} \\ &= -V_0 - V_0 \frac{1}{m} \left[1 + \frac{R_f}{R_1} + \frac{R_f}{R_2} + \dots + \frac{R_f}{R_N} \right] \end{aligned} \dots\dots(3)$$

Thus the percentage error due to the finite gain, $-m$, is:

$$\frac{100}{m} \left[1 + \frac{R_f}{R_1} + \frac{R_f}{R_2} + \dots + \frac{R_f}{R_N} \right] \dots\dots(4)$$

9.1.2. Integrator

Consider the integrator shown in Fig. 17(b) in which R is the input resistor and C is the feedback capacitor.

Then, assuming zero input current to the amplifier:

$$V'_0 = -mv' \dots\dots(5)$$

and $\frac{V'_1 - v'}{R} = \frac{v' - V'_0}{1/pC}$ (6)

Eliminating v

$$V'_0 [1 + 1/m + 1/pmCR] = -V'_1/pCR \dots\dots(7)$$

The required solution is $V'_0 = -V'_1/pCR$... (8) so that the percentage error due to the finite gain is

$$100/V_0 \cdot L^{-1}\{V'_0 [1/m + 1/pmCR]\} \dots\dots(9)$$

Consider two typical inputs:

(i) A step function of value V , i.e. $V'_1 = V/p$.

The required solution is

$$V_0 = -L^{-1}\{V/p(pCR)\} = -Vt/CR \dots\dots(10)$$

where t is the time of computation in seconds.

From (9) the percentage error is therefore

$$\frac{100 L^{-1}\{V'_0 [1/m + 1/pmCR]\}}{Vt/CR}$$

i.e. the percentage error is

$$100\{1 - (mCR/t)[1 - \exp(t/CR(1+m))]\} \dots(11)$$

or, approximately,

$$100 \left\{ 1 - \frac{m}{m+1} \left[1 - \frac{1}{2} \frac{t/CR}{(1+m)} \right] \right\} \dots\dots(12)$$

The relationship between percentage error and finite gain is shown in Fig. 18 for various values of t/CR .

(ii) A sine wave given by $V_1 = V \sin \omega t$,

i.e. $V'_1 = V \cdot \frac{\omega}{p^2 + \omega^2}$ (13)

From equation (7)

$$V'_0 [1 + 1/m + 1/pmCR] = -V'_1/pCR$$

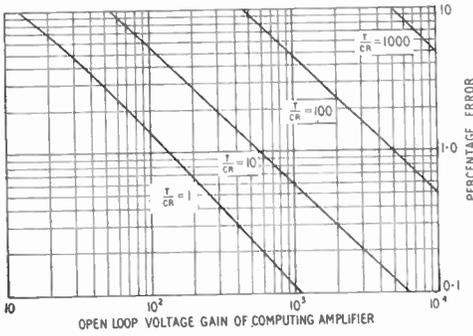


Fig. 18. Relationship between percentage error and open loop gain for integrator with step input function.

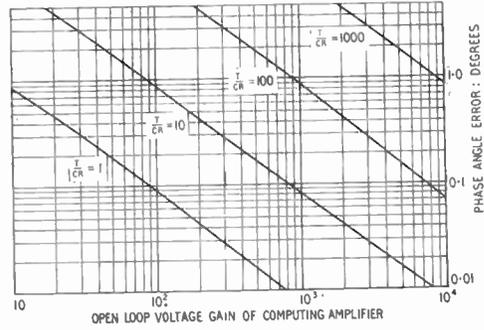


Fig. 19. Phase angle error due to finite open loop gain in integrator with sinusoidal input.

Substituting (13) in this equation

$$V'_0 = -V\omega m \left[\frac{1}{pCR(1+m)+1} \cdot \frac{1}{p^2 + \omega^2} \right]$$

The solution to this equation is given by:

$$V_0 = \frac{V\omega m}{1 + \omega^2 k^2} \left[k e^{-t/k} - (k^2 + 1/\omega^2)^{1/2} \cos \omega t + \varphi \right] \dots\dots\dots(14)$$

$$\text{or } V_0 = \frac{V\omega m k}{(1+m)\sqrt{(1+\omega^2 k^2)}} \cdot \frac{1+m}{\omega k} \times \left[\cos(\omega t + \varphi) - \frac{\omega k}{\sqrt{(1+\omega^2 k^2)}} \cdot e^{-t/k} \right] \dots\dots\dots(15)$$

where $k = CR(1+m)$ and $\varphi = \tan^{-1}(1/\omega k)$.

The desired solution is $V_0 = \frac{V(\cos \omega t - 1)}{\omega CR}$

$$\text{or } V_0 = \frac{V(1+m)(\cos \omega t - 1)}{\omega k} \dots\dots\dots(16)$$

It will thus be noted that the solution given by the integrator differs from the desired solution in three respects:—

(a) The overall amplitude is reduced by a factor

$$\frac{\omega m k}{(1+m)\sqrt{(1+\omega^2 k^2)}} \dots\dots\dots(17)$$

(b) An exponential error term exists having a value

$$1 - \frac{\omega k}{\sqrt{(1+\omega^2 k^2)}} \cdot e^{-t/k} \dots\dots\dots(18)$$

(c) There is a leading phase angle error

$$\varphi = \tan^{-1}(1/\omega k) \dots\dots\dots(19)$$

The relationship between phase angle error and open loop voltage gain, for various values of τ/CR (i.e. ratio of periodic time of input signal to integrator time constant CR), is shown in Fig. 19.

9.2. Influence of Finite Bandwidth

Consider an equation of the form

$$d^2y/dt^2 + \omega_n^2 y = F(t) \dots\dots\dots(20)$$

A computing arrangement for solving this problem is shown in Fig. 20. When the

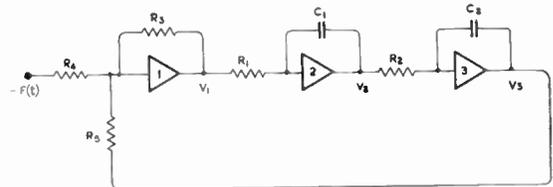


Fig. 20. Set-up for second-order equation.

integrator coefficients $1/R_1 C_1 = 1/R_2 C_2 = \omega_n$, the response of the system to a step function input of value V is given by

$$V'_1 = \frac{1}{1+p\tau} \left\{ \frac{R_3}{R_5} V'_3 + \frac{R_3}{R_3} \left[-F'(t) \right] \right\} \dots\dots\dots(21)$$

if the adding amplifier has a simple time constant τ .

If, also, $R_3=R_4=R_5$ and the integrators are assumed to be perfect, then

$$V'_1 = -\frac{1}{1+p\tau} \left[V'_3 - \frac{V}{p} \right]$$

$$V'_2 = -\frac{\omega_n}{p} V'_1$$

$$V'_3 = -\frac{\omega_n}{p} V'_2 = \frac{\omega_n^2}{p^2} V'_1$$

i.e. $V'_1 = p^2 V'_3 / \omega_n^2$

Substituting this value of V'_1 in equation (21)

$$\frac{p^2 V'_3}{\omega_n^2} = -\frac{1}{1+p\tau} \left[V'_3 - \frac{V}{p} \right]$$

i.e. $V'_3 = \frac{\omega_n^2 V}{\tau} \frac{1}{p [p^3 + p^2/\tau + \omega_n^2/\tau]}$

The solution of this equation is given by

$$V_3 = V \left\{ 1 - \frac{4\alpha^2}{\omega_n^2 + 4\alpha^2} e^{-\gamma t} - \frac{\omega_n^2}{\omega_n^2 + 4\alpha^2} e^{\alpha t} \cos \beta t - \frac{\omega_n^2 - 8\alpha^2}{\omega_n \left[\frac{\omega_n^2}{\alpha} + 4\alpha \right] \left[1 - \frac{5\alpha^2}{2\omega_n^2} \right]} e^{\alpha t} \sin \beta t \right\} \dots(22)$$

where

$$\gamma = \frac{\omega_n^2}{2\alpha} \left[1 - \left(\frac{2\alpha}{\omega_n} \right)^2 \right]$$

$$\beta = \omega_n \left[1 - \frac{5\alpha^2}{2\omega_n^2} \right]$$

$$\frac{1}{\tau} = \frac{\omega_n^2}{2\alpha} - 4\alpha$$

An alternative form of the solution is

$$V_3 = V \left[1 - \frac{4\alpha^2}{\omega_n^2 + 4\alpha^2} e^{-\gamma t} - (A^2 + B^2)^{1/2} e^{\alpha t} \cos (\beta t - \varphi) \right]$$

where $\dots(23)$

$$A = \frac{\omega_n^2}{\omega_n^2 + 4\alpha^2}$$

$$B = \omega_n \left[\frac{\omega_n^2}{\alpha} + 4\alpha \right] \left[1 - \frac{5\alpha^2}{2\omega_n^2} \right]$$

and

$$\varphi = \tan^{-1} (B/A)$$

The solution which would be obtained from an adder with infinite bandwidth (i.e. $\tau=0$) is given by

$$V_3 = V (1 - \cos \omega_n t) \dots(24)$$

The actual solution is therefore in error in five respects:—

(i) There is a d.c. shift of the solution given by $-\frac{4\alpha^2}{\omega_n^2 + 4\alpha^2} e^{-\gamma t} \dots(25)$

(ii) The frequency of the solution should be $\omega_n/2\pi$, but is in fact $\beta/2\pi$ where $\beta = \omega_n \left[1 - \frac{5\alpha^2}{2\omega_n^2} \right] \dots(26)$

(iii) There is a lagging phase angle error φ , where $\varphi = \tan^{-1}(B/A) \dots(27)$

(iv) An amplitude error of $\sqrt{(A^2 + B^2)}$ exists in the cosine term. $\dots(28)$

(v) The solution is divergent, as shown by the term $e^{\alpha t} \dots(29)$

Of the various errors present, the most significant is the last.

If α is small, then $1/\tau \cong \omega_n^2/2\alpha$ and if the cut-off frequency of the adding amplifier is

$$f_c = \omega_c/2\pi, \text{ then } \omega_c = 1/\tau = \omega_n^2/2\alpha$$

i.e. $\alpha = \omega_n^2/2\omega_c$

As an example of the order of this effect, consider the case where the amplitude must not change by more than 1 per cent. during a computation time t seconds.

Then $0.01 = \alpha t = (\omega_n^2/2\omega_c) t$

i.e. $\omega_c/\omega_n = 100\omega_n t/2$

or $f_c/f_n = 100\pi f_n t \dots(30)$

Alternatively the number of cycles which can be displayed during a time t is given by

$$N = f_n t$$

If $f_c = 15 \text{ kc/s}$, then $f_n = 6.92/\sqrt{t}$

or $N = 6.92\sqrt{t} \dots(31)$

9.3. Effects of Phase Shift

It can be shown that the phase shift round any particular computing loop due to the finite bandwidth of the amplifier, stray capacitance or inductance associated with the feedback components or leakage paths, will result in errors in the solution. If the computing loop shown in Fig. 20 is again considered and the phase shift error round the loop is ζ radians, it

can be shown to a first approximation that

$$\alpha = \omega_n \zeta / 2 \quad \dots\dots\dots(32)$$

The logarithmic decrement is therefore

$$\alpha \cdot 2\pi / \omega_n = \pi \zeta \quad \dots\dots\dots(33)$$

i.e. $\pi \times$ (total phase shift round the loop).

Alternatively, if the amplitude of the solution must not change by more than 1 per cent. during a computing time of t seconds, then

$$\alpha t < 0.01$$

i.e. $\omega_n \zeta t / 2 < 0.01$

Therefore $\zeta < 0.02 / \omega_n t$ radians

$$\zeta < \frac{1.8}{\pi^2 f_n t} = \frac{0.181}{f_n t} \text{ degrees} \quad \dots\dots\dots(34)$$

or $N = f_n t = 0.181 / \zeta \quad \dots\dots\dots(35)$

where N is the number of cycles displayed in the time t .

It may be noted that a negative phase angle error round the loop results in a divergent error term in the solution while a positive phase angle error results in a convergent error term.

The sign convention taken here is that a leading phase change is regarded as positive for a system where the forward direction round the loop gives integration.

9.4. Effects of Drift (Offset) Voltage and Input Grid Current

Consider the circuit shown in Fig. 21 where i_1 is the input current and i_2 the current flowing through Z_2 .

Then $i'_1 = i'_2 + i'_g$.

Therefore $(V'_1 - v') / Z'_1 = (v' - V'_0) / Z'_2 + i'_g$

or $v' \left[\frac{1}{Z'_1} + \frac{1}{Z'_2} \right] = -i'_g + \frac{V'_0}{Z'_2} + \frac{V'_1}{Z'_1}$

But, since $V'_0 = -m(v' - V'_d)$
 $v' = V'_0 / m + V'_d$

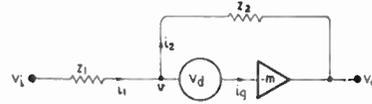


Fig. 21. Circuit for investigation of drift and grid current.

By elimination

$$-\frac{V'_0}{m} \left[\frac{1}{Z'_1} + \frac{1}{Z'_2} \right] - \frac{V'_0}{Z'_2} = -i'_g + \frac{V'_1}{Z'_1} -$$

$$-V'_d \left[\frac{1}{Z'_1} + \frac{1}{Z'_2} \right]$$

or $-V'_0 Z'_1 (1+m) + Z'_2$
 $= -i'_g m Z'_1 Z'_2 + V'_1 \cdot Z'_2 - V'_d m (Z'_1 + Z'_2)$

i.e. $V'_0 = -\frac{V'_1 m Z'_2}{Z'_1 (1+m) + Z'_2} + \frac{i'_g m Z'_1 Z'_2}{Z'_1 (1+m) + Z'_2} +$
 $\frac{V'_d m (Z'_1 + Z'_2)}{Z'_1 (1+m) + Z'_2} \dots\dots\dots(36)$

In the case of an integrator

$$Z'_1 = R \text{ and } Z'_2 = 1/pC$$

If i_g and V_d are both constant

$$i'_g = i_g / p$$

and

$$V'_d = V_d / p$$

Therefore $V'_0 =$

$$-\frac{V'_1 m}{1 + pCR(1+m)} + \frac{i_g m R}{p[1 + pCR(1+m)]} +$$

$$+\frac{mV_d(1 + pCR)}{p[1 + pCR(1+m)]}$$

If ϵ represents the error due to grid current and drift voltage combined

Then $\epsilon' = \frac{(i_g m R + mV_d) + mpCRV_d}{p[1 + pCR(1+m)]}$

The solution of this equation is given approximately by

$$\epsilon = m(i_g R + V_d) [1 - \exp\{t/CR(1+m)\}] \dots\dots\dots(37)$$

AUTOMATION, EDUCATION AND TRAINING

THE above title was the theme of a Conference held in London recently by the British Association for Commercial and Industrial Research. The Association was founded in 1919 to promote and advance schemes of education and training in industry and commerce, and to encourage co-operation between industry and education authorities, colleges and teachers. The Association's latest Conference was based on three papers dealing with the engineering, educational and economic aspects of automation.

Mr. Frank Woollard, who was for many years a Director of Morris Motors and General Manager of the Engines Division, defined automation as "the system and method of making processes automatic by the employment of self-acting, self-controlling means for performing necessary operations." He pointed out that automatic machinery has existed for almost 200 years, and that automation was merely an evolution of this process, accelerated by the intensive wartime study of automatic control.

Mr. Woollard stressed the fact that the introduction of automation in the manufacturing industry saves labour, floor area, spoilage; gives greater control of processing, better control of stock, and fuller utilization of plant. However, it also brings its problems, mainly in management. He stressed, too, the need for longer-term planning, with a more co-operative and communicative management, and for much closer team work between all departments. It is also important that all levels of management, all engineers and technicians, should have an appreciation of the principles of flow production and industrial administration.

At one time Under-Secretary and Head of the Further Education Department at the Ministry of Education, Mr. Frederick Bray, C.B., dealt with the education required in an age of automation.

The main problem was a shortage of students of the right calibre, and Mr. Bray suggested that what was needed was a greater investment in schools and education. This could be done by carrying out the recommendation of the 1944 Education Act, for an ultimate school-leaving age of 16 years, which would result in a raising of the level of education. It was also important that industry should look very closely at its apprenticeship schemes: many of them based on craft training schemes adopted years ago were still being used for student apprentices who were

reaching professional status in their academic studies. An efficient system of apprenticeship was absolutely essential if industry was to provide the personnel able to develop, build, install and maintain automation equipment. It was not enough to have an apprenticeship scheme—there must be several schemes for the various grades of personnel being trained.

In ten years' time the school-leaving population will temporarily increase by one-third, compared with today's figure of 640,000. The present training facilities, which are inadequate, will be quite incapable of dealing with this "bulge." Industry must now try to increase its training facilities, so that engineering and technology could maintain the percentage of recruits from the school-leaving population.

There was a possible wastage of intellect in female population. Many of them were quite up to the standard required, and capable of benefiting from scientific and technological training.

Mr. E. Shenfield, Economic Adviser to the Federation of British Industries, dealt with the impact of automation on industrial development. He expressed the opinion that the introduction of automation would be limited by a number of factors—the cost of capital outlay, the shortage of technologists and technicians, and the cost of production. To be successful, automation must lead to an increasing production at a lower price, thus reaching a wider market. In the question of markets, tariff policies and external trade were bound to have a very considerable effect.

There was a discussion concerning the need for re-training of mature and middle-aged operatives and semi-skilled personnel. They undoubtedly experienced embarrassment by attending technical colleges. This was obviously a case for training-within-industry courses.

The Conference was summed up by Mr. Geoffrey Mills, Deputy Comptroller to J. Lyons and Co., who had wide experience on the computer operations of office work. He referred to the conflicting views on the speed of introduction of automation equipment. In his view, its introduction would be not fast enough for the management, too fast for the workers, and much too fast for the educational system. It was therefore essential that industry and the technical colleges should immediately make preparations for the proper training of the necessary personnel.

A LOW-NOISE TRAVELLING-WAVE TUBE AMPLIFIER FOR THE 4000-Mc/s COMMUNICATIONS BAND*

by

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*A paper read before the Scottish Section of the Institution on 24th March, 1956.
In the Chair: Mr. R. H. Garner, B.Sc.(Eng.)*

SUMMARY

The factors which influence the performance of a low-noise travelling-wave tube are discussed and design data are developed for a particular tube. The mechanical design is considered and some performance curves presented.

1. Introduction

In a travelling-wave tube an electron beam is allowed to interact with an electromagnetic wave travelling in the same direction and at nearly the same velocity as the beam. To slow the wave down it is made to travel round the turns of a helix while the beam takes the shorter path along the axis. Since at high frequencies we may think of the wave travelling along the wire with the velocity of light (c) it follows that the resolved velocity along the axis is $c \sin \psi$ where ψ is the pitch angle of the helix. Divergence of the beam due to space-charge forces is prevented by using an axial magnetic field. The helix is not the only slow wave structure but it has the advantage that it is electrically smooth and its bandwidth can be made very great. The enormous potential bandwidth of the travelling-wave tube is its principal virtue.

2. Gain of Travelling-wave Tube

The axial field due to the signal on the helix has a longitudinal component so that electrons in the beam may be either in an accelerating region or a retarding region. The mean acceleration of an electron as it travels through the wave will depend on the relative velocity of electron and wave. If the electrons are injected with a velocity exceeding that of the

wave they are, on the average, slowed down and the kinetic energy which they lose is converted into field power in the wave.

The accelerating and retarding forces acting on the beam produce bunching effects which in turn give rise to additional longitudinal fields. To simplify the picture we may ignore the magnetic field if we replace it by a positive ion distribution having a density equal to the steady value of the space-charge density in the beam. The positive ions, because of their greater mass, will be hardly affected by the high-frequency fields, but since the total charge is zero the steady space-charge fields will be zero and we can thus neglect the radial forces tending to make the beam diverge. Now we can see that the bunching forces will produce alternate regions in the beam in which there is an excess of positive or negative charges. This separation of charge produces restoring forces with the result that superimposed on the steady drift of the electrons down the tube is a small oscillatory motion.

It may be shown that the electron beam can support two longitudinal waves; one travels slightly faster than the beam and the other slower. The helix, like any other transmission line, can also support two waves travelling in opposite directions so that we have in the travelling-wave tube a system of four waves. These are coupled together (Appendix 1). The effect of the coupling is to allow power to flow from the beam into the circuit waves with the result that one forward wave on the helix becomes a growing wave.

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With the assumptions, appropriate to a low-noise tube, of small signal level, small beam diameter, small coupling and negligible losses, the gain in decibels is given by the expression

$$G = A + B CN$$

Here A and B are constants and C/N is a function of the current and voltage of the beam and the helix form and length. Usually we know the gain and the voltage required and can choose helix dimensions so that the electron velocity is similar to the wave velocity. It is apparent that the initial electron velocity must be greater than the wave velocity of the increasing wave for the tube to give gain; in practice the tube voltage is adjusted when operating to give maximum gain.

3. Noise

3.1. Sources of Noise

In a properly designed microwave amplifier the cathode is the most serious source of noise. However, there are other possible sources which can become important if adequate precautions are not taken. We will consider them briefly:

(a) Partition noise due to the random collection of electrons by the gun electrodes or the first few centimetres of the helix can be serious since less than 1 per cent. of the total current intercepted will give a minimum noise factor of more than 10 db. This effect can be reduced to negligible proportions by good mechanical alignment of the gun with respect to the helix and by allowing adequate apertures in the gun anodes.

(b) A form of partition noise can arise due to electron collisions with gas molecules but it has been found that this effect is negligible if the pressure of gas in the tube is less than 10^{-6} mm of mercury.

(c) A further source of noise is "induced" partition noise due to random transverse excursions of electrons in the beam between regions of large and small axial r.f. fields. Complete freedom from radial movement of electrons in a magnetically constrained beam is virtually impossible, but the effect may be kept small if the diameter of the beam is small compared with that of the helix and the beam is in the centre of the helix. Again, accurate mechanical alignment ensures this.

3.2. Noise Reduction

We have already suggested that a magnetically constrained beam of electrons is a medium which will support two longitudinal waves. These waves can be excited similarly by a noise source in the gun, and will interfere to give periodic maxima and minima of a.c. velocity, alternating current, a.c. charge density or longitudinal field. These maxima and minima correspond to the normal kinematic bunching of klystron theory.

If the potential of the beam is changed suddenly we find that the amplitude of the interference pattern may change. In Appendix 2 we obtain the following expression for this change at such a jump:

$$\frac{\hat{E}_2^2}{\hat{E}_1^2} = (r_1/r_2)^2 (V_1/V_2)^{1/2} \sin^2(\omega_{q1}z/u_{01}) + \cos^2(\omega_{q1}z/u_{01})$$

where V is the voltage, u_0 the beam velocity, r the plasma-frequency reduction factor, z the distance along axis, and subscript 1 applies to the region before the change and subscript 2 to the region after.

If $V_2 > V_1$ and we choose a value of z such that $\cos(\omega_{q1}z/u_{01}) = 0$ then there will be a reduction in noise. It is interesting to note that at z such that $\sin(\omega_{q1}z/u_{01}) = 0$ we may reduce the potential without increasing the value of E and in practice this method may be useful. This method of reducing noise by means of potential jumps was suggested by Watkins.¹

It might appear that E could be reduced indefinitely by cascading jumps and drift tubes, but this is true only for a single source of noise: there may be others which are independent.

For a cathode operated under space charge limited conditions there are two sources of noise. These give rise to velocity and current variations in the electron beam at the potential minimum in front of the cathode.² The more important of these has maximum mean square velocity fluctuations at this plane given by

$$\bar{v}_a^2 = (4 - \pi)\eta(KT_c/I_0)\Delta f$$

The other source has maximum mean square current fluctuations given by

$$\bar{i}_a^2 = 2R^2eI_0\Delta f$$

where R^2 is an effective "space charge smoothing" factor defined by Watkins.

Both sources excite plasma waves on the beam, the envelope of one being out of phase by $\pi/2$ with the other. The gun is designed to reduce the major source (that due to velocity fluctuations), but the presence of the other uncorrelated source provides a finite minimum limit for the noise factor.

4. Design

The complete design for a low-noise travelling-wave amplifier for communication purposes falls naturally into five distinct sections.

- (a) The slow-wave structure,
- (b) The electron gun,
- (c) Focusing methods,
- (d) Coupling to external transmission lines and matching,
- (e) The attenuating section.

It is proposed to deal with each of these in turn.

4.1. The slow-wave structure

The tube under consideration is for communication purposes and may be incorporated in links carrying 600-channel telephony. It is therefore important that phase distortion and group delay variations over a frequency band of up to perhaps 40 Mc/s are very small. A dispersionless structure is thus required for the tube and of the many known types of structure the helix and the interdigital line alone fulfil this requirement. The effective coupling impedance between beam and circuit wave for the latter type is low since it has to rely on the first forward harmonic for broadband amplifier application. The simple helix thus becomes the natural choice of structure. The dimensions of the helix are set mainly by the frequency band over which the tube is required to operate and also by the required synchronous velocity. For lowest noise-factor the gain per unit length of the helix should be high and this can be

achieved by making the helix diameter small. An important parameter is $\omega a/u_0$ where a is the mean helix radius and u_0 is the mean electron velocity; ω is the radian signal frequency. We find that if $\omega a/u_0 < 0.8$ the structure becomes dispersive leading to possible group delay distortions. Pierce and others have shown that for optimum operation of a helix, for constancy of gain over a bandwidth, with a small beam diameter, $\omega a/u_0$ should be about 1.5. With this value and a predetermined synchronous velocity the radial dimensions of the helix are uniquely determined. It then remains to calculate the other relevant parameters of the tube and the helix length in order to achieve the required gain. The wire gauge is chosen to give a spacing between turns of slightly more than the thickness.

The expression for the gain $G = A + BCN$ assumes a structure free from both distributed and lumped attenuation. Pierce³ has shown that losses may be allowed for approximately by introducing two further terms. The gain relation now becomes

$$G = A + BCN + \alpha L + \zeta \quad \text{decibels}$$

in which αL and ζ describe the losses introduced by the finite resistance of the helix wire and the attenuating section respectively. In this tube $A + \alpha L + \zeta \cong -17$ db. An estimate of the gain to be obtained under certain conditions then depends on the ability to calculate the "gain parameter." C , which involves a knowledge of the effective "coupling impedance" from beam to wave. Pierce has considered this for a lossless sheath helix in free space but his results must be greatly modified to account for the effect of space harmonics (present when a wire or tape helix is used), the effect of the dielectric supports, etc. Tien⁴ has considered this in some detail and using his results a more accurate estimate can be made. In practice, dividing Pierce's value by 2 gives a very good estimate.



Fig. 1. Low noise travelling-wave tube amplifier for 4,000 Mc/s communication band. (Type VX 8168.)

4.2. The electron gun

In a tube of this type the electron gun has to fulfil two functions. It must first produce an electron beam with parallel flow and fairly clearly defined boundaries and in addition it must be a compromise between the points outlined in Section 3 to ensure that its noise modulation is minimized before it reaches the helix.

In the gun designed for the present tube (Fig. 1) the position and shape of the cathode electrode and accelerator are designed to produce a parallel beam. The anode, operating at a potential of 120 volts, is followed by a drift space at the same potential, which is made nearly a quarter plasma wavelength long in order that an a.c. field maximum appears at the end furthest from the cathode. This drift tube is terminated in an apertured disk one millimetre from which is situated a further aperture at approximately half the potential. This gap, occurring at an a.c. field maximum, enables a downward jump in potential to be made, so that the following upward jump can be more effective. The fact that this jump occurs at the field maximum implies that the noisiness of the beam will be unaffected (see Section 3.2).

The second anode, operating at approximately 60 volts, is followed by a further drift space at the same potential which is again made a quarter plasma wavelength long. Another apertured disk terminates this space and an upwards voltage jump takes place between this and the end of the drift tube which is connected to the helix. This upward jump from approximately 60 to 600 volts gives an effective noise reduction of $(60/600)^{3/2} \cong 1/31$ (see Appendix 2).

After leaving this second velocity jump the electron beam travels down the drift tube at the same potential as the helix and then enters the helix itself. The approximate length of this drift tube is found by calculation and it is finally adjusted empirically for lowest noise output.

The calculation of the plasma wavelength at the various potentials has been made from curves published in Watkins' paper.¹ Where possible the drift tubes surrounding the beam have been made large in diameter compared with the beam in order that the plasma

frequency is almost independent of the operating signal frequency of the tube. The gun can thus be "broadbanded" to a certain extent.

4.3. The focusing system

An important requirement is that the electrons in the beam do not make large transverse excursions during their path through the gun anodes and the initial few centimetres of the helix. This can best be arranged (although at the expense of using a high magnetic field) by immersing the gun and helix regions in a uniform magnetic field of magnitude several times that required for Brillouin flow, which in this case is about 80 oersteds. An arrangement of three coils is used to produce a suitable uniform field of about 320 oersteds.

A photograph of the coils with the waveguide circuit attached is shown in Fig. 2.

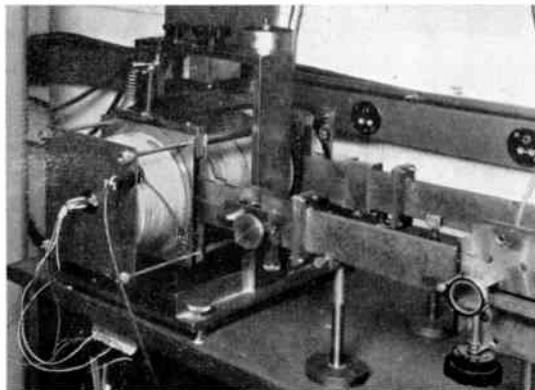


Fig. 2. The coils for the travelling-wave tube and the associated waveguide circuits.

Undoubtedly the best method of focusing for practical application is the spatially-periodic system using permanent magnets. This wastes no power and has the added advantage that the external stray field is small. With this method, however, some scalloping of the beam is inevitable and this can materially increase the noise factor. Recent work by Chang⁵ suggests that low-noise operation for this type of tube is possible with only a small degradation of performance, so long as there is a solenoid providing a uniform field in the gun region. Experiments are in progress in these laboratories with the object of designing low-noise

tubes focused solely by a periodic permanent magnet system.

4.4. *Coupling to the helix*

Since this tube was intended primarily for communication purposes it was considered that a direct transition from helix to waveguide at input and output was essential as the feeder from the receiving aerial in a link repeater would certainly be waveguide.

The system used for the helix-to-waveguide transition is a scaled-up version of one originally used in these laboratories at X-band. The requirement in this tube is for a match of better than 0.9 input v.s.w.r. over any 40 Mc/s within the band 3.6-4.2 kMc/s. The system actually provides a v.s.w.r. of better than 0.6 over the whole band and with a single tuning screw in the waveguide this is improved to the required match of better than 0.9. A sketch of the matching system is shown in Fig. 3. The narrow dimension of No. 12A waveguide is reduced to give an effective impedance of about 50 ohms to match that of the coaxial line formed by the drift tube and the outer brass tube of the circuit, into which the travelling-wave tube fits. This junction is broadbanded by the addition of a "doorknob," the rear section of which performs the function of a choke and provides an effective contact at the waveguide plane. The inner conductor of this coaxial line is coupled by means of a one step binomial coupler (or $\lambda/4$ transformer) to the helix. This transformer takes the form of a section of helix, a quarter wavelength long at mid-band, wound at a pitch to give the correct impedance intermediate between that of the coaxial line and the interacting helix. A multi-step binomial coupler could be used for broader band operation.

Typical plots of the broad-band and narrow-band matches with and without the external tuning screw are shown in Figs. 4 and 5. An identical coupling system is used at the output end of the tube. The curves shown are for the tube under normal working conditions.

4.5. *Attenuating section*

Attenuation is necessary in the helix so that power reflected at the output and input does not give sufficient feedback for oscillation. The helix has a loss of about 8 db and additional

attenuation is provided in an intermediate attenuating section. The modulation is carried past the attenuation in the forward direction by the beam. The attenuating section must not only provide sufficient loss to prevent instability in operation but must be well matched to the active sections of the helix at either end. Any severe reflections cannot be compensated by matching components in the external waveguide as the active sections of helix are necessarily several wavelengths long.

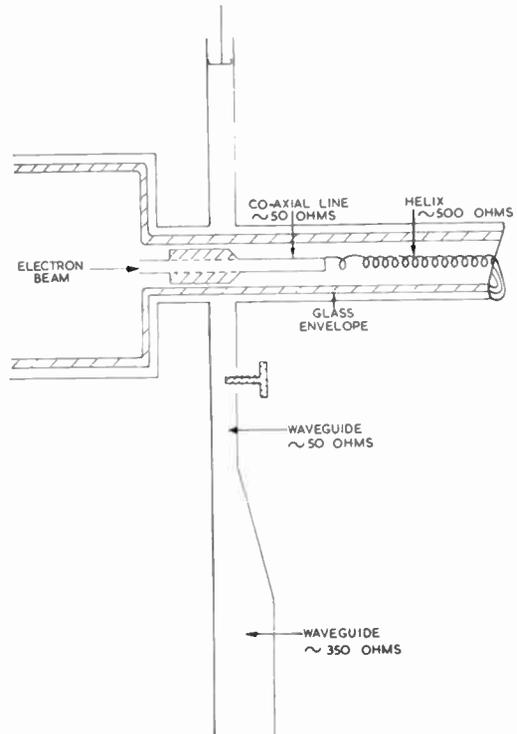


Fig. 3. Matching system for the tube.

The helix in this tube is wound from Kanthal wire, copper plated over its interaction sections with the attenuating section left unplated. Theoretical considerations would suggest that there is a small change in impedance between the conducting and resistive regions. There is also a change in thickness of the wire. In practice, there is enough reflection to make the matching difficult. The junctions between the active sections of the helix and the attenuator in this tube are arranged to be reflectionless by tapering the thickness of the copper plating.

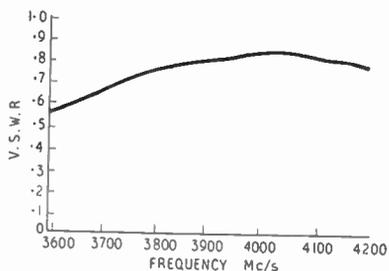


Fig. 4. Broad-band matching.

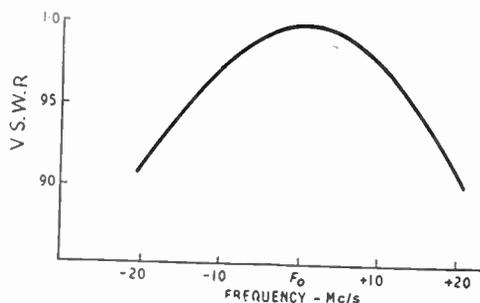


Fig. 5. Narrow-band matching.

5. Mechanical Design

We have stated that interception of beam current at the input of the helix will lead to high noise factors because noise generated here will receive full amplification.

The fact that the helix must intercept no beam current over at least the first part of its length makes considerable demands upon the accuracy of construction of the tube. There are three main conditions to fulfil: (1) the helix must be supported to be as straight as possible over its whole length: the supports should introduce as little dielectric loss or disturbance of the helix fields as possible; (2) the gun assembly must be rigidly constructed and arranged so that its axis is accurately determined; (3) the axis of the gun and helix must be coincident.

Figure 1, which shows a

sectioned drawing of the complete tube, indicates how these are achieved. A photograph of a typical tube is shown in Fig. 6.

The helix is held within a low-loss glass spiral whose inner and outer diameters are accurately maintained throughout its length. This glass spiral is fixed inside a length of precision-bore glass tubing which also forms the vacuum envelope. The interaction helix is thus held concentric with the inside bore of the envelope.

The electrodes of the gun, which are provided with accurately punched holes for the beam aperture and fixing rods, are assembled on precision ceramic tubes and jugged by means of the cathode electrode assembly. The cathode is held within this by quartz spacers. At each end the gun is provided with mica discs which are made a push-fit into precision-bore glass tubing of larger diameter. By arranging that the two sections of the envelope are sealed together on a precision carbon mandrel concentricity between the gun and helix axis is assured.

The assembly is completed by sealing in the Kovar collector which also serves to locate the end of the tube in its circuit.

6. Life

For a tube to be useful for continuous service

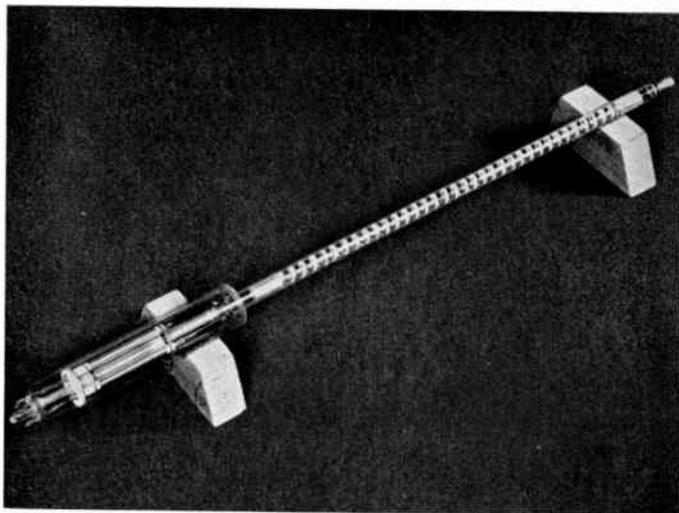


Fig. 6. General view of a typical tube.

in a communication link it must have a life of not less than 5,000 hours and preferably should be usable for as long as 10,000 hours.

The life of a low-noise travelling-wave tube amplifier can be considered terminated when the tube becomes too noisy for its application. Assuming that at this point the gas pressure has not risen to greater than 10^{-6} mm Hg, the increase in noise factor will almost certainly be due to deterioration in the emission properties of all or part of the cathode surface. This emission failure can be caused both by the steady evolution of gas from the cathode and other parts of the tube during life, and also by the build-up of an interface layer between the active surface and core metal of the cathode. Emission can also be impaired by mechanical disruption of the cathode surface due to ion bombardment. This effect is probably more serious in travelling-wave tubes than in other microwave devices due to the greater chance of gas ionization within the long beam lengths employed.

Gas evolution can be minimized by very careful outgassing of all parts of the tube on the pump and by operating at low potentials and low currents. Operation at low voltages and currents will also assist in reducing the effects of ion bombardment.

The results from life tests have been encouraging and in fact tubes operating under normal conditions have completed 5,000 hours without significant change of noise factor.

7. Further Aspects

A number of tubes have been made in an endeavour to establish whether tubes of this type show consistency from sample to sample. Tests of about twenty tubes have shown noise factors ranging from 8.5 to 16 db though each tube was of identical geometry. Investigations have shown that this scatter is due to non-uniformity of emission density over the surface of the cathode. It has, in fact, been possible to correlate high noise factors with only very slight departures from the ideal current/voltage characteristics for a space charge limited diode. Where complete activation of the cathode has been obtained at a higher temperature the noise factor has been high due to the increased temperature which gives rise to an increased

noise velocity component at the cathode.

Experimental evidence shows that for consistent low-noise performance great care must be taken over the processing and cathode activation. Practical results have also shown the importance of having means of controlling the heater voltage applied to the tube. The noise factor is a function of the cathode temperature; if the cathode is cooled the noise factor rises very sharply as the temperature limited region is reached whereas at high temperatures the velocity distribution of emission is increased which again raises the noise factor, though less rapidly. It may therefore be necessary for equipments using such a tube to incorporate a variable heater voltage control on which a few small changes may have to be made during life. At the present time conventional oxide-coated cathodes are used as these allow operation at the lowest temperatures. Experiments are in hand to investigate different cathodes which, although requiring higher temperature operation, may give greater uniformity.

8. Design Details of Tube

Helix

Outer diameter	1.84 mm
Inner diameter	1.64 mm
Wire	Kanthal
Wire thickness	0.1 mm
Pitch	92 t.p.i.
Length	18.3 cm (including attenuator)

Beam

Diameter	0.5 mm
Current	120 μ A
Synchronous potential	600 volts
Cathode	0.5 mm oxide-coated
Cathode current density	60 mA/cm ²

Focusing

Uniform field	320 oersteds
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9. Performance

Operating frequency range	3.6-4.2 kMc/s
Gain	15-18 db (at $I_{coll} = 120 \mu$ A)
Power out (saturated)	2 mW
Noise factor	8.6 db
I_{coll}	120 μ A

I_{helix}	$< 1\mu A$
$I_{A1} + I_{A2}$	$< 1\mu A$
Input match	v.s.w.r. > 0.92 over any 40 Mc/s in the band
Output match	v.s.w.r. > 0.88 over any 40 Mc/s in the band

Some performance curves are shown in Figs. 7, 8 and 9.

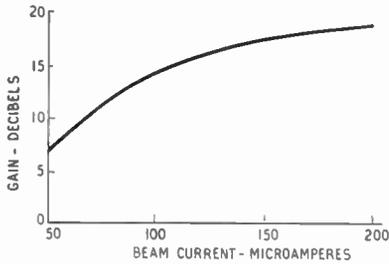


Fig. 7. Gain of tube against beam current.

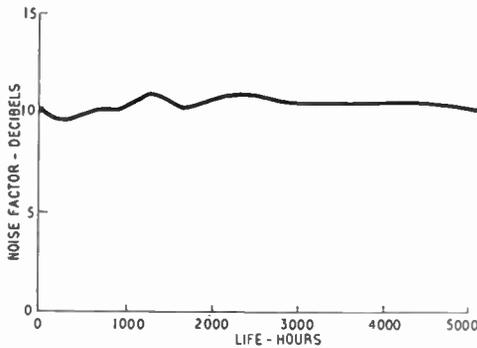


Fig. 8. Variation of noise factor with life.

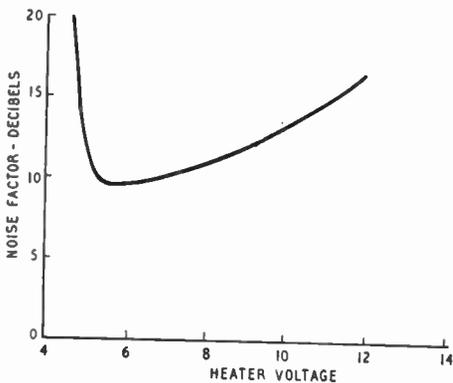


Fig. 9. Variation of noise factor with heater voltage.

10. Acknowledgments

The authors thank the Admiralty, the Manager of Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this paper. They would like to thank Mr. G. Knott for many helpful discussions and also the members of the Laboratory who have constructed and tested the experimental tubes.

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12. Appendix 1: Space Charge Waves on a Beam

A beam of electrons will diverge, because of space charge forces, unless constrained in some way. In a travelling-wave tube the beam is usually constrained by a longitudinal magnetic field. Although the following analysis assumes complete space charge neutralization, a similar result is obtained for the magnetically focused case if the cyclotron frequency is very different from the plasma frequency.

We think of a beam with a charge density ρ of positive charge and $-\rho$ of negative charge. We assume a neutral plasma in which the positive ions are heavy and do not move appreciably in the time of an oscillation. If the beam extends in the z direction, we also assume that all vector quantities have only z components. This assumption of zero radial field means, in effect, that all conductors are far removed from the beam because, if electron displacement occurs in the beam, image charges will appear on the conductors, and together with the charges in the beam, will produce radial fields.

In the unperturbed case the electric field E is zero. If a perturbation occurs along the z axis, E will have a component in that direction given in Rationalized M.K.S. units by

$$\frac{\partial^2 E}{\partial z^2} + \frac{2}{u_0} \frac{\partial^2 E}{\partial z \partial t} + \frac{1}{u_0^2} \frac{\partial^2 E}{\partial t^2} = \frac{\rho_0 \eta}{\epsilon u_0^2} E \quad \dots\dots(1)$$

Here u_0 is the average velocity of the electrons, ρ_0 the charge density, η the magnitude of the charge-to-mass ratio of an electron and ϵ the dielectric constant.

This has solutions of the form

$$E = \hat{E} \exp j(\omega t - \beta z) \quad \dots\dots(2)$$

where $\beta = (\omega \pm \omega_p)/u_0$ and $\omega_p^2 = |\rho_0 \eta|/\epsilon$

Equation (2) indicates that a disturbance, of angular velocity ω , will propagate in both directions with velocity relative to the stream

of approximately $(\omega_p/\omega)u_0$. For a full discussion of space charge waves reference is made to work by Ramo.⁶

We have assumed that vector quantities only have z components. If the diameter is finite and there are transverse fields, analysis shows that the value of ω_p is reduced. It is usual to define an effective radian plasma frequency, ω_q , which will depend on the beam and shield diameters.⁷

The waves on the circuit (in this case a helix) will satisfy the equation

$$\frac{\partial^2 E}{\partial z^2} - \frac{1}{v_0^2} \frac{\partial^2 E}{\partial t^2} = 0 \quad \dots\dots(3)$$

with solutions

$$E = \hat{E} \exp j(\omega t \pm \omega z/v_0) \quad \dots\dots(4)$$

where v_0 is the velocity of the wave in the axial direction. Here we have again neglected losses.

We rewrite equation (1) as

$$\frac{\partial^2 E}{\partial z^2} + \frac{2}{u_0} \frac{\partial^2 E}{\partial z \partial t} + \frac{1}{u_0^2} \frac{\partial^2 E}{\partial t^2} = -\left(\frac{\omega_q^2}{u_0^2}\right) E \quad \dots\dots(5)$$

for the waves on the beam.

Now, coupling will exist between the beam and the circuit, and if we introduce a coupling constant K , we obtain two simultaneous equations,

$$\frac{\partial^2 E_1}{\partial z^2} + \frac{2}{u_0} \frac{\partial^2 E_1}{\partial z \partial t} + \frac{1}{u_0^2} \frac{\partial^2 E_1}{\partial t^2} = -\left(\frac{\omega_q^2}{u_0^2}\right)(E_1 + KE_2) \quad \dots\dots(6)$$

and

$$\frac{\partial^2 E_2}{\partial z^2} - \frac{1}{v_0^2} \frac{\partial^2 E_2}{\partial t^2} = -\left(\frac{\omega^2}{v_0^2}\right)KE_1$$

If we try a solution $E = \hat{E} \exp(j\omega t - \Gamma z)$, we have

$$\Gamma^2 - 2j \frac{\omega \Gamma}{u_0} - \frac{\omega^2}{u_0^2} = -\frac{\omega_q^2}{u_0^2}(1 + K\hat{E}_2/\hat{E}_1)$$

and
$$\Gamma^2 + \frac{\omega^2}{v_0^2} = -\left(\frac{\omega^2}{v_0^2}\right)K\hat{E}_1/\hat{E}_2$$

Eliminating \hat{E}_2/\hat{E}_1 we obtain

$$\left(\Gamma - j \frac{(\omega + \omega_q)}{u_0}\right) \left(\Gamma - j \frac{(\omega - \omega_q)}{u_0}\right) \left(\Gamma + j \frac{\omega}{v_0}\right) \left(\Gamma - j \frac{\omega}{v_0}\right) = \left(\frac{\omega^2}{v_0^2}\right) \left(\frac{\omega_q^2}{u_0^2}\right) K^2 \quad \dots\dots(7)$$

This is a quartic in Γ and the four values of Γ correspond to four possible waves. We see that if the coupling is zero the four waves obtained are the two for the beam as solutions of equation (5) together with two for the circuit as solutions of equation (3). We are interested in a solution of equation (7) which has a positive real part to account for the gain.

For a discussion of the solution of equation (7) with small but finite coupling leading to an expression for the gain in the form $G = A + BCN$, the reader is referred to Pierce's solution of his equation (2.23).³

We arrived at equation (7) by a plausible argument. For a full treatment of mode coupling, where the electron stream and circuit are treated as coupled transmission lines, reference should be made to Pierce⁸ and Heffner.⁹

13. Appendix 2: Boundary Conditions at a Potential Jump

The charge approaching a boundary is equal to the charge leaving it provided that there is no current drain at the boundary. Hence $J_1 = J_2$, where subscript 1 refers to before the jump and subscript 2 to after the jump. In addition we have, for longitudinal fields and an infinite beam, $\nabla \times H = 0$.

Therefore $\nabla \times H - \partial D / \partial t = J$
 becomes $\partial D / \partial t = -J$
 Therefore $\epsilon \partial E_1 / \partial t = \epsilon \partial E_2 / \partial t$

From this it follows that $E_1 = E_2$ (8)
 which is the first boundary condition.

A second condition may be deduced from kinematic considerations. We shall use subscript 0 for average d.c. quantities, with a further subscript to distinguish values before and after the potential jump. Conservation of energy gives

$$\frac{1}{2}m(u_{01} + \hat{u}_1 \sin \omega t)^2 = \frac{1}{2}m(u_{02} + \hat{u}_2 \sin \omega t)^2 - e(V_1 - V_2)$$

if the gap is short. Here the u_0 and the u are d.c. and a.c. velocities, m the mass of an electron and the symbol $\hat{}$ is used for the moduli of the a.c. quantities. V_1 and V_2 are the d.c. voltages on each side of the jump.

Expanding and neglecting terms in $u^2 \sin^2 \omega t$ we have

$$\frac{1}{2}m(u_{01}^2 + 2u_{01}\hat{u}_1 \sin \omega t) = \frac{1}{2}m(u_{02}^2 + 2u_{02}\hat{u}_2 \sin \omega t) - e(V_1 - V_2)$$

Equating coefficients of $\sin \omega t$

$$u_{01}\hat{u}_1 = u_{02}\hat{u}_2$$

If the fields are wholly longitudinal as before

$$\epsilon \partial E / \partial t = -J = -\rho_0 u - u_0 \rho$$

also $\epsilon u_0 \partial E / \partial z = u_0 \rho$

Therefore $\epsilon dE / dt = \epsilon(\partial E / \partial t + u_0 \partial E / \partial z) = -\rho_0 u$

Therefore $u_0^2 dE / dt = -u_0^2 \rho_0 u = -u_0 u J_0$

We have already stated that the current is continuous, and shown that $u_0 u$ is continuous. Hence $u_0^2 dE / dt$ is continuous or, considering the potential, V ,

$$V_1 dE_1 / dt = V_2 dE_2 / dt \dots\dots\dots(9)$$

which is the second boundary condition.

13.1. Interference standing waves

The two waves obtained as possible solutions of the differential equation for the beam in Appendix 1 were

$$E_1 = \hat{E}_1 \exp [j(\omega t - (\omega + \omega_q)z / u_0)]$$

and

$$E_2 = \hat{E}_2 \exp [j(\omega t - (\omega - \omega_q)z / u_0)]$$

When these waves are equally excited so that

$$\hat{E}_1 = \hat{E}_2 = \hat{E}$$

then

$$E = E_1 + E_2 = 2\hat{E} \cos(\omega_q z / u_0) \exp [j\omega(t - z / u_0)]$$

E will be zero at $\omega_q z / u_0 = \pi / 2, 3\pi / 2$ etc., and E will be a maximum at $\omega_q z / u_0 = 0, \pi, 2\pi$ etc. It will be noticed that the group velocity of the resulting wave is equal to that of the beam velocity u_0 .

13.2. Effect of potential jump on plasma waves

Let the incident wave at the jump be

$$E_1 = \hat{E}_1 \cos(\omega_{q1} z / u_{01}) \exp [j\omega(t - z / u_{01})]$$

If the transmitted wave is

$$E_2 = \hat{E}_2 \cos(\omega_{q2} z / u_{02} + \theta) \exp [j(\omega(t - z / u_{02}) + \theta)]$$

then, from the boundary conditions (8) and (9),

$$E_1 = E_2$$

and

$$V_1 dE_1 / dt = V_2 dE_2 / dt$$

Thus

$$\hat{E}_1 \cos(\omega_{q1}z/u_{01}) \exp[j\omega(t-z/u_{01})]$$

$$= \hat{E}_2 \cos(\omega_{q2}z/u_{02} + \varphi) \exp[j\omega(t-z/u_{02} + \theta)]$$

and $\omega_{q1} V_1 \hat{E}_1 \sin(\omega_{q1}z/u_{01}) \exp[j\omega(t-z/u_{01})]$
 $= \omega_{q2} V_2 \hat{E}_2 \sin(\omega_{q2}z/u_{02} + \varphi) \exp[j\omega(t-z/u_{02} + \theta)]$
 or, considering the envelope of the standing plasma-wave,

$$\hat{E}_1^2 \cos^2(\omega_{q2}z/u_{01}) = \hat{E}_2^2 \cos^2(\omega_{q2}z/u_{02} + \varphi)$$

and $\omega_{q1}^2 V_1 \hat{E}_1^2 \sin^2(\omega_{q1}z/u_{01})$
 $= \omega_{q2}^2 V_2 \hat{E}_2^2 \sin^2(\omega_{q2}z/u_{02} + \varphi)$

Therefore

$$\hat{E}_2^2 / \hat{E}_1^2 = \cos^2(\omega_{q1}z/u_{01}) +$$

$$+ (\omega_{q1}^2 / \omega_{q2}^2) (V_1^2 / V_2^2) \sin^2(\omega_{q1}z/u_{01})$$

However

$$\omega_{p1}^2 / \omega_{p2}^2 = u_{02} / u_{01} = (V_2 / V_1)^{1/2}$$

Therefore

$$\left(\frac{\hat{E}_2}{\hat{E}_1}\right)^2 = \cos^2\left(\frac{\omega_{q1}z}{u_{01}}\right) +$$

$$+ \frac{(\omega_{q1} / \omega_{p1})^2}{(\omega_{q2} / \omega_{p2})^2} \left(\frac{V_1}{V_2}\right)^{3/2} \sin^2\left(\frac{\omega_{q1}z}{u_{01}}\right)$$

Hence if $r = \omega_q / \omega_p$

$$\left(\frac{\hat{E}_2}{\hat{E}_1}\right)^2 = \cos^2(\omega_{q1}z/u_{01}) +$$

$$+ (r_1/r_2)^2 (V_1/V_2)^{3/2} \sin^2(\omega_{q1}z/u_{01})$$

Here z is measured from a field maximum or a velocity minimum. Values of z and of V_1/V_2 can be chosen so that \hat{E}_2 is reduced compared with \hat{E}_1 .

This paper was one of a number presented at the Scottish Section's Symposium on "Industrial and Other Applications of Electronics," which was held in Glasgow and Edinburgh during March, 1956.

The Papers Committee invites the submission, for possible publication, of written contributions discussing papers published in the *Journal*, whether or not they have been read before meetings.

of current interest . . .

V.H.F. Broadcasting in the Midlands

The B.B.C.'s v.h.f. sound broadcasting station at Sutton Coldfield, near Birmingham, started test transmissions on 20th December. The station uses the same mast and adjoining buildings at the Midlands television station and it will transmit the Midland Home Service on 92.7 Mc/s, the Light Programme on 88.3 Mc/s and the Third Programme on 90.5 Mc/s. As at other B.B.C. v.h.f. sound broadcasting stations, the transmissions will be horizontally polarized.

The area in which satisfactory reception is expected under normal service conditions has a population of almost 7½ million people. It includes the whole of the counties of Warwick, Worcester, Leicester and Staffordshire; most of Derbyshire, Nottinghamshire and Shropshire, and parts of adjoining counties in Wales and the East and South Midlands.

Plans for the Expansion of Reed's School

Plans for enlarging Reed's School, Cobham, were announced by the Chairman of the Governors, Mr. D. S. Stacey, at the 143rd annual festival dinner held in London recently. A new block of classrooms and dormitories for another 100 boys is to be built, bringing the total number of boarders up to 250.

Mr. Stacey said that a new Science block had already been built which would enable the School to take all boys up to University entrance level in Science. The Trustees of the Institution's Benevolent Fund have provided the Physics Laboratory with equipment for conducting experiments in vacuum work.

Courses on Noise at Southampton University

The Aeronautical Engineering Department of the University of Southampton plans to hold a course from April 1st-6th, 1957, on Acoustics. The course is intended chiefly for the aeronautical industry, but there is adequate material to assist all those who have to deal with the increasingly important problems of noise.

Emphasis is laid on the mechanism of noise generation, in order to give a thorough understanding of the problem. Together with a general background on the subject, there will be a series of lectures on advanced measurement technique, and industrial noise control methods will receive special attention. Further details may be obtained from Mr. D. M. A. Mercer, Physics Department, The University, Southampton.

I.E.S. Trotter-Paterson Lecture

The Fourth Trotter-Paterson Memorial Lecture of the Illuminating Engineering Society will be given on Monday, February 11th, next, at the Royal Institution, Albemarle Street, London, W.1, at 6 p.m.

The lecture will be given by Sir Lawrence Bragg, F.R.S., Director of the Davy-Faraday Laboratory at the Royal Institution, and will be entitled, "Some Experiments on the Interference of Waves." Members of the Brit.I.R.E. are invited by the Council of the Society to be present and those wishing to attend should apply to 9, Bedford Square, London, W.C.1.

Courses for Industry in Nuclear Engineering

New opportunities for industry to acquire basic knowledge of nuclear energy are offered by a special course which has been arranged at the Harwell Reactor School. From 3rd June to 12th July, 1957, a six weeks' course is to be run entirely separately from the present standard course on reactor physics and engineering, and some sixty places will be available. Besides covering the basic principles of the physics and engineering of reactors, lectures will be given on such subjects as reactor metallurgy, shielding and health physics. In addition, emphasis will be placed on special topics such as the instrumentation of reactors and the use of radio-isotopes in industry. The latter topic is not included in the standard Reactor School course.

This special course is designed primarily for people of degree standard who are not directly concerned with overall reactor design. It should appeal especially to members of firms whose interest lies in the making of ancillary equipment for reactors. Application forms for places may be had on request from the Reactor School, A.E.R.E., Harwell, Berkshire. The students who are to participate will be selected at the beginning of March, 1957.

Films on Automation

The December issue of *Scientific Film Review*, the bi-monthly journal of the Scientific Film Association, contains details of 80 films on various aspects of Automation. This is the first list of its kind and should prove a useful work of reference for managements concerned with either the technical or human aspects of the subject. Copies, price 3s. 9d. each, are obtainable from the Association, 164 Shaftesbury Ave., London, W.C.2.

. . . Radio Engineering Overseas

621.314.63.012.8

Theoretical consideration on the experimental data of equivalent circuits of semi-conductor diodes at high current densities. W. GUGGENBUHL. *Archiv der Elektrischen Übertragung*, 10, pp. 483-485, November 1956.

The a.c. characteristic of a *p-n* junction diode has an inductive part when the diode is operated at high current densities. This effect cannot be explained by Shockley's theory of junction diodes. Starting from a recent theory of the d.c. performance of a junction diode at high injection levels by Herlet, the transient response of a diode in the high-current region is discussed. Some theoretical results are compared with previous experimental data and satisfactory agreement is found.

621.317.789.029.64 : 621.372.832.43

A microwave power monitor. J. SWIFT. *Proc. Instn Radio Engrs Aust.*, 17, pp. 424-428, December 1956.

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

621.319.4

Metallized paper capacitors. A. KOHLER. *Onde Electrique*, 36, pp. 842-851, October, 1956.

The conducting surfaces of metallized paper capacitors comprise a fine metallic film formed by vaporizing the metal *in vacuo*. The layer makes possible self-sealing in the case of breakdown, thus permitting a considerable reduction in specific volume compared with normal paper capacitors. The advantages of this capacitor are of special note when dealing with low voltages, when it is possible to use a single layer on the paper dielectric. The method of manufacture is described and the electrical characteristics discussed. The energy losses in the metallic layer are calculated and compared with those of ordinary paper capacitors.

621.319.43

Problems met in the manufacturing of variable capacitors. J. JOFFRE. *Onde Electrique*, 36, pp. 852-859, October 1956.

The author reviews the successive stages of variable capacitor development, with particular reference to commercial radio applications. He then proceeds to show how industrial-grade models were evolved from the commercial ones. Beginning with a definition of variable capacitor characteristics—law of variation, variable capacitance, residual capacitance—the author examines how the different construction factors effect their electromechanical characteristics and in particular, their microphonic susceptibility, temperature coefficient and resistance to corrosion. The paper includes a description of manufacturing control methods and, in particular, of the measurement of very slight capacitance drift.

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

621.37/8:789.9

Some prospects in the field of music. L. BERIO. *Elettronica*, 5, pp. 108-115, August/October, 1956.

The author distinguishes besides traditional vocal and instrumental music another kind: that of music electronically produced directly on magnetic tape without an interpreter. Several methods of doing this, and their effects, are illustrated in an l.p. record which accompanies this issue.

621.372.2

Measurement of the coupling impedances of delay lines. R. MULLER. *Archiv der Elektrischen Übertragung*, 10, pp. 424-428, October 1956.

Two methods of measuring the coupling impedance of delay lines are known. A dielectric rod is put into the electric field of the line and (a) the variation of the resonance frequency or (b) the variation of the propagation velocity, are determined. It is shown that the application of the known formulae is not justified under all circumstances. The correct relations and the limit of their application are presented. The theoretical and experimental results are compared and discussed.

621.373.5:537.228.1

On the concepts of resonance in piezo-electric crystal oscillators. G. BECKER. *Archiv der Elektrischen Übertragung*, 10, pp. 467-477, November 1956.

The paper discusses the relationship of series resonance and anti-resonance on piezoelectric crystal resonators with the physical resonances and shows that at these frequencies no physical quantity is at a maximum. The concepts of series resonance and anti-resonance are subsequently extended, the characteristic properties of series resonance and anti-resonance are outlined, and criteria are given which allow a clear discrimination to be made between series-resonance and anti-resonance oscillators.

621.385.16

Conditions for minimum noise figure in travelling-wave tubes. J. LABUS, R. LIEBSCHER and K. POSCHL. *Archiv der Elektrischen Übertragung*, 10, pp. 486-490, November 1956.

The theory of the travelling-wave tube forecasts a minimum for the noise figure; for its realization not all design parameters are fixed. This leaves a possibility for saving tube length by making the drift space before the interaction space as short as possible. The calculation is carried out for the practical case that two anodes with adjustable direct potentials are located between cathode and drift space.

621.385.16:621.317.74

A method of measuring reflections and stability in high-gain travelling-wave tubes. W. KLEIN. *Archiv der Elektrischen Übertragung*, 10, pp. 477-484, November 1956.

A report is given on a method of determining inner reflections on high-gain travelling-wave tubes in a manner deviating from the customary methods. This measuring method allows a separation of inner and outer reflections and a determination of the place of these irregularities. With this method the stability of a tube can also be checked in a very simple manner. The method is useful not only in developing the tubes, but might also be employed to advantage for checking the quality of the tubes or the entire travelling-wave tube amplifier.

621.373.444:621.387

Cold-cathode trigger tubes. C. H. TOSSWILL. *Philips' Technical Review*, 18, pp. 128-141, October 1956.

After a brief introduction dealing with the operation and properties of cold-cathode trigger tubes, work is described on the development of two special trigger tubes for use in a radiation monitor. Two of the measures taken to improve reliability and performance are dealt with: the use of molybdenum sputtering to achieve better stability and the provision of a priming current to eliminate the statistical lag. Both tubes are designed to work in self-quenching circuits. The stabilizer tube, a tetrode with an auxiliary cathode, is designed for providing a stabilizing supply for Geiger tubes. The circuit conditions for satisfactory operation are described. The advantages of this stabilizer are its voltage range (200-300 volts), its current range (5-150 microamps), and the fact that the stabilized output voltage can readily be adjusted by changing the values of two resistors. The rate meter tube is designed to handle Geiger tube pulses. It is also a tetrode but differs from the stabilizer tube in that its priming electrode, again an auxiliary cathode, lies in the path of the main discharge and provides a "direct" priming discharge. The special characteristic of this tube is its extremely high sensitivity; in the negative pulse circuit, the input charge sensitivity is 3×10^{-11} coulomb. This tube can also be used for voltage stabilization.

621.395.623:681.84.081.48

A magneto-dynamic gramophone pick-up. I. Construction. N. WITENBURG. *Philips' Technical Review*, 18, pp. 101-109, October 1956.

The new magneto-dynamic pick-up has a small rod-shaped magnet as its moving system; the rod is magnetized perpendicularly to its axis, about which it can turn, and is mounted between the ends of a yoke of magnetically soft material. A needle arm is fixed to the rod magnet whereby the lateral movement of the needle as it follows the groove in the record is converted into an angular movement of the magnet. An alternating flux is thus produced in the yoke, giving rise to a signal voltage in the coils wound on it. The angular movement of the magnet is provided for by an upper flexible bush of rubber and a bearing at the lower end of polyvinyl chloride. The rubber bush gives the magnet a restoring couple and the p.v.c. bearing provides the necessary damping against undesired resonances. The departure from

linearity between the angular movement of the magnet and the induced signal voltage is extremely slight. The sensitivity of the pick-up (ratio of r.m.s. voltage to peak needle velocity) is about 4 mV per cm/sec. A study of F_p , the force with which the needle presses on the disc, yields an optimum value for F_p of approximately 10 grammes weight in gramophones for domestic use, this value being sufficient to prevent de-tracking and giving minimum wear.

621.395.623.73

New investigation of "son rauque" in loudspeakers. J. B. FISCHER. *Archiv der Elektrischen Übertragung*, 10, pp. 441-454, October 1956.

The paper shows that the subharmonic excitation occurring with single-tone testing of dynamic loudspeakers with cone diaphragms can be tolerated to a certain extent without degrading the quality of reproduction. A new interpretation of subharmonic excitation is subsequently given by measuring the oscillating mode of the loudspeaker diaphragm and a method is proposed which suppresses "son rauque" (or "hoarse sound") on loudspeakers with cone diaphragms.

621.395.625.3

The development of a miniature battery-operated tape recorder. W. R. NICHOLAS AND A. D. HILDYARD. *Proc. Instn Radio Engrs Aust.*, 17, pp. 367-372, November 1956.

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

621.396.677.6

The total error of Adcock direction-finding systems. K. BAUR. *Archiv der Elektrischen Übertragung*, 10, pp. 491-493, November 1956.

It is shown why the bearing error of an Adcock direction-finding system in the upper part of the frequency range must be many times above the value of the natural antenna system error expected from theory. The result of the calculation further provides compensating means for reducing the error below the present theoretical limiting value.

621.397.8:621.372.553

The black-white edge in single-sideband television transmission. D. BUNEMANN and W. HANDLER. *Archiv der Elektrischen Übertragung*, 10, pp. 457-466, November 1956.

The paper covers analytical and experimental investigations into single-sideband transmission of television signals. The quality of the picture can be checked by the reproduction of a step, e.g. a black-white edge. The three quantities: build-up time, overshoot, and streaking, characterize the build-up transient of the step signal and accordingly the quality of the picture. In the proposal concerning the European Gerber standard, the tolerances and their apportioning to the individual elements of the television system had been left open. The tolerances are established for drawing up a specification for television transmitters.