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THE SYLLABUS OF NATIONAL CERTIFICATE EXAMINATIONS

TECHNICAL colleges throughout the country have recently been informed by the Joint Committee for the Award of National Certificates and Diplomas in Electrical Engineering that their syllabuses are out of date.* The Committee suggests that too much time is devoted in the Ordinary National Certificate syllabus to a detailed study of d.c. machines, which is inappropriate for those students who wish to proceed to the Higher National Certificate in Electronics. For these students it is now recommended that 80 per cent. of the electrical content of the final year (S3) be devoted to general electrical theory, machines *and electronics*. The remaining 20 per cent. would be concerned with machines or electronics, *or* a combination of both.

The Institution has still not been invited to give evidence to, or to serve on, this Joint Committee. Nevertheless, the present proposals follow, albeit belatedly, some of the recommendations made by the Institution at a Ministry of Education conference held in 1945. The desirability of wider consultations on such important matters as syllabuses was supported by *Wireless World* in January 1958.† Few engineers in management will dispute the need for an increasing quantity of trainees.

So far as the Higher National Certificate is concerned, flexibility in technical college syllabuses and curricula has been permitted for some years. As a result, many colleges throughout the country have tailored their courses to enable students to satisfy the Institution's examination requirements. Some colleges have arranged courses specifically designed to meet a local need and the Institution's recommendations, and have submitted the course details to the Institution for assessment.

Many students who progress some distance along the National Certificate road are concerned to find that professional institutions radically change their exemption requirements in terms of National Certificates. Students find themselves aiming at a moving target—in sight at one moment and out of range the next. There are very good reasons for this, not the least being the phenomenal rate at which the science of radio and electronics has evolved. In addition, there have been many fresh applications of existing and new principles.

It has been established that the Ordinary and Higher National Certificates alone form the basis of a very suitable qualification for the senior technician. Additional endorsements are required for the professional engineer, and the number and extent of these endorsements is constantly being revised.

To give an example, a need in the radio and electronics industry is for engineers to have a greater knowledge of the principles of technical management. This might well form an endorsement subject.

The National Certificate, with appropriate specialization and endorsements, is likely to form an important source of recruitment for the radio and electronics industry and for the Institution for many years to come.

* Document 5200/394.

† Editorial: "Training Technologists," *Wireless World*, **64**, p. 1, January 1958, and "Quantity and Quality," *Wireless World*, **65**, p. 1, January 1959.

World Radio History

INSTITUTION NOTICES

Symposium on Radio Apprentice Training

The following papers will be given at the Symposium on the training of radio apprentices which is being arranged by the South Western Section of the Institution in conjunction with the Royal Air Force Radio Apprentices School, Locking, near Weston-super-Mare, on Wednesday, 7th October:—

"Royal Air Force aspects of radio apprentice training," by Sqdn. Ldr. D. J. Garland, B.Sc., Associate Member (Training Officer, Aircraft Apprentices No. 1 Radio School, R.A.F.).

"Industrial aspects of radio apprentice training," by A. J. Robb (Education and Training Officer, E.M.I. Electronics Ltd., Wells).

"Psychological aspects of training for radio engineers," by H. C. A. Hale (Applied Psychology Research Unit, Medical Research Council, Cambridge).

The Symposium will be under the chairmanship of the President of the Institution, Professor E. E. Zepler, and will commence at 10.30 a.m. Further information may be obtained from the Symposium Organizer, Flt. Lt. D. R. McCall, B.Sc., A.M.Brit.I.R.E., 27 O.M.Q., R.A.F., Locking, Weston-super-Mare, Somerset. Early application is advisable as accommodation is limited.

Wellington, New Zealand, Section Committee

The Annual General Meeting of the Wellington Section was held on July 15th last, when the following were elected to the Committee for the year 1959:—

B. A. Bernon (Associate Member), *Chairman.*
J. P. Carter, B.Sc., M.Sc. (Hons.), (Associate Member), *Vice-Chairman.*
R. E. Greene (Associate), *Honorary Secretary.*
K. G. Duncanson, B.Sc. (Associate Member).
G. I. Hitchcox (Member).
W. C. Lee, B.Sc. (Associate Member).
D. J. Logan (Associate Member).

Readership of the *Journal*

A review of the circulation of technical journals was recently made in *World's Press News*. This showed that for the first half of 1959 the average circulation per issue of the *Brit.I.R.E. Journal*, as certified by the Audit Bureau of Circulation, was 7,064.

The Central Office of Information conducted during 1956 a survey on behalf of the Department of Scientific and Industrial Research on the use of technical literature by technologists in the electrical and electronics industry. The report of the survey was published early this year, and showed that 1,082 individuals, spread over 127 industrial organizations, were interviewed.

During the course of the investigation into journal readership, one of the questions was "Could you say which journals/papers you would have regularly if you did not have more than three?" In reply to this question the *Brit.I.R.E. Journal* is shown as sharing, *without prompting*, the highest rating of periodicals in the radio and electronics field.

Commonwealth Technical Training Weeks

In his capacity as President of the City and Guilds of London Institute, H.R.H. The Prince Philip, Duke of Edinburgh, summoned a representative meeting to discuss his proposal to hold technical training weeks in 1961. Apprenticeship weeks have been held for some time in the State of Victoria, Australia, and His Royal Highness considered this scheme to be capable of extension throughout the Commonwealth.

The object of the scheme is to stimulate awareness of the responsibility of the community towards young people entering employment; to give young people the opportunity to learn of the training and education schemes available in industry and commerce; to stress the importance of schemes of induction and training; to give increased opportunity to young people to learn of the opportunities available to them and generally to enhance the status of craftsmen and technicians, to promote production and prosperity.

The scheme was warmly welcomed by those present, who included High Commissioners, Ministers of the Crown and representatives of industrial, commercial and educational bodies. It will be referred to the respective Commonwealth Governments, and a small panel is being set up to consider the implementation of the scheme.

World Radio History

1959 Convention Diary†

Saturday, 4th July

The day's issue of "Brit.I.R.E. Convention News" contained references to comments, both serious and light-hearted, made by the speakers at the banquet held the previous evening in Caius College. The Immediate Past President, Mr. George Marriott, was particularly nostalgic in stating that he had returned to his old college and at last had the opportunity of speaking from High Table.

It was recalled also that this was the first occasion on which the Institution had had the opportunity of hearing Professor Mott. The hope was expressed that he might be able, possibly in the course of a formal address, to develop the theme of his after-dinner speech.

Reference has already been made in the Convention Diary to the sterling work of Mr. V. J. Cooper as Chairman of the Convention Committee, and in his first speech at an Institution Dinner in reply to Mr. Marriott's Toast he revealed himself as a most humorous after dinner speaker.

Mr. Cooper combined his speech with proposing the toast of "The Guests." His remarks, as a graduate of London University, on the role of "provincial" universities evoked much amusement and his witty comments on the place of science and industry in broadcasting proved a much appreciated topical quip on the theme of the Convention.

The morning sessions on Saturday dealt with industrial and scientific applications of television, and equipments and techniques used in applying it to astronomy, the exploration of space, observation of nuclear reactors and radioactive materials and to X-ray image amplification, were described. Delegates found it particularly interesting to learn at the same time of the comparable work in these fields which had been carried out in the U.S.S.R. One of the Russian delegates presented a survey paper and took a prominent part in the discussion.

During the afternoon transmitting aspects of television were discussed, papers being given on microwave and cable links, aerials, and the design and operation of satellite transmitters in this country and the U.S.S.R. A last-minute addition to the programme was an account of the B.B.C.'s use of facsimile methods for sending film extracts over the trans-Atlantic telephone cable.

It was indicated in the timetable that the official programme of the Convention should end at 5 p.m. However, among matters which delegates wished to pursue further was the discussion on the d.c. component which had been started two days earlier, and this was continued for another hour by some thirty interested engineers.

Sunday, 5th July

The President and Officers of the Institution were joined by delegates at the Morning Service in the University Church of Great St. Mary, and were welcomed by the Vicar, the Rev. Canon F. J. Fison.

In his address Canon Fison referred to the lessons to be learned from the life and work of James Clerk Maxwell. Deprecating worship of the individual, the Vicar laid stress on the value of a man's work rather than the man himself. He referred to Clerk Maxwell's early family life and the influence which his father, a man of the Church, most certainly had upon him. In many of his writings—as had been brought out in some of the Memorial Lectures—Clerk Maxwell made evident his own feelings of insignificance in grappling with the bountiful gifts of nature. Perhaps it was because of his Christian humility that Clerk Maxwell did in fact become possessed of the power to unlock some of the greatest secrets of nature.

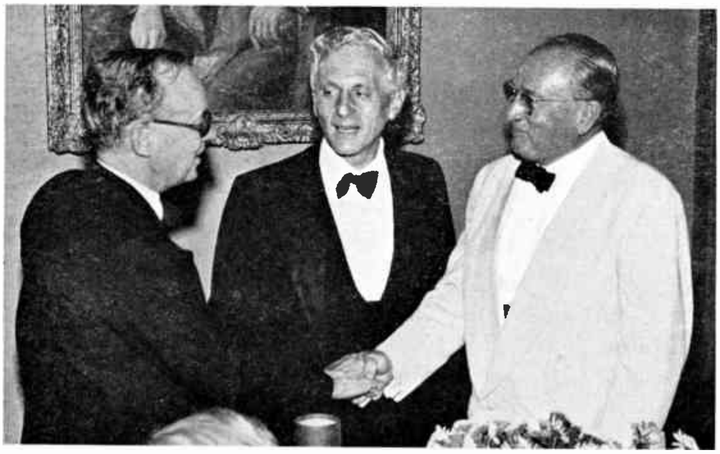
Upon such an inspiring address ended the Institution's fifth post-war Convention. It was certainly one of the most successful, and the entire proceedings demonstrated the even greater part that television will play in Science, Industry and Broadcasting.

† Convention Diary notes were published in the July and August issues of the *Journal*.

World Radio History

ELECTION OF DR. V. K. ZWORYKIN TO HONORARY MEMBERSHIP

THE announcement in the June *Journal* that the Council proposed the election of Dr. Vladimir Kosma Zworykin to Honorary Membership was received with considerable pleasure by the membership. It was felt to be especially appropriate that this honour should be bestowed on one of the pioneers of television at the Television Engineering Convention.



The President and the new Honorary Members after the Signing of the Roll. *Left to right*: Mr. E. K. Cole, Professor E. E. Zepler, Dr. V. K. Zworykin.

It is, of course, customary for such elections to be confirmed by members at a general meeting, and this formality was observed on the evening of Wednesday, 1st July, in the Cavendish Laboratory when Dr. Zworykin delivered his Clerk Maxwell Memorial Lecture "The Human Aspect of Engineering Progress."

In opening the meeting the President, Professor E. E. Zepler, said that among the highlights of a President's term of office, none surpassed the pleasure of taking the chair when a historical paper was being presented. He continued: "When our Institution took the initiative in establishing this Memorial Lecture, we laid down the principle that we would only invite speakers who had made a great contribution to the advancement of our profession. It is no exaggeration to say that the author of tonight's Lecture is, in fact, one of that small select band of men who practically founded the art and science of radio engineering."

Proposing a vote of thanks after the lecture, Professor Zepler said that the diverse themes of the lecture had demonstrated Dr. Zworykin's breadth of vision, and showed the enormous applicability of electronics techniques. He felt it was particularly stimulating to hear of the applications of these techniques for the benefit of mankind.

The President then invited those present to endorse by their applause the Council's recommendation that Dr. Zworykin be elected an Honorary Member of the Institution.

At the Banquet in Downing College on the following evening the President called on Mr. John L. Thompson, Vice-President, to read the citation on Dr. Zworykin's Certificate of Honorary Membership, as follows:—

"The development of Television Engineering starts with scientific work on photo-electric cells and cathode-ray tubes. In this field Vladimir Kosma Zworykin was a pioneer and it is internationally accepted that television cameras in the sense in which the name is now accepted became a possibility with the invention of the storage tube by Dr. Zworykin in 1923.

"These are but instances of the many services and contributions which Vladimir Kosma Zworykin has made in the application of scientific principles to practical engineering achievement.

"Typical of Dr. Zworykin's contributions to the advancement of Science for the benefit of mankind is his more recent work in promoting an international body for the further application of electronic science in the fields of biological research and diagnosis.

"The Institution pays tribute to the work of Vladimir Kosma Zworykin by electing him an Honorary Member. The presentation by the President of this Scroll of Honorary Membership followed delivery by Dr. Zworykin of the Fourth Clerk Maxwell Memorial Lecture in the Clerk Maxwell Lecture Theatre of the Cavendish Laboratories, Cambridge."

Dr. Zworykin then signed the Roll of Honorary Members and was formally presented with his Certificate.

World Radio History

THE FOURTH CLERK MAXWELL MEMORIAL LECTURE

"The Human Aspect of Engineering Progress" †

by

VLADIMIR K. ZWORYKIN, PH.D., D.SC., HONORARY MEMBER‡

Delivered on 1st July 1959 in the Cavendish Laboratory, Cambridge, during the Institution's 1959 Convention.

In the Chair: The President, Professor E. E. Zepler.

Introduction

IT is a great honour for me to address you, the members of the British Institution of Radio Engineers, on this occasion honouring the memory of James Clerk Maxwell at the scene of his activity in Cambridge.

We all recognize in Clerk Maxwell the great pioneer who laid the foundation for the modern science of communications with his mathematical conception of the electromagnetic field. But we also see in him a human benefactor whose sense of obligation to his fellows led him to undertake a great variety of tasks and to perform them with high ability and diligence.

In reflecting upon his many contributions to our modern heritage, it is this sense of obligation so amply demonstrated in the career of Clerk Maxwell that inspires my remarks to you today. This compulsion to utilize his skills and knowledge for the benefit of society must be the hallmark of every engineer and scientist in this age of technology. The opportunities for such service are limitless. Let me illustrate by sketching here today three widely separated areas in which the electronics engineer may make his most effective contribution. These are the fields of medicine, of public safety, and of public affairs. In the latter area, I shall suggest a technical step by which the engineer may help to strengthen the democratic process through broader application of the communications techniques to which Clerk Maxwell contributed so fundamentally.

Electronics in Medicine

The maintenance of health and the prevention or relief of suffering present a challenge to all of us. They present in particular a challenge to the electronics engineer, since he possesses specialized knowledge which can make a material contribution to the advancement of medicine. Up to the present, medical electronics has lagged far behind its potentialities. The reasons for this are easier to identify than to overcome. First of all, to contribute effectively to the field of medicine, the electronics engineer must overcome the barrier of language. Not only must he be proficient in his own field, he must know also something of medicine in order to communicate with medical men and life scientists and learn of their problems and difficulties. On the other side of the coin, the medical man or biological investigator must learn enough of electronics to collaborate effectively with electronics engineers and instrument designers. The gravity of this problem is widely recognized at present and some headway is being made in overcoming it by the creation of specialized institutes and curricula in medical engineering.

The second obstacle to progress in medical electronics is economic. New types of medical

† Manuscript received 1st July, 1959. (Address No. 18.)

‡ Director of the Medical Electronics Center, Rockefeller Institute of Medical Research, New York; Honorary Vice President, Radio Corporation of America.

U.D.C. No. 621.37/9: 612+656.1+342.847.

equipment find general acceptance only after exhaustive clinical tests. Consequently, there is a long delay between the original engineering investment and an adequate financial return from the sale of the finished article. The result is understandable reluctance on the part of instrument manufacturers to undertake the development of new types of medical apparatus. The obvious answer to this problem appears to be the assumption of development and testing costs by government or philanthropic sources.

However, let us emphasize the opportunities rather than the obstacles. We may conveniently distinguish here between the development of essentially new instruments for specific purposes, and the application of familiar electronic techniques to novel objectives in medicine.

The measurement of the extremely small potential variations which accompany vital processes within the body presents a direct opportunity for new instrumentation. Indeed, Einthoven's string galvanometer was developed expressly for the measurement of heart-muscle potentials or electrocardiography long before the advent of electronics. It has generally been replaced by instruments with electronic amplification which serve, at present, for the diagnosis of heart conditions as well as for the localization of brain tumours and the indication of epilepsy.

There can be little question that a further development of such instruments can provide much additional information. A most promising approach is the use of area display to replace the linear representation of the potential variation between selected points. Different types of area displays have been developed by Stanford Goldman and his associates in the Massachusetts Institute of Technology, by Harold Shipton at the Burden Institute, and by J. C. Lilly at the National Institutes of Health. In the last instance, a rectangular array of electrodes attached to the exposed cortex of a cat was connected with a similar array of indicator tubes, whose brightness was modulated by the potential of the corresponding electrode (Fig. 1). The indicator panel was then photographed by a high-speed motion picture camera and the film was reprojected at a lower speed. It was thus found that various stimuli produced characteristic sweeps of potential waves across the panel.

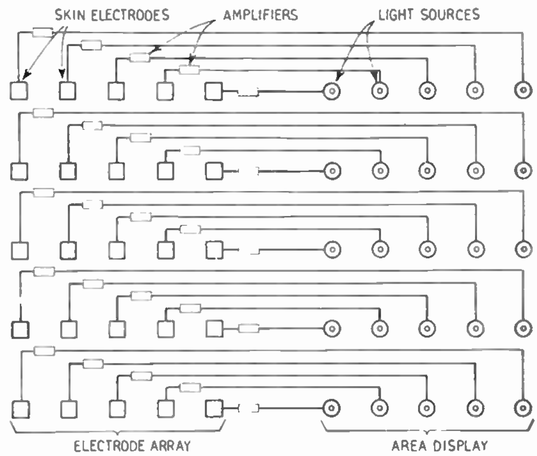


Fig. 1. Area display of skin potentials.

This technique, already valuable for a study of the nervous system and for the localization of defects in the brain in spite of its present limitations, clearly calls for further development, particularly by the use of new techniques made possible by solid-state technology.

Electronic techniques are not limited to a purely passive role in the investigation of our nervous system. As Wilder Penfield of the Montreal Neurological Institute has shown, the application of appropriate electrical stimuli to different parts of the cortex can cause recollection of remote memories. This represents one of the first steps in an experimental approach to the problem of memory storage. As another example of active control of the nervous system, we may cite non-convulsive electro-shock therapy for mental disorders and its outgrowth, sleep therapy. Both of these techniques appear to function by blocking normal cortical conduction. In the U.S.S.R. devices called "Electrosonne" are apparently in routine production and use for insomnia therapy.

Electronics, in fact, has gone much farther than this in influencing the functioning of the nervous system. It has been found that, in certain disease condition, relief can be obtained by the destruction of specific neural tissues. These may lie deep within the brain, at or beyond the limits of conventional surgery. In such cases, focused beams of ultrasonic waves have been found helpful in destroying tissue at the

World Radio History

beam focus without injury to the intervening layers. In the equipment developed by W. J. Fry of the University of Illinois, four 1-Mc/s sound beams are focused by polystyrene lenses at a carefully predetermined common point within the patient's skull, which is perforated to provide entrance ports for the beams (Fig. 2). This technique has been employed successfully in the treatment of Parkinson's disease.

These are a few examples of essentially new instruments which have been devised with specific medical or biological problems in view. There are many more instruments which employ familiar techniques but are novel in their application. An example is the radio pill which was developed co-operatively by the Medical Electronics Center of the Rockefeller Institute,

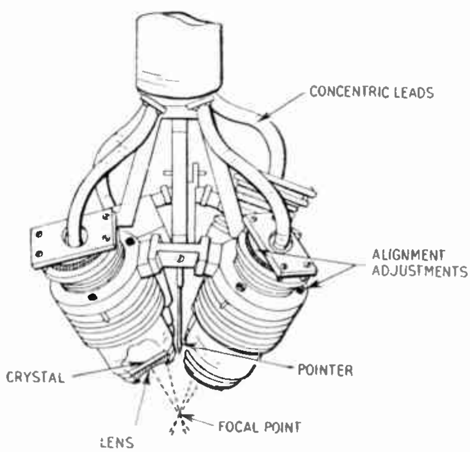


Fig. 2. Ultrasonic focusing irradiator after W. J. Fry (W. Welkowitz, "Ultrasonics in medicine and dentistry," *Proc. Inst. Radio Engrs.* **45**, p. 1059, 1957).

Dr. J. T. Farrar of the Veterans Administration Hospital in New York, and R.C.A.'s Industrial Electronic Products Division (Fig. 3). It is essentially a minute transistor oscillator whose frequency is modulated by small pressure changes and whose leakage field is detected by a frequency-modulation detector at a distance of the order of a foot away. The pressure-sensitive element is a flexible diaphragm supporting the armature of the circuit inductance.

All these components are quite conventional. The special problem to be solved before the

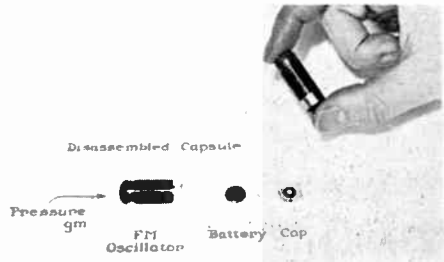


Fig. 3. The radio pill.

radio pill could become a practical tool for the study of the gastro-intestinal tract was to encapsulate the oscillator in a passive envelope small enough to be swallowed without discomfort. This envelope had to include a power source sufficient to maintain oscillations for the time of passage through the tract—from one to several days. The actual pill is about 1 in. in length and 0·4 in. in diameter. The firm plastic envelope is perforated and covered by a membrane at one side to communicate pressure variations to the interior. Power is provided by a 1·2 volt rechargeable cadmium battery. The pressure variations are registered by a recording galvanometer attached to the frequency modulation detector. Studies on 28 patients have shown that the radio pill does not affect digestive processes and is not a source of discomfort. It can thus be employed for patients too ill to permit the use of more conventional techniques for the study of the tract, such as gastroscopy.

Another new medical technique relying upon familiar engineering principles is the employment of pacemaker pulses to overcome heart standstill. If the heart condition resulting in standstill is momentary, it is generally possible to restore regular heartbeat by the external application of pacemaker electrical pulses with the normal heart rhythm. This is no longer practical when the condition is so chronic that stimulation has to be continued for a long time, since this results in the undesirable stimulation of other muscle tissue. At the same time, direct insertion in the heart muscle of exciting electrodes from an external pulse generator has proved unsatisfactory because the penetration of the electrodes through the skin creates a source of irritation and infection.

World Radio History

The simple solution has been the surgical insertion of the secondary of a pulse transformer under the skin supplying stimulation to the heart muscle, with an external primary connected to the power source (Fig. 4). No doubt

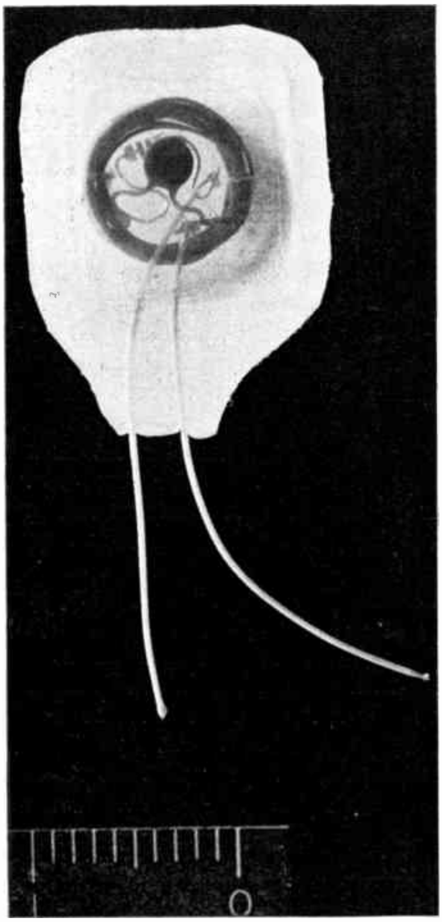


Fig. 4. Heart muscle stimulator. The part shown is inserted surgically and coupled electromagnetically to the external power source. (*Courtesy of Dr. Alexander Mauro.*)

the application of transistor techniques and compact batteries may lead to portability of the entire pulse generating system, so that the patient may move about freely.

Other familiar electronic techniques have greatly reduced the possible injury received from excessive X-ray exposure in fluoroscopic examinations. In conventional fluoroscopy the light emitted by the X-ray screen is so inefficiently utilized by the observer's eye that the discrimination of sufficient detail may demand exposure to levels of radiation a thousand times more intensive than those required for X-ray photography. Electronic amplification can raise the brightness level of the observed image by a comparable factor and thus reduce the necessary X-ray exposure of the patient to relatively safe levels.

Present methods of electronic image amplification are largely borrowed from television practice. In fact, one of the most flexible techniques employs simply a very large-aperture lens to image the conventional fluoroscopic screen on the photocathode of a high-sensitivity image orthicon camera. The image is viewed on the screen of a television monitor at a light level which is essentially independent of the screen brightness. In another system, the X-rays impinge on a fluorescent screen in intimate contact with the photocathode of an image tube and the photoelectrons are accelerated and focused on a small fluorescent screen. The small screen is then either observed directly with a magnifier, or indirectly with a compact television chain. The acceleration of the electrons and the size reduction of the image combine to yield an increase in brightness of the viewing screen over the X-ray screen approaching a thousand.

Still another system, which is perhaps the most interesting but the least developed of the three, employs a photoconductive layer in series with an electroluminescent film (Fig. 5). Alternating voltages are applied across a sandwich of the two films, and the X-ray pattern is projected on the photoconductor. Light is emitted by those portions of the electroluminescent layer

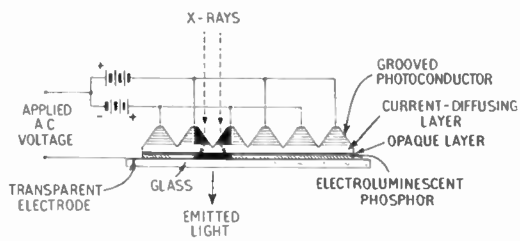


Fig. 5. Electroluminescent panel X-ray amplifier (B. Kazan. "A solid-state fluoroscope screen," *R.C.A. Review*, **19**, p. 19, 1958).

World Radio History

which are backed by photoconductor subjected to X-ray radiation; the X-rays increase the conductivity of the photoconductor, effectively shorting it out and increasing the alternating fields across the electroluminescent layer. Since both the photoconductor and the electroluminescent layer have highly non-linear electrical characteristics, the X-ray panel amplifier delivers pictures with high contrast. Brightness gains better than 100 have been achieved at low

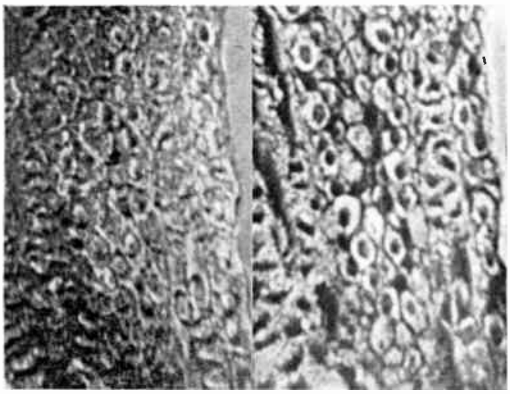


Fig. 6. Identical sections of kidney tissue, photographed with television microscope with 4000 Å and 2537 Å illumination, respectively.

radiation levels. However, before the X-ray panel amplifier can become a practical device replacing the conventional X-ray screen, improvements must be made in speed of response and in graininess of the image.

Television and the image tube have contributed materially to the effectiveness of the microscope—the chief research tool of the biological scientist. One of the chief difficulties in the study of organic preparations at high magnifications is the fact that most organic materials are equally transparent in thin sections making discrimination very difficult. In the past, this difficulty has generally been overcome by the use of specific stains which would combine chemically with different substances. Unfortunately, this technique generally proves lethal to living substance and may also cause structural changes. On the other hand, it is well known that substances which appear quite transparent in the visible commonly exhibit strong characteristic absorptions in the ultraviolet portion of the spectrum. An example is provided by identical sections of kidney tissue observed under ultraviolet illumination with the wavelength of 2537 Å and with visible illumination of 4000 Å (Fig. 6). The cell nuclei stand out clearly in the ultraviolet picture.

In the past, the advantages of ultraviolet illumination were largely nullified by the fact that photography was required to make the picture visible. Instantaneous observation with ultraviolet illumination became possible only with the introduction of ultraviolet-sensitive television cameras. Here the sensitive target of the camera tube replaced the photographic plate, and the visible picture was observed on a television monitor. The pictures of the kidney tissue were prepared with a system of this type. However, a much simpler solution of the problem was recently achieved with the introduction of the Ultrascope.

The Ultrascope consists essentially of an image tube (Fig. 7) and magnifier which, in an ultraviolet microscope, replaces the eyepiece of a conventional instrument. The image tube converts the ultraviolet image projected on its photocathode into a greenish-yellow image of

Fig. 7. Ultrascope image tube.

high brightness on its fluorescent screen. Thus, with this electronic device, microscopy in the ultraviolet becomes as simple as in the visible range of the spectrum.

The picture observed with the Ultrascope is, of course, in monochrome. Thus, while it does away with the necessity of staining with possible adverse effects on the specimen structure,

World Radio History

it does not permit the immediate differentiation of various components which might be achieved with a complex multiple staining technique. Differentiation by colour, without the drawbacks of staining, can, however, be achieved by colour translation, each of the three primary colours, red, blue, and green, is correlated with one of three selected ultraviolet wavelengths, for which the components of the specimen exhibit material differences in absorption. Ultraviolet images of the specimen formed with these three wavelengths are converted into red, blue, and green images, which are viewed in super-position. Thus, cyan, yellow, and magenta portions of the compound image immediately indicate the presence of substances which strongly absorb the ultraviolet radiations correlated with red, blue, and green respectively. Substances absorbing any two of the selected wavelengths will appear tinted with the primary colour corresponding to third. In brief, the colour of different substances is determined by their ultraviolet absorption spectrum, much as in conventional microscopy the colour is determined by the absorption spectrum in the visible.

Earlier efforts at practical ultraviolet colour translation employed photographic methods. They necessarily involved long time delays between specimen exposure and observation, and demanded large expenditure on materials. Instantaneous observation with colour translation became possible only with the application of television techniques (Fig. 8). Here, the specimen is illuminated in successive vertical fly-back periods with ultraviolet radiation of the three selected wavelengths. An ultraviolet-sensitive television camera receives the image of the specimen. The picture signals are channelled successively to the corresponding colour channels of a television monitor. The red, blue, and green pictures formed on the screen of the colour kinescope are fused by persistence of vision into a compound colour picture of the specimen.

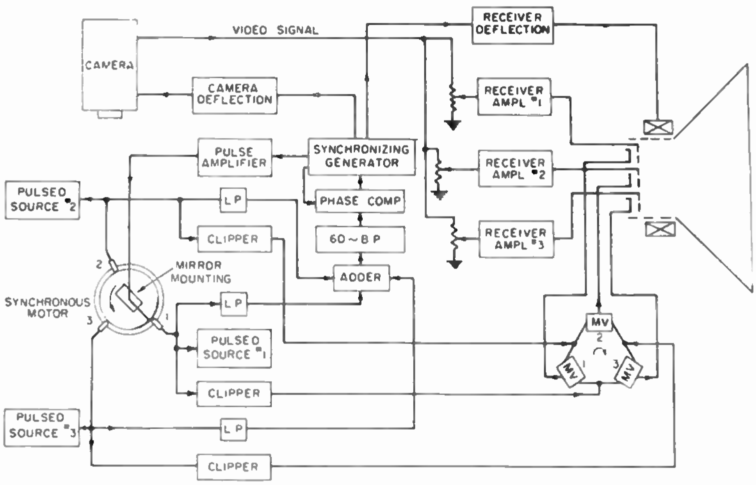


Fig. 8. Block diagram of ultraviolet colour translating television microscope.

In the instrument constructed at the Rockefeller Institute in New York, illumination is provided by three separate ultraviolet monochromators provided with thyratron-controlled mercury-arc sources. The radiation from the exit slits of the monochromators is directed into the substage of the ultraviolet microscope by a rotating 45-degree mirror. The striking of the mercury arcs, the vertical deflection of the television systems, and the gating of the three colour channels are all controlled by commutator contacts on the mirror mounting, which has a total rotation period of 1/20 sec. A high-sensitivity image orthicon with an ultra-violet-transmissive front end serves as camera tube. The monitor is a conventional colour television

World Radio History

receiver converted to field-sequential operation. A quantitative idea of the relative absorption of the different constituents of the specimen at the selected ultra-violet wavelengths may be obtained by applying the signals from the three colour channels to a line-selector oscilloscope. This gives a visible representation of the variation in signal amplitude along a particular scanning line.

The primary aim of the various forms of ultraviolet microscopy which have been described is the accentuation of contrast, in the form either of brightness differences or of colour differences, between different components of organic specimens. The increased resolution obtainable, in view of the somewhat shorter wavelengths of ultraviolet as compared with visible radiation, is generally a secondary consideration. Whenever very high resolution is required in biological investigations, another electronic instrument fills the gap, the electron microscope. While this instrument has, for a long time, been the object of independent development, it too, in its early stages, represented an application of techniques familiar from cathode-ray oscilloscope and television studies.

The examples given so far are drawn from problems in which we have made appreciable headway. There are many others where we are reasonably confident that electronics can make an essential contribution, although an effective approach has eluded us so far. An important problem of this type is the measurement of the intraocular pressure in a manner not painful to the patient. The recognition of small increases in this pressure is important for the recognition of early stages of glaucoma, a blinding disease endemic in an ageing population. Attempts to measure the effects of pressure increase optically or by the changes in the propagation time of ultrasonic pulses through the eyeball or along its periphery are under investigation. There is thus ample scope for new ideas and novel techniques.

There are, finally, a number of basic needs in the field of medicine which are immediately obvious to the engineer and which demand his co-operation in seeking a solution. One of these is the more efficient utilization of skilled hospital personnel and the reduction of hospital operating costs. Principles widely applied in industrial automation can make a material contribution here. Thus, an undue amount of the nurses' time is consumed at present by the routine of taking the patient's temperature and pulse. Furthermore, present methods of taking the temperature are quite unnecessarily slow. Replacement of the conventional thermometer by low heat-capacity thermistor probes could accelerate the process greatly. Furthermore, automatic temperature and pulse recording equipment employing punched cards would permit the use of untrained workers for carrying out this routine function. Automatic data transfer to the patient's record would both save time and eliminate error. These are just a few examples how the application of accepted engineering principles could contribute to increased economy and efficiency in hospital operation.

An even more urgent problem is presented by the enormously rapid increase in accumulated medical knowledge. It is becoming increasingly true that no physician can carry in his memory the vast store of knowledge which has bearing on the proper diagnosis of disease from the symptoms exhibited by a particular patient. What he needs is the possibility of recourse to a much larger memory containing the countless correlations between symptoms, disease, and therapy which are now scattered over the hospital records and the medical literature of the entire world. Modern computer technology has provided us with the vast, high-speed access memories which are required for such a purpose and our communication system is such as to permit a single memory store to serve physicians distributed over a great area. On the other hand, extensive research is required to determine the best method of coding the available information in computer language and in planning the computer logic so that it may provide the needed answers.

A small beginning in this direction has been made. A group of physicians at the New York and Mt. Sinai Hospitals, in co-operation with the Medical Electronics Center of the Rockefeller Institute and the R.C.A. Electronic Data Processing Division, have set up an experimental programme for the differential diagnosis of the hematological diseases, making use of available

World Radio History

hospital records. Ninety-eight different symptoms were employed as indicators for some 30 diseases. To obtain a diagnosis for a given hospital case, the observed symptoms were recorded on a tape and compared by the computer with the symptom listings for the several diseases, also recorded on a magnetic tape. On the basis of this comparison, the computer printed out a list of diseases consistent with the observed symptoms, as well as those untested symptoms which might facilitate a positive diagnosis. It was found that, in the great majority of test cases, a positive diagnosis or a relatively short list of possible alternative diseases was provided. It is expected that much further work will be done to refine and extend the procedure.

Electronic data processing has numerous applications in medicine apart from diagnosis. Some of these are the maintenance of standardized health records for the population, the charting of the origin and course of epidemics and the analysis of statistical data on the effect of new drugs and health measures. The possibility of obtaining needed information more quickly than heretofore through the use of electronic computer techniques is particularly important for the extension and the rapid application of medical knowledge.

Electronics in Transport

I shall now turn to an entirely different field which can properly claim the attention of the electronics engineer—transportation. To emphasize the need for electronic aids, let me quote some statistics relating to highway accidents in the United States. Granting material differences between the situations in the United States and the United Kingdom, I think that conditions are not so different that my conclusions will be without local application.

At the present time approximately 40,000 people are killed annually in automobile accidents in the United States. This is the population of a town half the size of Cambridge. The blotting out of life in a town of this size year after year should be cause enough for serious consideration of ways of reducing the holocaust. A survey of other effects of automobile accidents does not improve the picture. Thus, the number of those injured annually with disablement extending beyond the day of accident is 1½ million. The anguish of the victims of the accidents and their families defies measurement. However, to obtain some gauge of the amount which could properly be expended for the prevention of accidents we may turn to the direct economic loss resulting from automobile accidents. This is approximately 6,000 million dollars annually in the United States—a figure comparable to the entire fuel cost for operating the automobiles.

It is thus clear that there is ground for concern. However, it would be pointless for the electronics engineer to consider the matter further unless there is some indication that he may be able to improve the situation. To answer this question we must examine the accident statistics in greater detail.

The curve in Fig. 9 representing the total number of fatalities per 100 million passenger miles shows that the fatality rate has remained practically constant in the last four years. This may occasion surprise since, in this period, the

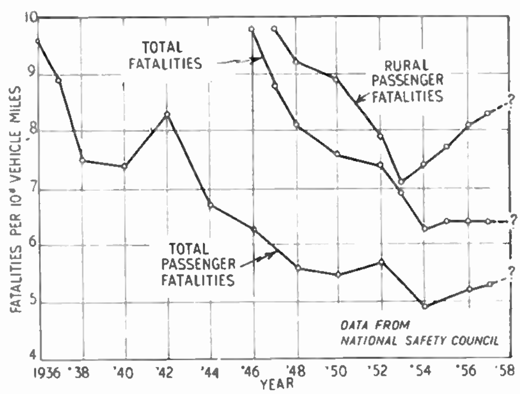


Fig. 9. Statistics of fatal highway accidents in the United States.

construction of divided and limited-access highways has greatly reduced the number of head-on collisions and crossing accidents, and engineering improvements in automobiles have rendered mechanical failure a minor factor in accident causation. A further breakdown shows that, in fact, pedestrian fatalities and fatalities on urban

World Radio History

roads have decreased. On the other hand, the curve representing rural passenger fatalities shows a material rise in the period in question.

A further examination of fatal rural accidents indicates that the accident distribution is quite different for turnpikes and for the road system in general. The figures shown in Table 1 refer to a particular turnpike system. It was found that, here, by far the commonest form of fatal accident was a rear-end collision: 40 per cent. of all fatal accidents on the turnpike were of this nature, whereas for the total road system only 7 per cent. were rear-end collisions.

Table 1

Fatal Rural Accidents in 1956

	Total	*Turnpike*
Left Road	40%	28%
Overturned	19%	13%
Struck object	15%	13%
Struck other car	26%	55%
Head on	10%	5%
Rear end	7%	40%
Sideswipe	8%	8%

The logical conclusion is that at the high speeds which are regarded as nominally safe on the improved highways, we are coming hard up against a human limitation, the reaction time of the driver. According to the American Automobile Association, an alert driver requires $\frac{3}{4}$ sec to move his foot from the accelerator to the brake. In this interval a car moving at 70 miles per hour moves 75 feet. The drowsiness and inattention induced by turnpike driving is likely to lengthen materially the reaction time of the driver and the distance which the car advances at undiminished speed upon a stopped or slowed vehicle ahead of it. The result: a rear-end collision. The only practical way of preventing accidents of this type consists in eliminating the driver's reaction time from the response of the car to an altered traffic situation ahead of it. The elimination of the human factor from a mechanical operation, which is here required, is precisely the type of problem which the electronics engineer encounters in the design of electronic controls for industrial automation. There is thus good reason to believe that his skills can make a material contribution to the reduction of the accident toll.

At the beginning of our attack on the problem, which was initiated over six years ago, we considered a broad range of means which might achieve the desired control. We soon came to the conclusion however, that individual devices attached to the car, such as anti-collision radar, ultrasonic devices, and infra-red detectors, would not accomplish our objective. Instead, the essential features of a satisfactory control system could be realized only with the primary installation in the road itself, under the skilled supervision of highway construction and maintenance personnel.

What are these essential features of the control system? They are:

(1) means for keeping the car in its lane;

(2) means for preventing the car from colliding with a vehicle ahead of it; and

(3) the possibility of introducing improvements in roadways and vehicles gradually, permitting cars with and without control equipment to use roads with and without control installations, without detracting from the safety of any vehicle.

Each of the first two features, guidance and collision prevention, implies the existence of a separate automatic control system. The third feature, which we may call compatibility, imposes conditions on the nature of these separate automatic control systems.

The problem of providing guidance is relatively simple. In the system devised by us it is solved by the insertion of a cable carrying high-frequency current in the centre of the traffic lane. Simple detectors mounted on the car then provide control signals to the steering mechanism which will keep the car centred on the lane. The assigning of different frequencies to different lanes provides, at the same time, a method for route selection.

Collision prevention is a more complex matter, since it involves the interaction of two vehicles with the road as intermediary. The road itself must be the agent for transmitting the signal from the car ahead, which may or may not be equipped with control devices, to the electronically controlled car under consideration, since the signal must be unaffected by curves and hills. Furthermore, the active source of the signal must be within the road since a

World Radio History

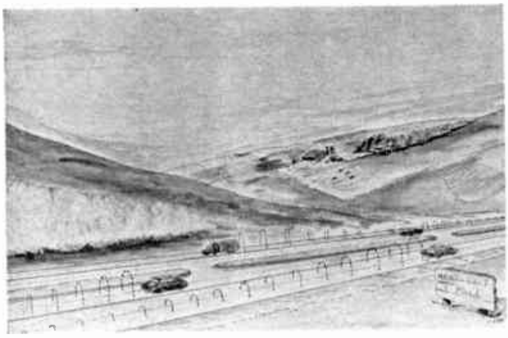


Fig. 10. Generation of flying-tail signal on the highway.

perfectly passive, unequipped vehicle must makes it presence and relative speed known to any equipped vehicle following it. The equipment of the controlled car must consist of a detector for the warning signal, and a servo system for adjusting the car speed in response to it.

The road installation for collision prevention has thus the joint function of detecting vehicles at any point along the road and of generating warning signals over a prescribed distance of road behind every detected vehicle. The detection is accomplished in our system by a sequence of wire loops embedded in the surface of the highway. The loops are about the size of a car and spaced a few feet apart. As a vehicle passes over any loop, the inductance of the loop is altered by the metal of the vehicle. This change is detected by an electronic circuit placed alongside of the road.

In response to the change in inductance of the loop a "flying tail" of electromagnetic warning signals, decreasing in amplitude with distance, is generated behind the vehic'e (Fig. 10.) This is received by a following car provided with the proper equipment and will act on its accelerator and brakes whenever the separation of the cars falls below a maximum permissible value established for the absolute and relative speeds.

The third feature of the system, compatibility, is best illustrated by visualizing three stages in its development. In the first of these only the roadway is modified. The vehicles are completely unequipped. Detectors in the roadway actuate warning lights on the side of the road which indicate to the driver the presence of a car ahead of him in his own lane or of other vehicles approaching in the opposite direction (Fig. 11). Such arrangements are of special value wherever visibility is obscured by a curve, a hill, or some other obstacle and may be equally useful in areas subject to fog or smog.

In the next stage, some or most of the vehicles are equipped with detectors for the warning or guidance signals, controlling visual or auditory indicators within the cars. The series of lights by the side of the road are now supplemented or replaced by the automatically generated flying tail of electromagnetic signals which has already been mentioned. Both detecting the vehicle ahead and maintaining the car centred in the lane remain possible even under conditions of extremely bad visibility.

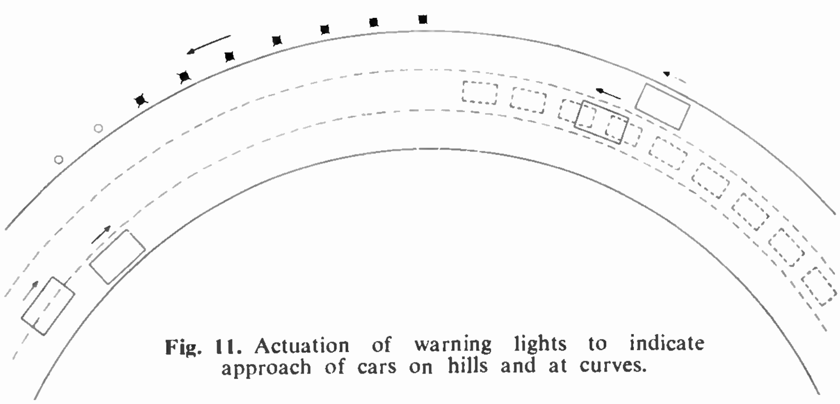


Fig. 11. Actuation of warning lights to indicate approach of cars on hills and at curves.

In both these stages the driver continues to exercise the essential function of controlling his

World Radio History

vehicle in response to information supplied by the control system. In the third, fully automatic stage, this function is assumed by servo systems acting on the steering and speed-control mechanisms, responding directly to the guidance and collision prevention signals. Thus, as long as the vehicle is within the electronically controlled road system, the driver is relieved of his responsibilities and the vehicle proceeds along a preset course. When the vehicle leaves the controlled road system, it shifts back from automatic to manual control. The situation is essentially similar to the piloting of long-range aircraft, where the autopilot may take over the controls except near take-off and landing.

In practice, the three stages of development which have been described may be expected to coexist for many years. Vehicles in all stages of automation may be expected to use the highways simultaneously. Compatibility has been built into the system in such fashion that this leads to no difficulties.

The general plan of the road installation for a single traffic lane is indicated in Fig. 12. The detector loops, which are its most prominent feature, may be cast with the concrete in new roads or inserted by a diamond-saw technique in the finished pavement. Phase changes in the loop current resulting from the presence of a car above the loop effect the opening of a gate in the control unit which applies a frequency modulation signal to the high-frequency currents continuously fed to a sequence of short wire antennas buried in the roadway. The modulation control signal is propagated rearward with exponentially decreasing intensity along a line of diodes, so that the modulation of the antenna currents shows a similar exponential decrease with distance from the car influencing the loop currents. This modulation constitutes the flying-tail signal. Following cars equipped with detectors can infer the distance of the vehicle ahead from the amplitude of the modulation. In a fully-controlled vehicle an electronic computer derives control signals for the acceleration and braking mechanism from the modulation amplitude, its rate of change with time, and the speed indication of the controlled car.

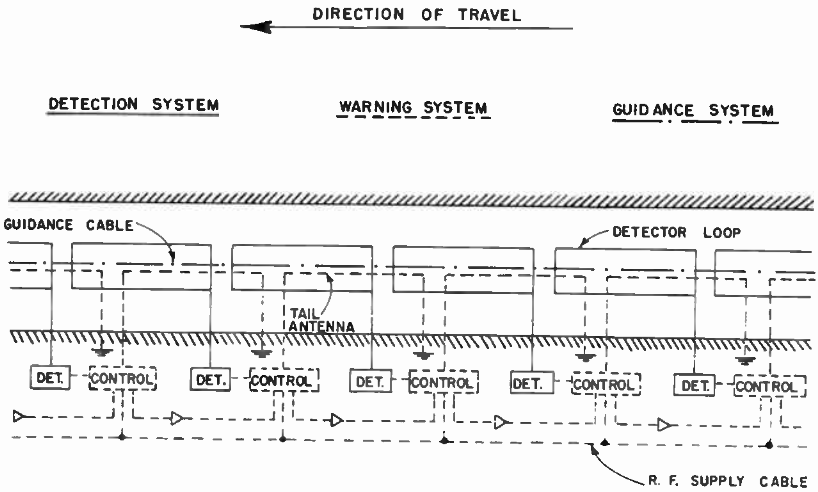


Fig. 12. Electronic control installation for one traffic lane.

Control signals for the steering mechanism of a controlled car are derived from a pair of tuned antennas mounted on the left and right of the front bumper. The difference in the current induced in the two antennas determines the correction applied to the orientation of the front wheels. Thus, with properly stabilized servo controls, the controlled car proceeds down the centre of the lane without wavering unless the driver takes deliberate action which supersedes the automatic guidance mechanism.

There are a number of incidental advantages obtained with the system as described. The cables in the road traversed by high-frequency currents provide a convenient communication system for the highway authorities and the driver. By means of it, the driver may be advised of adverse road or weather conditions, special events, facilities provided for his comfort, and other items contributing to his safety and enjoyment. Furthermore, the amplitude of the modulations constituting the flying-tail signal may be adjusted to fit the road conditions. When the roads are icy or wet it can be increased, so that the spacing between cars is increased automatically. In this manner the chances of pile-ups resulting from skids may be minimized. Finally, the detection of the cars by the loop circuits makes it possible for the traffic authorities to obtain a complete, instantaneous representation of traffic conditions over a wide area and to adjust routing instructions accordingly.

The individual components of the system also have their special uses. Thus, a pair of loops and detector circuits may be employed for measuring the speed of passing cars; such a system has been in operation on the entrance road to the R.C.A. Laboratories in Princeton for over a year, actuating a warning sign whenever a car exceeds the prescribed speed limit. Incidentally, experience with this system has shown that the transistor circuits employed in the detector unit function satisfactorily over a wide range of temperature and under all weather conditions.

Another use for the detector units is as traffic counters along intersecting roads, providing control information for traffic lights, so that the green-light periods for each road may be made proportional to the relative traffic flow density along that road. A more elaborate application is now being worked out for the Airways Modernization Board. The problem is here the most efficient routing of planes in airports, after landing and before take-off.

After this description of the traffic control system you will naturally ask how much has actually been accomplished in realizing and testing the system.

I have already had occasion to show you the detector units and to mention some tests to which they have been subjected. The first complete road installation of the system was demonstrated in the Fall of 1957, in co-operation with Mr. L. N. Ress, State Engineer, Department of Roads, State of Nebraska. The demonstration was carried out on a 300 foot section at a newly-constructed intersection in Lincoln, Nebraska. Test vehicles were provided with dashboard indicators for the guidance and anti-collision signal. With their aid, drivers were able to guide their cars along the centre of the lane, with speed adjusted to that of preceding vehicles, even when the windshield was completely obscured.

The guidance function has more recently been tested by the General Motors Corporation on a full-scale test track in Warren, Michigan, with regular Chevrolet cars whose power steering was controlled by signals derived from two antennas on the front of the car. The difference signal, determined by the car position, was applied along with a car speed signal from a tachometer generator and a signal indicating the orientation of the front wheels to a computer in the glove compartment. The output of the computer provided a command voltage corresponding to the proper front wheel angle to correct the car path error. The signal actuated the hydraulic-power positioning cylinder through an electro-hydraulic servo valve. Considerable information was gained from the tests with respect to the required accuracy of positioning of the guidance cable and its preferred form at approaches to curves.

We are prepared to extend tests of the various aspects of the system at the R.C.A. Laboratories in Princeton. On the other hand, it is clear that the satisfactory development of this or any other traffic control system will demand the close co-operation of highway construction engineers, the automobile manufacturers, the electronic industry, and other groups concerned with traffic problems. With this in view, the plan here outlined has been presented to numerous groups of engineers and traffic specialists, including the Highway Research Board in Washington and a special Conference called by the Safety Education Project of Columbia University in New York. The Conference recommended further studies and appointed a Representative Committee drawn from all interested groups to advance the programme. We hope that the interest evinced by the various groups which have been contacted may soon result in the

World Radio History

establishment of a large-scale test facility. Such a test facility could provide the background knowledge necessary for initial installations of a control system on our highways. In the meantime, I have been greatly encouraged by the parallel work undertaken in England by the Road Research Laboratory in Harmondsworth, Middlesex, under Mr. C. G. Giles, Head of Surface Characteristics Section.

In brief, having seen the need of measures to curtail traffic accidents and recognized that electronics could make a material contribution to such life-saving measures, we have worked out a plan for an electronic control system which appeared to hold promise for attaining our objective. More than that, we have designed and constructed essential portions of the electronic hardware and subjected them to such tests as were within our reach. Finally, to fulfill our social obligation in this matter, we have sought the co-operation of other groups needed to effect the practical realization of the plan.

Thus, while electronic control of motor vehicles is still far from being realized in practice, we can at least look back on a considerable period of experimentation and testing and of discussion with other groups concerned with the problem.

Electronics in Public Affairs

The last subject which I wish to present as an example of the application of the engineer's sense of obligation toward society is still entirely a vision for the future. Nevertheless, engineering principles involved are so straightforward and accepted that I have little hesitancy in presenting it as a practical project. Finally, it seems particularly appropriate for this occasion, since it concerns the subject of radio communication, for which Clerk Maxwell established the theoretical basis, and the telephone, the theme of his last public lecture.

The essential objective of the project is to make democratic processes work more effectively in the modern world. If we look back to the beginnings of democracy, in the city states of Greece, we find the entire citizenry coming together to make joint decisions on the issues which faced their city. The citizens of Athens did not congregate in the Agora just to hear speeches of their leaders, but to express their will through the vote. Such direct democratic procedures have survived only in isolated instances, such as the town meetings of New England and Switzerland, and their influence is limited to purely local matters. The major decisions, affecting the welfare of the entire nation, are made by bodies with delegated authority; the electorate as a whole has an opportunity to express its will only at rare intervals and then in a manner which lumps issues and estimates of personality. The public will on any one issue remains subject to arbitrary interpretation.

It is often said, with some justification, that the giant nation states of today—and, indeed, world government—are made possible by our advanced techniques in communication. Indeed, these techniques make it possible to inform the entire population almost instantly of important events going on anywhere in the world. Furthermore, they permit governments, which control the media of communication directly, or indirectly, to transmit orders and instructions to everyone under their sovereignty within and beyond the nation's boundary. On the other hand, they have not given the people, in whom theoretically the ultimate power resides, a direct controlling voice over the actions of their government. The manner in which popular control is exercised today in our representative democracies dates back to the days when it was impossible for the ordinary citizen to keep abreast of events. The best he could do under such circumstances was to select, along with his neighbours, a man whose general good judgment he trusted to speak for him at the distant seat of government. There was, and is, no continually effective check on the actions of a representative so chosen until he comes up for re-election and, indeed, no official way in which the representative can gauge the will of his constituents.

This is, however, not at all necessary. Our present communication system, augmented by some simple accessories derived from computer technology, makes it possible to make the popular will felt on issues of both national and

World Radio History

local importance. It is only proper that the engineer who realizes this fact and is convinced of the desirability of more direct popular control of government should work toward its realization.

The key to the problem lies in the parallel existence of two communications systems which establish contact with practically the entire citizenry. The first of these is the broadcasting system, comprising both radio and television. It permits the practically instantaneous addressing of questions and instruction to the entire population. The other is the telephone system, which enables the citizen to talk back, identifying himself uniquely by his telephone number. Occasionally, these two communication systems are used in combination. This occurs, for instance, when the master of ceremonies on a television programme asks a question and announces that a prize will be awarded for the first correct answer telephoned to the studio.

This particular application, however, is not only trivial, but clearly limited. If a large fraction of the television audience would try to reply then and there, the telephone switchboards would be hopelessly jammed. We shall see however, that this difficulty can be overcome by a relatively simple addition to the telephone installation.

Before going into details it is only proper, however, to establish that the telephone system, like the broadcasting system, is indeed universal in its coverage. My figures will apply again to the United States, since I am not so familiar with the statistics applying to the United Kingdom. There are, at present, approximately 60 million telephones in operation in the United States. Since its total population is about 175 million and since about four people can be counted per family, it is evident that the great majority of American families are in possession of telephone service. Citizens of voting age not so covered could be assigned to public telephones in their area of residence.

The sequence of the voting operation would be as follows. To begin with, the questions on which a vote was required would be presented over the regular television and radio broadcast channels, with a time limit set for the voting period. The same information would, of course, be disseminated by public media, including newspapers and posters. The individual voter would then record his vote on his "tele-voter," a box provided with a series of yes-no push buttons and attached to his telephone. To unlock the mechanism of his "tele-voter" he would have to insert his individual voting card, coded with an appropriate key. In general, a tele-voter would be provided with several blocks of push buttons, each assigned to a particular voter in the family and unlocked by that voter's card.

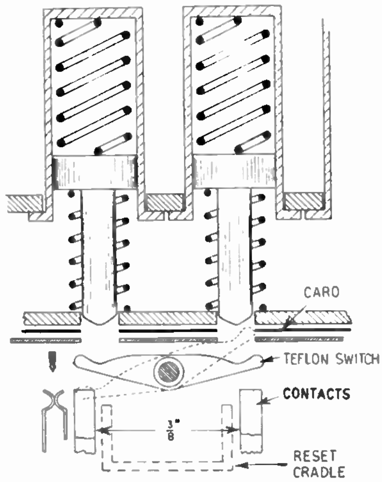


Fig. 13. Yes-no switch in the "tele-voter."

The voting information would be retained in the tele-voter in the form of a series of closed and open switch positions. A possible arrangement of a pair of yes-no push buttons is shown in Fig. 13. Here all the switches are closed in the neutral position, preceding a vote. This protects the electrical contact surfaces and facilitates electrical checks of the equipment. Depression of the doubly-sprung push-button opens the switch by inserting a teflon plunger between the contact surfaces. The switches may be returned to their neutral positions by the actuation of a cradle.

At the end of the specified voting period, interrogating pulses are transmitted from the telephone exchange to all the telephones in its district in sequence. These trigger the transmission of a series of pulses corresponding to the

World Radio History

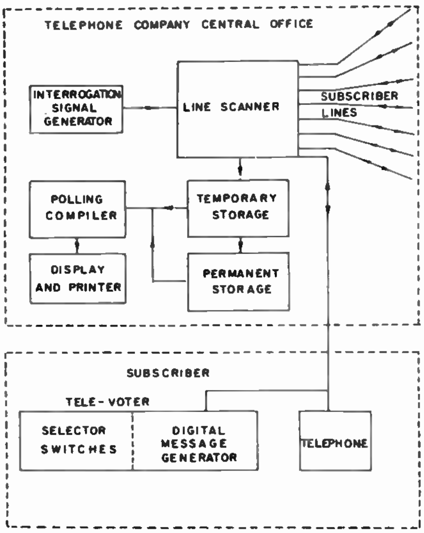


Fig. 14. Block diagram of vote recording system.

switch positions in the tele-voter, which are summed and tabulated at the exchange. The results of the counts of the several exchanges are transmitted in similar fashion, over broadband communication channels, to a central station, which thus obtains a record of the national vote (Fig. 14).

Suppose that the individual tele-voters transmit signal pulses at the rate of 1,500 per second and that there are 30 push buttons on every tele-voter. Then the interrogation time of one telephone will be 1/50 second. Since a telephone exchange serves at most 10,000 telephones, the polling of its entire district would have to take only about four minutes. If the polling of the exchanges by the central station takes a similar time, the national results of the vote will still become available within a small fraction of an hour after the expiration of the voting period. It should be noted that the interrogation would not interfere in any way with normal telephone traffic; the 1/50 of a second interruption would at most be noted as a faint click. Incidentally, the interrogation pulse would also actuate the reset cradle which would return the switches to their neutral position after the transmission of the voting record.

The principles of the tele-voter operation is best illustrated by a mechanical model such as that in Fig. 15. Here the spring-driven arm of a commutator contacts in turn the several selector switch leads and causes the transmitted oscillation to be increased or decreased in frequency, depending on whether the selector switch in question is open or closed; the arm is released in response to the interrogating pulse transmitted by the central office and is reset manually before the next vote is recorded. The transmitted sequence of pulses of frequency modulation thus reflects the selector switch settings on the tele-voter.

With the electromechanical system, approximately a second would be required to transmit the pulse pattern for a single telephone station, so that approximately three hours would be required to scan all the lines in a telephone exchange. It has other more serious drawbacks, such as contact deterioration. Much faster and more satisfactory operation can be realized with the aid of a magnetic-core or thyristor shift register (Fig. 16). Here the interrogating pulse from the central station is shifted by shift pulses from a 1,500 c/s electronic "clock" from one selector switch position to the next; the clock is started by the interrogating pulse and is stopped by the same pulse as it returns from the shift register. The pulses transmitted by the selector switches are applied to a frequency shift key oscillator (f.s.k. carrier oscillator) and produce here a sequence of frequency-modulated pulses on a carrier wave. The basic simplicity of the thyristor shift register is illustrated by the four-stage unit shown in Fig. 17. The selector switch leads would here be connected to the collectors of the thyristors. The compactness of the equip-

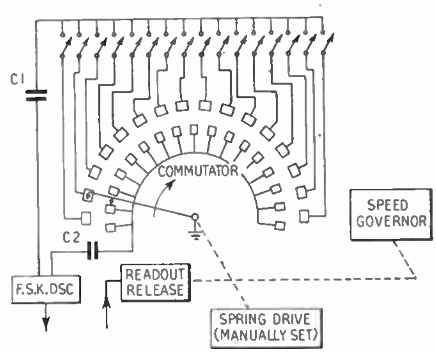


Fig. 15. Mechanical model of the tele-voter.

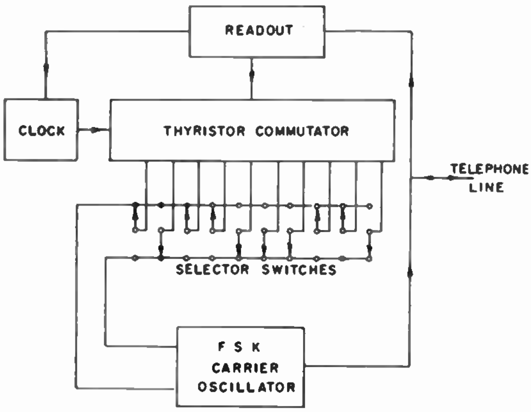


Fig. 16. Thyristor model of the tele-voter.

ment is illustrated by the fact that one stage of the thyristor shift register can be cast in a block of plastic material 0·1 cubic centimetre in size.

An attractive feature of the electronic polling system here described is its flexibility. Thus, to obtain votes on purely local issues, such as the building of roads, the construction of schools, and changes in the tax rate, the interrogation is limited to the telephone lines in the voting districts in question. Furthermore, simple modifications permit its use for such objectives as stockholder voting in corporations and the automatic transmission of meter records of public utilities for billing purposes. These could be of considerable economic value. However, the primary conclusion which I wish to stress at this point is that the broadcasting system and telephone system in combination provide us with an opportunity of giving entire peoples a direct control over their destiny which heretofore has only been enjoyed in relatively small, compact communities.

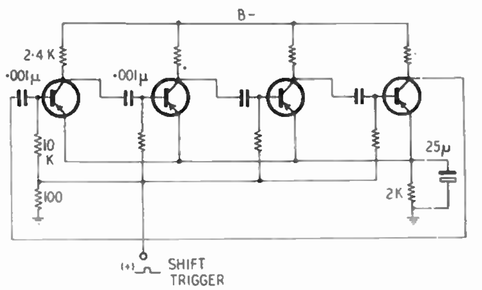


Fig. 17. Thyristor shift register.

Conclusion

In summary, I have indicated by three examples, how an electronics engineer, by examining areas which might be judged remote from his usual occupation, can find fertile fields for the application of his skills and thus make a greater contribution to the welfare of his fellow man. Clerk Maxwell amply recognized this need for breadth of outlook among scientists, advising them in his last public lecture, "we cannot do better than improve the shining hour in helping forward the cross-fertilization of the Sciences." Medicine, transportation, and the mechanics of democratic government, all offer the electronics engineer worthwhile opportunities to follow this advice.

While I have chosen these three fields of engineering endeavour for my examples, it will now be clear that much vaster opportunities exist for engineering and for the engineer, not merely as an artisan in physical science, but as an architect of the progress of humanity.

World Radio History

Communications in Independent Television†

by

L. F. MATHEWS, MEMBER‡

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

Summary: The effect of expanding demand on the design of a network for the conveyance of vision, sound and the control of programmes within a television programme organization is described. Reference is made to problems associated with exchange of programmes with similar organizations. Outside broadcast radio links and methods adopted whereby a number of contractors can participate in a composite link are discussed. A new microwave link for monitoring purposes is described, with particular reference to the automatic switching and monitoring facilities carried out over an associated u.h.f. link. Uses of the communication systems other than for scheduled programme transmission are described, and the paper concludes with a short account of future plans and developments.

1. Introduction

Although the Independent Television Authority had planned the areas to be served by the television transmitters prior to the commencement of independent television in September 1955, the only sites finally selected, and on which transmitters were being erected, were Croydon for the London Station, Lichfield for the Midlands, and Winter Hill for the North West of England. Links to connect these transmitters to main Post Office centres in London and Birmingham respectively were in the process of being installed, and an existing 900 Mc/s radio link operating between London and Birmingham was being refurbished in readiness for operation between these centres. The two London contractors were providing themselves with a skeleton network in order that studios and main technical areas could be connected to each other and to the London Post Office switching centre.

At this stage it was impossible to determine what the ultimate communication requirements would be when a large number of contractors were operating because it had not been resolved whether networking of programmes between contractors would become the accepted practice and if so how an interchange of programme in this way might be effected. This uncertainty of the future complicated the preplanning of the domestic and national networks; the problem was increased by the high costs involved in such networks, and the time interval before service could be provided. The provision of a network of vision circuits utilizing radio links, coaxial, balanced pair and telephone pair cables was, however, embarked upon, and a national communications system has been evolved which, although comparatively complex, does, in fact, enable British Independent Television communications to function in a satisfactory manner.

2. Independent Television Expansion

On February 17th, 1956 the Midland contractors commenced operation and programmes were transmitted from the Lichfield transmitter, the programme companies' transmission network at this time being as shown in Fig. 1. This network was mainly constructed using 0·375-in. coaxial tubes, 1-in. coaxial tubes, and twin coaxial cables operating in a balanced manner.

In addition to the video circuits described, music and control circuits were required from all the centres and these were provided either over the interstice pairs which surrounded the coaxial tubes in various cables or via a separate cable provided solely for this purpose.

† Manuscript received 13th March, 1959. (Paper No. 517.)

‡ Associated Television Ltd., 17 Great Cumberland Place, London, W.1.

U.D.C. No. 621.397.74

World Radio History

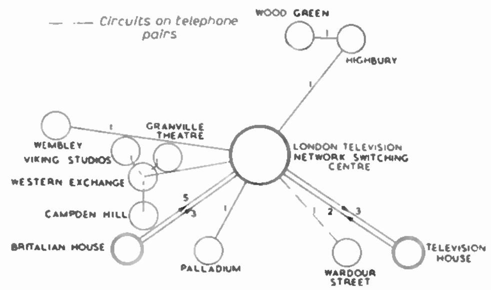


Fig. 1. Post Office network for programme companies in London. February 1956.

At this time the main network was as shown in Fig. 2. This, as has previously been explained, provided a bothway connection between London and Birmingham over a 900 Mc/s radio link, music and control circuit networks covering the same area having also been brought into operation.

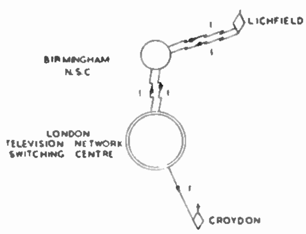


Fig. 2. Post Office network for Independent Television Authority. February 1956.

With the increase in the number of contractors and the amount of networking resulting, the situation both with regard to domestic communications and I.T.A. networks made further demands and at the present stage the independent contractors' London network is as shown in Fig. 3.

As a result of collaboration between the author's company and the Post Office, a new and satisfactory method of operating unbalanced coaxial cables at video frequencies over substantial distances has been evolved. This has proved to be of extreme value in many cases by eliminating the necessity for carrier equipment. Apart from its cost, this equipment was also in short supply.

It will be noted that a number of telephone pair circuits have been provided for the transmission of video signals to various offices and viewing centres in order to provide monitoring or other closed circuit facilities.

Figure 4 shows the present day condition of the network and, comparing this with Fig. 1, it will be seen that a great expansion has taken place in the matter of three years; the link has been extended to serve Northern England, South Scotland and the South and West of England. All links are now operating as two-way systems except for the Newcastle link on which the return circuit is still in course of construction. In addition, the main backbone of the network between London and Manchester has been increased by the addition of another link in either direction.

3. Networking

The planning of a network of both company and main links has been complicated by the fact that any one contractor may be providing a programme to all or any number of other contractors, demanding a complete re-arrangement of the network connections at a precise time. In order to provide for these conditions, and in the absence of instantaneous switching, it would have been necessary to adopt a very

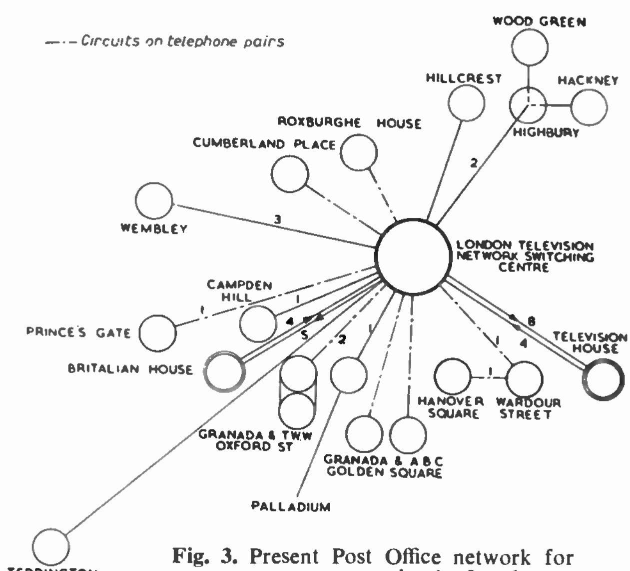


Fig. 3. Present Post Office network for programme companies in London.

World Radio History

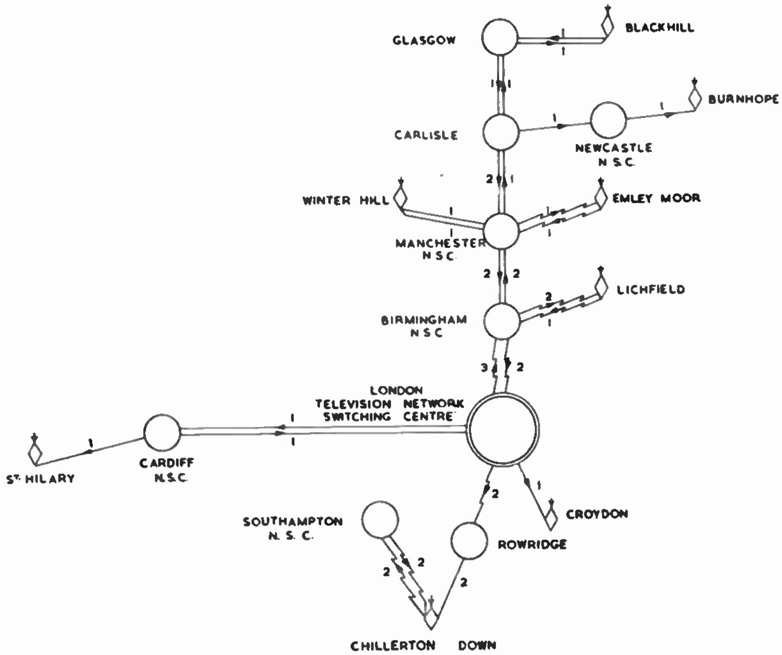


Fig. 4. Present Post Office network for Independent Television Authority.

large and complex network to cope with all the combinations likely to occur. However, the Post Office were able to undertake this switching and equipment was installed at the Post Office main centres whereby this operation could be carried out. Figure 6 shows the switching equipment recently installed at Museum Exchange for this purpose. Three separate programme chains can be set up in advance of connection time, the actual switching being carried out by means of a time clock, or by a manual operation. Each chain can select any source and the output be routed to any one of the destination channels. The equipment for setting up routes is shown in Fig. 5. By carrying out similar switching at Birmingham and Manchester the network can be selected to operate in a large number of different combinations, both in vision and sound.

4. Outside Broadcast Links

Outside broadcasts play an important part in the complete programme schedule and most of the contractors have a number of mobile camera units and outside broadcast radio link equipments. Radio frequencies for these purposes have been allocated by the Post Office as follows:

Frequencies for vision :

Table 1 shows five pairs of frequencies in the microwave spectrum for use between the ten present contractors. Some frequencies within the band 610 and 622·5 Mc/s can be made available for use, but at the moment very little equipment is working in this part of the spectrum.

Frequencies for music circuit link equipment : 76·825 Mc/s ± 0·01%, 50 kc/s channelling, and output power of 25 watts is allocated to be shared between all contractors except the East Anglian contractor, the latter having been allocated a separate frequency owing to the normal frequency being used in isolated areas in this region for the provision of Post Office

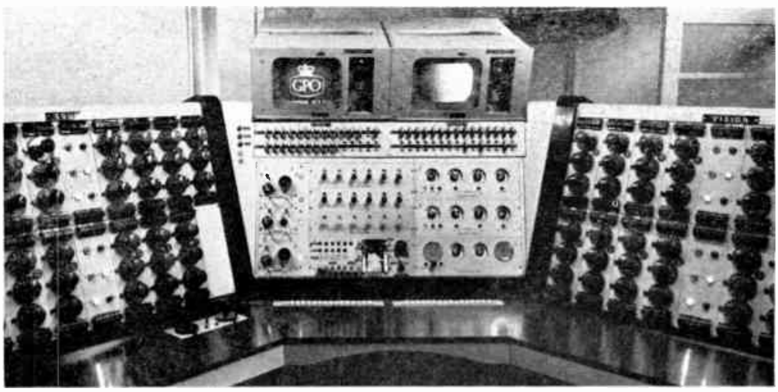


Fig. 5. Equipment used for setting up routes.

telephone kiosk service. Frequencies within 660 and 661·25 Mc/s are available to all contractors if required.

Frequencies for producer talk-back equipments: At the moment a frequency of 86·825 Mc/s has been allocated between all the existing contractors.

Frequencies for engineering radio telephone purposes: 160·9 Mc/s and 161·45 Mc/s have been allocated to provide two channels, the programme contractors being split into two groups and allocated either one of these frequencies.

As these frequency allocations are confined to a 50 kc/s channelling some present day equipment does not conform to these standards and the Post Office have agreed to contractors retaining a previous frequency allocation on either a 165·1 or 169·9 Mc/s on a channelling of 100 kc/s until January 1962, after which time the new frequency schedule must apply.

At the time of writing, frequency allocations are also being considered by the Post Office for use on "peepie-creepie" mobile cameras for transmission of vision and sound from cameraman to base receiver.

Due to the restricted number of frequencies allocated, it will be seen that frequency sharing has been unavoidable. This has made it necessary for contractors on occasions to plan their programmes in conjunction with each other in

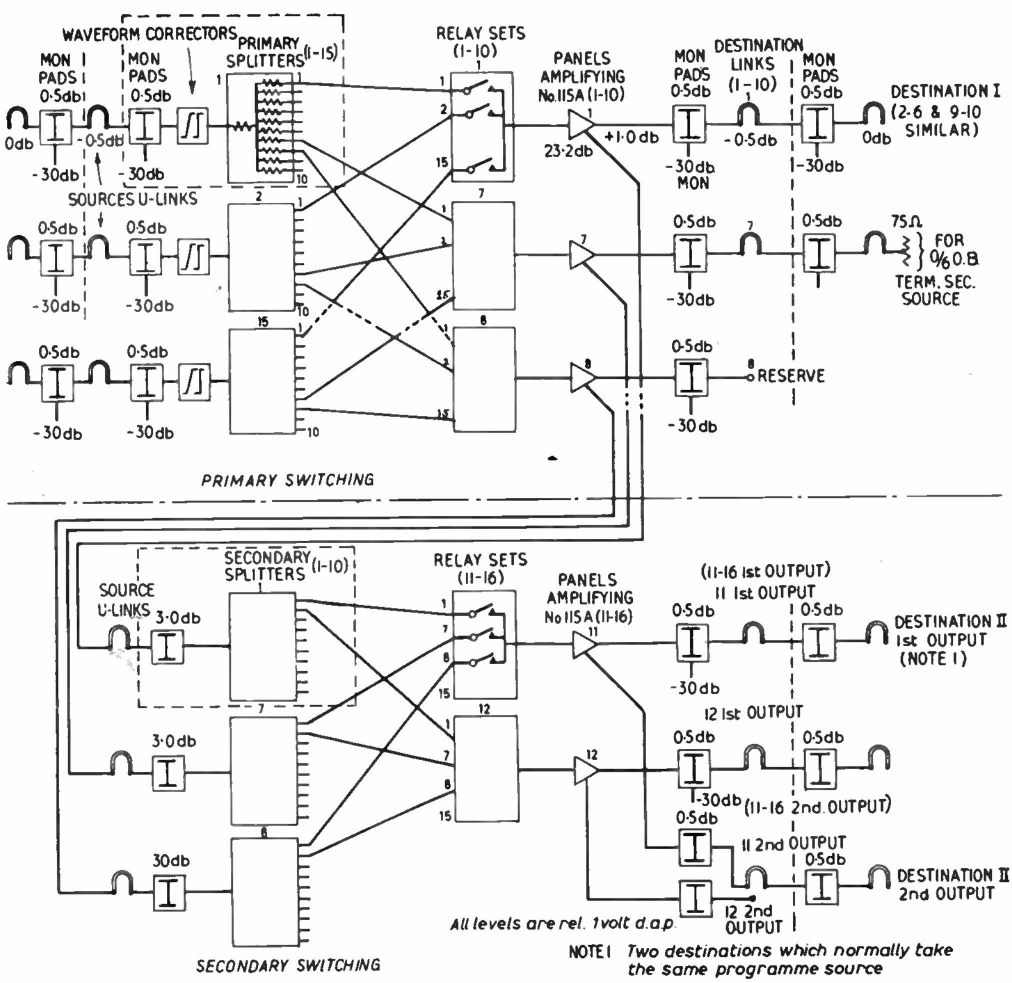


Fig. 6. Schematic diagram of automatic network distribution equipment for vision switching.

World Radio History

order to avoid interference between their respective operations. It has, at the same time, however, been possible to carry out composite programmes combining different contractors' equipments.

In order to augment the microwave communications a method has been developed by the author's company of running temporary cable circuits for distances up to two miles to carry signals from the site of an outside broadcast to the nearest high point where a microwave link can be effectively used. These circuits are normally provided over an inexpensive four-pair cable suspended over or buried in the ground, as may be necessary. This cable can be recovered and used over and over again and has a fairly long life.

Microwave equipment which is capable of being tuned on site to the various frequencies allocated has found favour with the contractors as it is then possible to re-tune the equipment at a particular site in order to operate with other contractors' equipments.

Table 1

Frequencies allocated for Programme Contractors' Outside Broadcast Vision Links.

7207·5 and 7267·5 Mc/s
7231·5 and 7291·5 Mc/s
7183·5 and 7243·5 Mc/s
7219·5 and 7279·5 Mc/s
7195·5 and 7255·5 Mc/s

5. Programme Monitoring Facilities

Due to the author's company operating both in London and Birmingham it has been found necessary to introduce a private link for vision and sound between these places purely for monitoring purposes. By this means programmes transmitted from Lichfield and which are not available in London except when networked, together with studio rehearsals, dry runs, etc., can be viewed in London at any time without occupying normal network circuits. This link which operates using frequencies 7304 and 7404 Mc/s is routed as shown in Table 2.

The system utilizes standard 4, 6 or 10 ft. paraboloids except at the Barkway repeater, where two 12 by 9 ft. passive reflectors are installed on the mast. The link also passes the associated sound programme by modulating a 6 Mc/s subcarrier prior to mixing with the vision signal. The microwave link is duplicated except for the aerial system which can be switched to the normal or standby equipment by "waveguide switches". The change-over is automatic in operation, and is achieved by transmitting over the microwave system a 7 Mc/s pilot signal which will instigate a change over to standby equipment when it encounters a fault condition.

Table 2

London-Birmingham Monitoring Link

Birmingham	to	Meriden	(Warwickshire)	13½ miles
Meriden	to	Cold Ashby	(Northants)	23½ miles
Cold Ashby	to	Barkway	(Hertfordshire)	51½ miles
Barkway	to	Highgate	(London)	31 miles

All transmitter output powers are 1 watt.

In addition to the microwave equipment each station along the link is equipped with a 460 Mc/s transmitter and receiver which is used to provide an engineering speech channel and an alarm system which indicates simultaneously the state of the link on annunciator panels at the Birmingham and London Studios and also at a maintenance depôt located at Cambridge. Briefly the whole system operates in the following manner:

At regular predetermined intervals a signal consisting of 1·8 kc/s and 3·1 kc/s tones is sent from Birmingham to all stations. In response to this signal all station alarm systems return a signal to Birmingham and also to London and Cambridge. Each remote station alarm system has a built-in delay which allows sequential reporting without overlap.

At each station four tones are generated—400 c/s, 850 c/s, 1·8 kc/s, 3·1 kc/s and each transmitter and receiver at each point is allocated two of these audio tones. When a remote station reports back, first a tone code is sent giving station identity, followed by two out of the four tones in sequence, depending upon units in operation. In the event of mains failure all four tones are returned. This sequence of events takes place from all stations in turn.

At Birmingham, London Studios and Cambridge the display panel presents on a mimic diagram information of the complete link; cold cathode tubes are used to prepare

World Radio History

the display. At the end of the scanning sequence the system is cleared in readiness for the next automatic check of the link.

The triggering automatic sequence can be overridden by mechanical operation at any time should an instantaneous test of the link be required. Figure 7 shows the layout of the annunciator panel in use at the monitoring stations.

6. Subsidiary Uses

Although the communication facilities are in general primarily intended for use with live or recorded television programmes, other types of operation can also be undertaken. For example, one such operation consisted of a closed circuit programme carried out from a manufacturer's premises in Cambridgeshire which was linked with a number of hotels throughout the country as shown in Fig. 8. Two microwave links were used to pass the programme vision into London, whence it was sent over the permanent cable to Museum Exchange, where a telephone pair vision circuit passed the programme to an hotel in Park Lane. The programme was also distributed from London to the main north-going network and to the circuit to Bristol. At Bristol the signal was passed via telephone pair circuit to monitors in an hotel; the north-going feed from London was passed to Lichfield over existing links from whence the vision was linked by a two-hop microwave path to Lincoln where it again terminated on monitor equipment in an hotel. The north-bound link from Birmingham extended the vision programme to Manchester, where by use of a further telephone pair circuit the programme was sent to a local hotel. The signal going northwards to Glasgow on the main link was passed to Black Hill from where two microwave link hops passed the vision

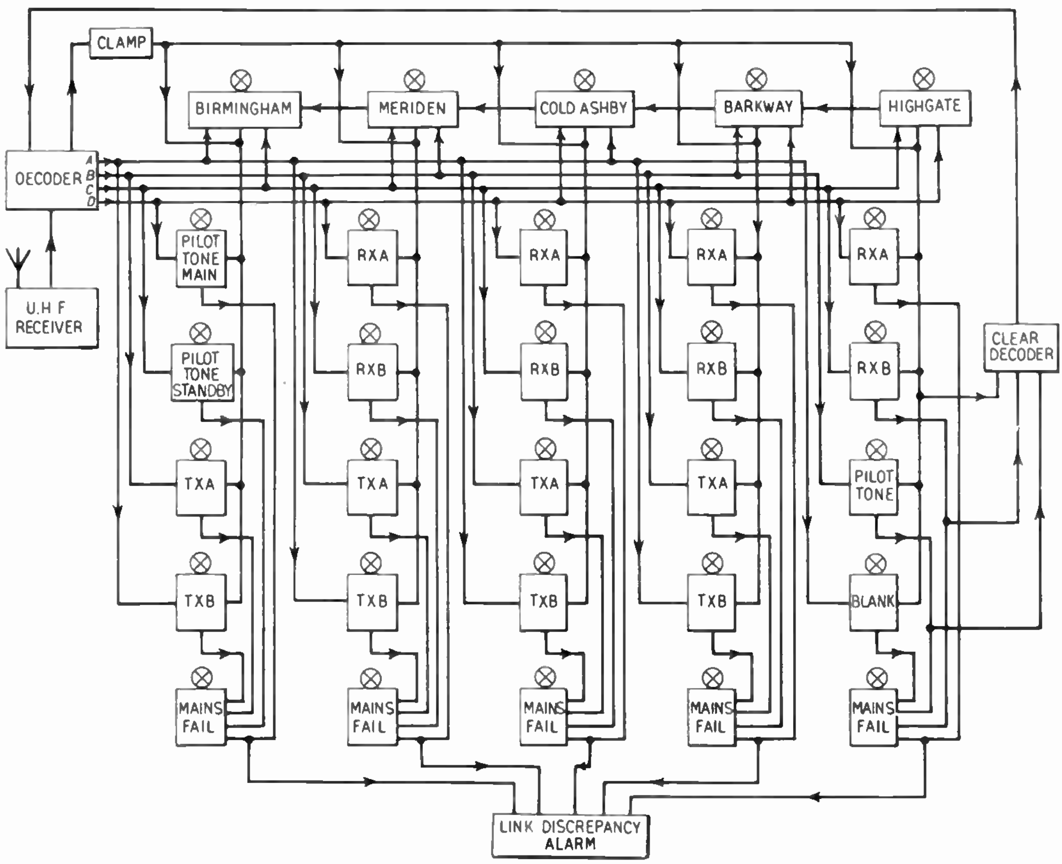


Fig. 7. Layout of annunciator panel at monitoring stations

World Radio History

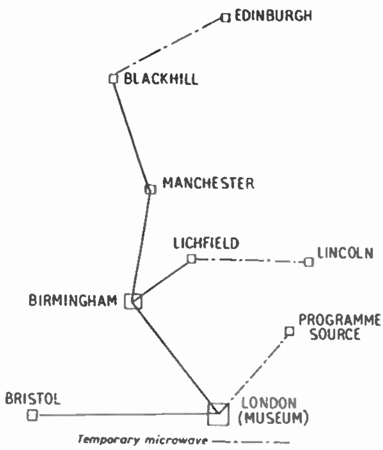


Fig. 8. Network used for closed circuit project.

programme to an Edinburgh hotel. The accompanying sound programme was fed from its source to London where, by utilizing Post Office music circuits, it was distributed to all the required destinations.

This operation was a fairly extensive one and, although the programme was of only half an hour's duration, it involved the use of approximately 500 miles of permanent network, either on cable or over radio link, and also required the use of seven outside broadcast type microwave links in addition to a number of vision link circuits on telephone pair and a fairly elaborate sound circuit network. In addition to the foregoing there was a two-way control circuit facility to all points from the source so that the audience in the hotels were able to pass questions back to the point of origin.

7. Future Developments

It has previously been shown that the expansion of the main network had been undertaken during a time when a number of problems in respect to the ultimate requirements and the future of networking were unresolved. Expansions are still taking place, in both the network and the number of contractors, which will alter the situation in the future. Figure 9 shows the condition of the main network as planned at this stage.

The problems connected with programme sharing, as previously described, will increase with the number of contractors, and a method of combating this situation might well be the adoption of a sub-carrier system whereby in addition to one video programme modulating a microwave transmitter carrier, a second programme is first modulated on to a sub-carrier and then modulated on to the same microwave carrier.

The advent of colour television and its effect on present-day links and future links must be considered, also the possibility of remotely controlling and switching all permanent networks on the lines of the telemetering circuits described previously.

In the future design of main network links, an important factor may be the frequency spectrum requirements needed to take account of higher definition and colour, and this is likely to produce development on both cable and radio link networks. The tendency with the latter is to go to even higher frequencies and 10,000 to 12,000 Mc/s are under active consideration. These have very definite advantages if the

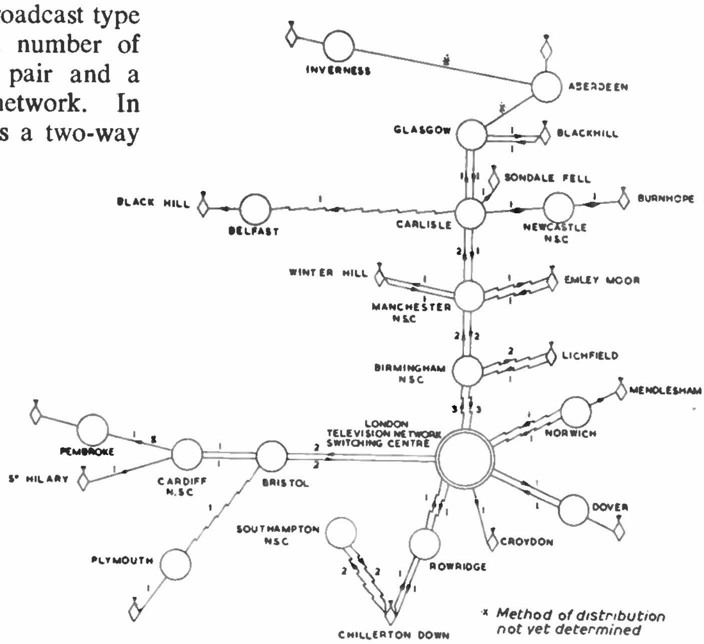


Fig. 9. Future development of network for Independent Television Authority.

requirements associated with propagation can be completely satisfied.

The advent of the travelling-wave amplifier is already revolutionizing the design of long distance links, and intermediate repeater stations, operating purely at microwave frequencies, using three travelling-wave amplifiers as (1) radio frequency amplifier, (2) frequency changers, and (3) local oscillator, are being investigated. These would have certain economic advantages over the present systems and at the same time would aid engineers in achieving higher technical standards.

No mention has been made in this paper so far of the Eurovision network, although Independent Television does participate in this operation. If some of the existing problems can be overcome, it should be possible to have an interchange of programmes on a much greater scale. The technical problems are complicated by the differences in scanning standards in use in different countries, and should these not become compatible in the future, the answer to this particular problem will, it is anticipated, be in the purely electronic type of standards converter which, not being dependent upon optics, is likely to introduce less degradation into the system than does the present type of standards converter.

Finally, much interest has recently been shown in transatlantic communication for television by using the forward-scatter principle. This particular method of television communication is a very exciting possibility, but not very practicable at present due to the problems associated with the unpredictable propagation conditions, the wide bandwidth, and the very large transmitter powers necessary.

8. Acknowledgments

The author wishes to thank the Managing Director and the Technical Controller of Associated Television Limited for permission to publish the paper. He would like also to acknowledge the help given by the Post Office in the preparation of this paper and to thank the Independent Television Authority for their courtesy in allowing reproduction of illustrations of their network.

World Radio History

THE 1959 BRITISH RADIO SHOW

THE 26th National Radio and Television Exhibition was held in London between 26th August and 5th September. Its emphasis on domestic radio and television, and audio equipment, was referred to by Mr. E. E. Rosen, Chairman of the Radio Industry Council, at the opening ceremony.

Referring to that part of the British radio industry which might be described as "home entertainment," Mr. Rosen stated that set manufacturers employed some 100,000 people; of this segment of the industry alone some £3½M per annum is spent on research and development, of which £2½M is expended on salaries for scientific and technical personnel. This commentary on the technical background to the Show is reflected also in the exhibits. In recent years there has been a tendency to reveal more of the technical background rather than concentrate on what has sometimes been referred to as "furniture display." The public advertising at exhibitions of instruments and technical equipment is increasingly showing awareness of the average man's interest in technical matters, and from the engineer's point of view, this tendency does of course greatly add to the interest of visiting what is primarily a public exhibition.

Television Receivers :—Typical of this type of approach was the presentation of the most striking technical advance in television receivers to be seen at this year's exhibition, namely, the general introduction of the 110° scanning angle picture tube. This has come into use entirely since last year's show. It seems unlikely that, for some time at any rate, there will be any tubes employing wider angle scanning than 110°, as the attendant complications of higher power and more difficult glass envelope design would outweigh the advantages of still further reducing the tube length. What does seem probable is that there will be changes in the electron gun construction to enable it to be sited much more closely to the deflection coils, thus shortening the neck of the tube. With 110° scanning the reduction in depth of a 17 in. tube is, in any case, about 4 in. compared with 90° scanning, and this shorter neck tube, examples of which were on view, will give perhaps another 1½ in. to 2 in. reduction. The deflection coils themselves now have to be taken further up the flare of the tube, and special arrangements have to be made for mounting the associated magnets used for picture shift. One interesting development in this connection is the use of flexible magnets made by moulding ferrite material in a plastics medium: the magnets have to be so thin that the rigid type would tend to warp, and it is also claimed that there is an appreciable saving in scanning power.

In the case of the transportable receiver, which has achieved reduced weight due to there being less glass in the tube and less material in the cabinet, a more easily carried shape has been obtained—one receiver using a 17 in. diameter tube is only 12¾ in. in depth and weighs only 31½ lb. For receivers not claimed to be transportable, the reduction in depth has made possible a set which *can* be hung in a wall recess. The position of the loudspeaker has become rather a problem with these sets, and twin speakers are in some instances mounted vertically on each side of the screen, which is, of course, acoustically more desirable than the side or top positions which have been used in recent years.

Reference was made in our review of last year's exhibition to the necessity for grouping the sub-chassis round the tube neck, and it is typical of the more realistic attitude of set manufacturers to the inevitable need for servicing that these units are easily accessible, and often are designed for plug-in replacement. Printed or plated wiring boards for sub-assemblies contribute both to the reduced size and weight and to the serviceability of the new sets.

The inclusion of v.h.f. radio channels in television receivers has been a popular feature, and now nearly all manufacturers include a set of this type as an alternative for all their models.

The remote control of television receivers has, of course, been talked about for some time, and one or two examples have been shown in the past. This year, however, two manufacturers showed control boxes which are connected to a receiver by a multiway cable, and enable channel changing to take place using a motor driven switch, as well as adjustment of volume and brightness. A third manufacturer has announced the impending production of a remote control device without connecting wires, using ultrasonic radiation to switch the set on and off, change channels and adjust volume within a range of 25 feet.

Other technical improvements which are becoming general include the wider use of the more sensitive "frame-grid" type of double triode r.f. valve. With the object of reducing the number of controls ultimately to only the programme selector switch, great attention has been paid to temperature compensation during the warming-up period. More stable time base circuits and improved automatic gain control all contribute to this aim of ease of operation. A future development, which may prove popular in this country as it has elsewhere, is the introduction of an automatic contrast control circuit demonstrated by a valve manufacturer. Here a small light-sensitive resistor is incorporated either in the a.g.c. circuit or between the video amplifier and cathode ray tube, to cause the brightness and contrast of the picture to follow automatically changes in the ambient light. One advantage of incorporating the resistor in the a.g.c. circuit is that it controls only d.c., hence its location and lead lengths are not critical. However, the other arrangement, as an attenuating circuit before the tube, does mean that the audio signal is not affected by changing light conditions.

Broadcast Receivers :—Broadcasting on the v.h.f. band is now firmly established, and in addition to the incorporation of these channels in television sets, all manufacturers showed table receivers for Band II as well as long and medium waves. The most noticeable feature, as far as receivers is concerned, was the great number of portable transistor sets. Developments of this type of receiver include an extra transistor to permit of greater sensitivity and improved sound quality through the use of negative feed back. A few models have a short wave band in addition to the usual long and medium wave bands. A transistor manufacturer demonstrated a portable receiver which also permitted reception of v.h.f. f.m. transmissions. It employed an r.f. amplifying stage preceding a self oscillating mixer; an advantage claimed for the transistorized v.h.f. set was that difficulties encountered due to ambient temperature changes during warming up are practically eliminated.

Sound Reproduction Equipment:—Increasing scope for stereophonic reproduction from records has led virtually every manufacturer to include a radio-gramophone containing two complete channels from pickup to speaker. In some cases both speaker assemblies are mounted at either end of a long cabinet, sometimes on slides enabling greater spacing to be obtained, and in other cases one or both of the speaker units is an extension unit. Even portable record players can now be obtained in stereo versions, with the speakers in a detachable unit.

Tape recorders are also available for stereophonic reproduction either from tape records or from the user's own recording. Probably the most interesting innovation in tape recording is the introduction of a tape cassette, equivalent to a standard 4 in. reel, which, at the cost of increasing the size over the standard reel, does mean that spool changing is no more difficult than putting an ordinary record on a gramophone turntable.

Other Features :—The Careers Stand this year was concerned primarily with training for radio and television servicing, and the demonstrations were arranged with this in view; careers in the radio and electronics industry for engineers received comparatively little attention. The Royal Navy, Royal Air Force and the General Post Office, however, treated the recruitment of technical people of all grades very seriously on their respective Stands. Technical demonstrations of particular interest included a large analogue computer constructed by the R.A.F. Technical College, an indication of some of the features of the radar installation in H.M.S. *Victorious*, and explanations of subscriber trunk dialling and submarine telephone cable measurements.

While the B.B.C. and the Independent Television companies were, as always, mainly concerned with programme matters, the presence of a mobile control room and a mobile television recording vehicle gave some technical interest to this important aspect of the industry.

World Radio History

The Use of Television for the Microscopical Examination of Radioactive Metals †

by

E. C. SYKES ‡

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

Summary: Closed circuit television for relaying the image produced by remotely controlled microscopes enclosed within thick lead shields, and used for the examination of highly radioactive irradiated fissile materials, e.g. uranium 235, is described. Brief details are given of two installations incorporating respectively c.p.s. emitron, and image orthicon camera chains. A device is described which accurately measures the microstructural features displayed on the monitor screens. The advantages and disadvantages of television compared with optical viewing are discussed, and brief mention is made of experiments with television for general remote handling.

1. Introduction

The behaviour of fissile materials in a nuclear reactor environment, and the development of improved fuels constitute an important part of the research carried out by the Metallurgy Division at A.E.R.E., Harwell. In this work, samples of nuclear fuels are irradiated in a reactor, and then subjected to a series of tests, including a microscopical examination of a specially polished surface. The sample becomes highly radioactive during irradiation, and to protect people from the dangerous radiation, both the sample and the microscope have to be enclosed by thick walls of lead. The problem then is to bring the image produced by the microscope out through the shielding wall to a suitable observation point. This paper describes two installations at Harwell utilising closed circuit television for this purpose.

2. Concerning the Choice of Television

In the early 1950's when the active metallography project was started, the only suitable way of observing the grain structure of uranium was to examine the specimen under polarized light. The characteristics of this type of examination are the care needed in the optical system to avoid flare and depolarization, which easily destroy the contrast in the image, and the very low light levels of the final image, possibly as low as 0·002 ft. candles. In those days, long cranked optical paths suitable for high resolution work were almost unknown and time did not allow their development. Moreover, transfer optics would have required replacement at intervals since most ordinary optical glasses lose their transparency and are discoloured by the radiation from active samples.

The rival claims of ordinary television and the flying spot technique were then considered; television was favoured because of better availability, greater simplicity, and avoidance of the difficulties associated with the provision of a very high intensity scanning illumination, required for the flying-spot method. The high-pressure mercury-vapour lamps used for metallography have an intrinsic brightness of about 15,000 candles/cm^2 from a source about 1 mm^2, and making due allowance for light collection efficiencies, the flying spot illumination tube would need to develop at least 500 candle power.

Next, the type of television pick up tube had to be decided. Tests with a 300 curie cobalt source showed that high γ fluxes did not appreciably affect the quality of the picture produced by orthicon, image orthicon, or vidicon pick-up tubes. To obtain the best possible resolution and a large field, a full size camera tube was indicated. The image orthicon tube of those days had a rather low signal/noise ratio and was not readily available in a convenient camera. The photoconductive tube was attractive, but the image "lag" could not be

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‡ Formerly with the Metallurgy Division, Atomic Energy Research Establishment, Harwell; now with Central Electricity Generating Board, London, S.E.1.

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World Radio History

tolerated since, during polarized light examinations, the specimen is continuously oscillated through 90°. An orthicon was therefore selected for the first equipment. However, experience has shown that, whilst the sensitivity is adequate for examining non-irradiated uranium under polarised light, it is not sufficient to produce a good picture from irradiated uranium, the available contrast of the latter being considerably less than that of the normal metal. It is for this reason that a later installation employs an image orthicon camera tube. Nevertheless, the first television equipment has provided a large amount of valuable information.

3. Description of Installations

The television equipment is used as shown diagrammatically in Fig. 1; the microscope projects a real image on to the sensitive area of pick-up tube and the usual camera lenses are not required.

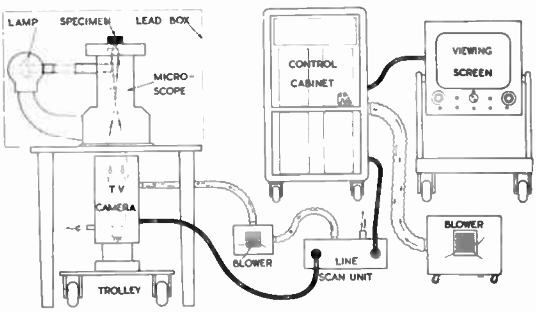

Fig. 1. Arrangement of microscope and television equipment.

The equipment shown in Fig. 2 was designed in 1953 for the examination of specimens of up to 10 curies of $\beta\gamma$ activity. Most of the interconnected shielded boxes are used when polishing the samples; the microscope is in the extreme right hand box. The television chain associated with this microscope is an EMI Type 10307 (essentially a modified outside broadcast unit) and works on a 625 line, 50 interlaced frames per second standard. The full size C.P.S. Emitron pick-up tube is housed, together with the head amplifier, in a trolley-mounted watertight casing below the lead floor of the microscope box. The line scan generator is a separate unit placed about 5 ft. from the camera

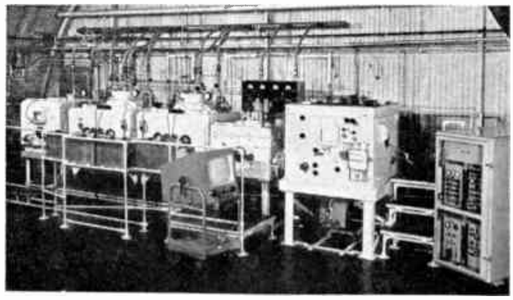

Fig. 2. Early equipment for the microscopical examination of radioactive metals.

and both are cooled by filtered air to ensure that the electronic components are kept free from air-borne contamination, a simple precaution which greatly simplifies maintenance, since rubber gloves and respirators may be dispensed with. For the same reason, the trolley mounted 14-in. monitor is sealed and surface cooled and the remaining four electronic units, all built as "suitcases", are enclosed in a large perspex-fronted cabinet through which filtered air is blown. The four suitcases contain the stabilised power supply, waveform generator, vision power and camera control units respectively.

The large cabinet also contains a special unit, designed by Mr. F. H. Wells of A.E.R.E. Electronics Division and shown in block diagram form in Fig. 3 (a). This unit produces two series of black-out pulses occurring either a uniform time after the start of each frame, or after the beginning of each line of the picture minitor; the former series are each sufficiently long (60 microsec) to black out a whole line and so produce a horizontal black line on the monitor screen, whilst the second series are of very short duration ($\frac{1}{4}$ microsec) and produce the effect of a vertical black line. The timing of these pulses, after the frame or line sync. pulses, and therefore their position on the monitor screen, may be varied by calibrated potentiometers mounted on the facia of the monitor. A gating circuit operated by the line synch pulse is incorporated to prevent the longer black out pulse being distributed over two lines.

Since the overall magnification of the microscope/television combination is known, the black lines may be used as cursors for measuring the size of objects appearing on the monitor screen, such as the microhardness indentation

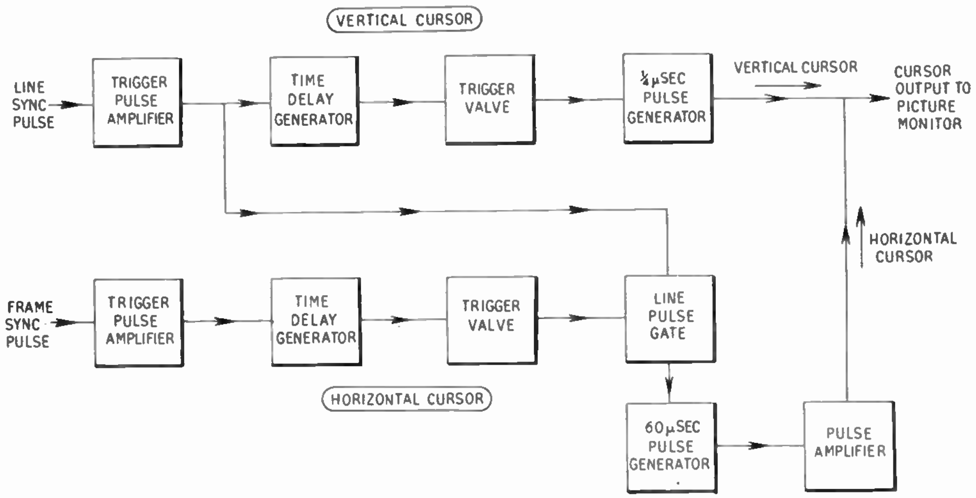

Fig. 3. (a) Block diagram of electronic cursors.

shown in Fig. 3 (b). Calibration indicates that the accuracy is better than $\pm \frac{1}{4}$ micron at an overall magnification of 3,000 times.

The microscope is mostly operated with lenses yielding a magnification between $\times$ 11 and $\times$ 350 which, with the $\times$ 10 enlargement introduced by the television results in overall magnifications in the range $\times$ 110 to $\times$ 3500 diameters. This range could be extended in either direction by use of other lenses, but in practice higher magnifications are of little benefit since the very short focus lenses necessary are rapidly discoloured and spoilt by the radiation from samples. The highest power lens in use, a 4 mm. dry achromat, has a maximum resolution of about 0·4 micron and in an image at $\times$ 3500 the smallest resolvable dot would spread over four lines on the monitor screen.

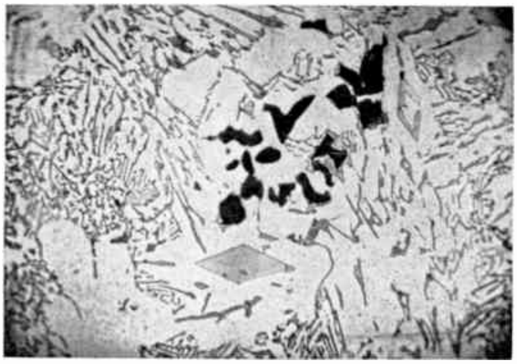

Fig. 4. Uranium-aluminium alloy (Magnification $\times$ 3000 on screen).

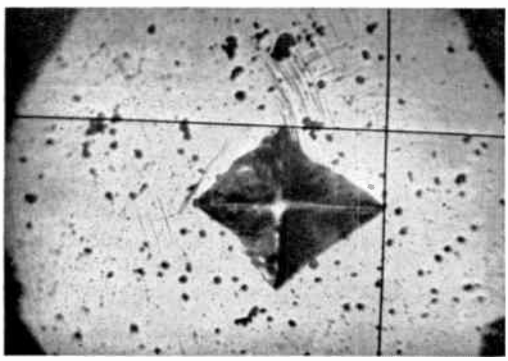

Fig. 3. (b) Cursors in use for measuring microhardness indentation (Magnification $\times$ 1500 on screen).

The images presented on the screen are recorded photographically using a specially built 5-in. $\times$ 4-in. plate camera incorporating a 90 mm. Schneider Angulon wide-angle lens. As might be expected when photographing a curved front cathode-ray tube, there is some image distortion towards the edge of the plate, but this is surprisingly small, and many photographs are later assembled into composite pictures. Figures 4 and 5 are typical photographs from this equipment.

A second metallography suite (Figs. 6 and 7) has recently been commissioned and is designed to accommodate specimens of up to 500 MeV-

World Radio History

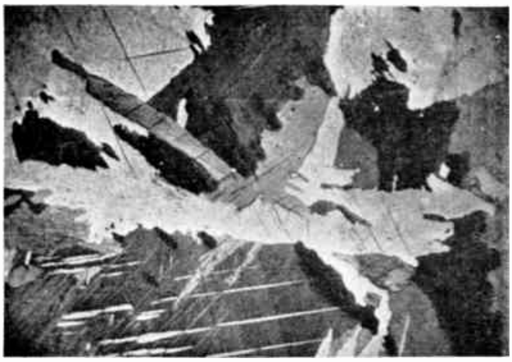

Fig. 5. Zirconium (Polarized light illumination. Magnification × 1250 on screen).

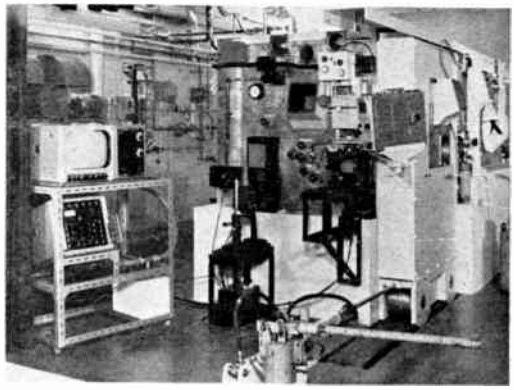

Fig. 6. Latest examination equipment—microscope control face.

curies of mixed radiations. For reasons already explained (Sect. 2) or to be discussed later (Sect. 4), this suite has been equipped with an image orthicon television camera rather than the orthicon type used previously, and with optical viewing.

In this apparatus, the microscope box and three of its 10-in. thick lead shielding walls are mounted on a concrete plinth of U plan form ; the shielding is completed by a 6 ton hydraulically-operated lead-filled door. The camera fits into the cut-out in the plinth and is confined by the lead door and the floor of the microscope box. It must therefore be appreciably less bulky than the normal broadcasting type camera and the camera fitted is, in fact, a slightly modified version of the Pye underwater unit. As with the earlier installation, all the electronic components are cooled with filtered air. The operator's work has been made somewhat easier by providing a 5-in. monitor for his personal use and by feeding the television circuits from a stabilized supply capable of dealing with greater mains variations than the built in regulator. An electronic cursor unit and a plate camera are also fitted. This image orthicon system goes a long way towards giving the desired quality of picture when using polarized light, but the intensity of light incident on the pick up tube is so low that noise is sometimes obtrusive.

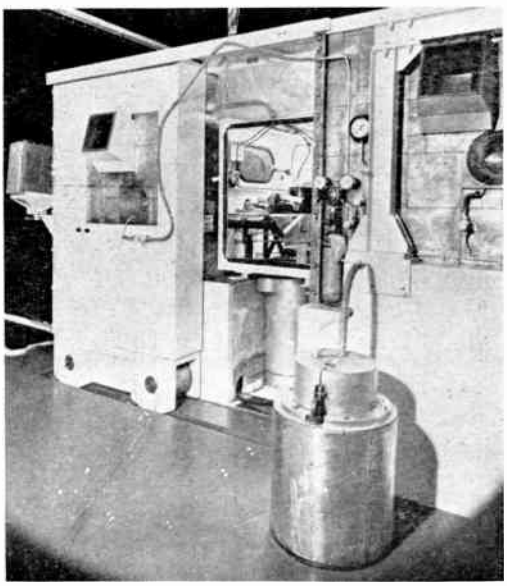

Fig. 7. Latest examination equipment—showing location of television camera.

4. Comparison of Television and Optical Systems

In comparison with the optical image transfer systems now available, the chief disadvantages of television are the poorer response with the low light levels of polarized light examinations, and the loss of colour effects. The principal advantages are the ability to present a large bright image suitable for simultaneous observation and discussion by numbers of people, and to relay this image over a greater distance than is possible optically.

The complex microstructures sometimes present in irradiated fissile materials make the possibility of group discussion a much more attractive feature than might be supposed.

World Radio History

Conversely, the loss of colour is less serious since it is usually possible to recognize features by their mono-chromatic tonal value.

Optically-relayed images almost invariably seem sharper than the equivalent televised picture because of the line structure and slight image spread in the latter, but such effects do not interfere appreciably with the metallographic value of the picture. The real image presented by television facilitates rapid and accurate estimates of the size of metallographic features, and if the electronic cursors are used, the measurements are free from the parallax errors present with most optical measuring systems, are much less tiring, and have a smaller operator error than if an optical micrometer is used.

One may thus conclude that the use of television with remotely controlled systems is practical and convenient, and is a technique which can, for many purposes, show definite advantages over purely optical image transfer systems.

5. The Use of Television for General Remote Handling

The general handling of radio-active substances is usually performed in large concrete cells with the aid of master-slave manipulators. Present day practice is to use large thick windows of glass and zinc bromide for viewing purposes, but the author would like to mention some experiments made several years ago using small industrial Staticon type television cameras. The experimental manipulations normally involved dropping 0 B.A. bolts into clearance holes or building towers of rubber bungs on a given spot. Regardless of the camera position, the sense of distance between objects seemed to be lost, even when using two cameras and monitors working as a beam splitter/polarized light stereo unit. In general, it was found that the shadows caused by strong oblique lighting in two directions were much more helpful than the stereo effects and tasks could be completed somewhat more quickly, more deftly, and with less fatigue if the shadows were used for positioning, although it must be admitted that the participants were all accustomed to working by direct visual contact, and might therefore be unconsciously biased. When the public were invited to try similar manipulations on an Open Day, using non-stereo television, most people found the tests reasonably easy when given the correct lighting conditions and advised to concentrate their attention on the shadows. From these experiments it is inferred that when television is used for remote viewing of handling operations the complexities and expense of stereo systems may be unnecessary if proper lighting is installed.

6. Acknowledgments

The work briefly described in this paper is part of a much larger project and many people have contributed in one way or another; their help is gratefully acknowledged. In particular special thanks are due to Mr. V. J. Haddrell who has been very closely concerned with this application of television.

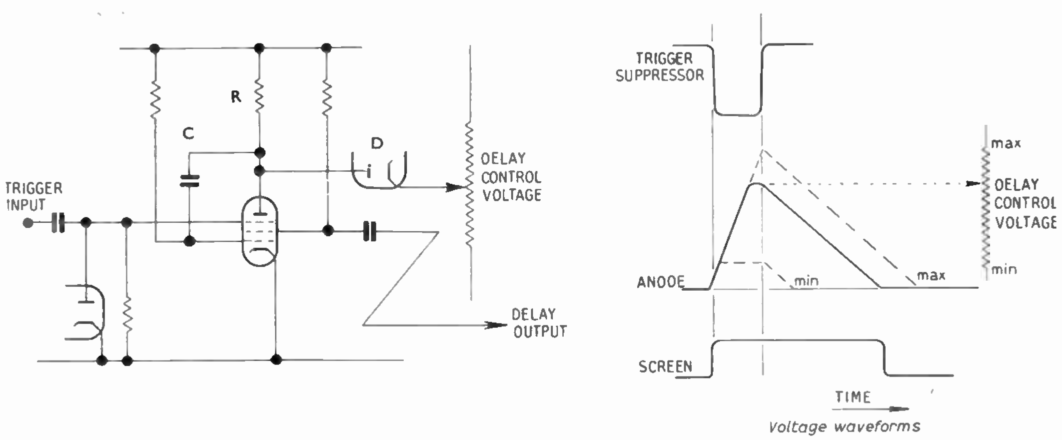

Fig. 8. Time-delay generator circuits and waveforms.

World Radio History

7. Appendix
Electronic Cursors

Whilst it is not the purpose of this paper to delve into details of the television circuits, it is felt that some portions of the electronic cursors (Sect. 3, Fig. 3 (a)) are sufficiently unusual to justify brief description, particularly the short pulse generator used to produce the vertical cursor line and the time delay generators which control the positioning of the cursors on the monitor screen.

discharging of C through a fixed resistance to the h.t. positive line. Since the grid voltage varies only slightly a constant voltage is applied to R and the rate of discharge of C is also constant. The time taken for the anode volts to bottom is therefore proportional to the anode voltage when the trigger pulse ceased, i.e., is proportional to the setting of the delay control potentiometer. The sharp drop in screen volts at the conclusion of this period is used to trigger the generators providing the black out pulses forming the cursors.

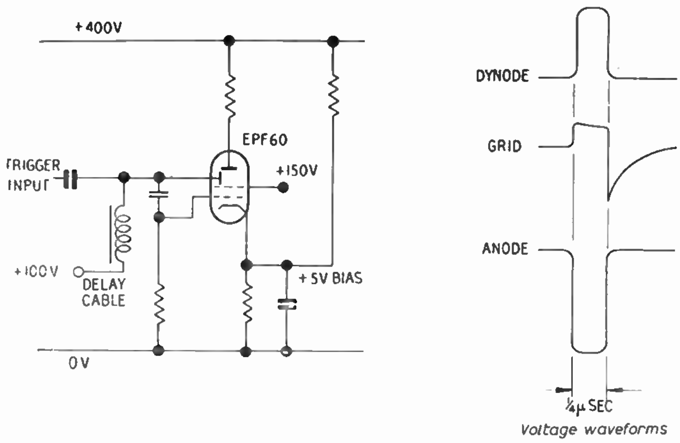

Fig. 9. Short-pulse generator.

The time delay generator circuit and waveforms are shown in Fig. 8. When the amplified trigger pulses derived from the television sync. pulses are impressed on the suppressor, the anode voltage rises to a value at which the diode D conducts, the value being controlled by the voltage set on the delay control potentiometer. As a consequence the capacitor C becomes charged to an extent also controlled by this setting. With the trigger pulse removed the anode voltage falls at a rate determined by the

The circuit of these generators is shown in Fig. 9. In order to obtain large sharp pulses, the circuit is essentially a positive feedback system based on a secondary emission pentode. Pulses are obtained of duration equal to twice the delay time of the delay line since the valve is effective only while the positive feedback voltage affects the dynode, i.e., for the time required to travel down the delay line and be reflected back to the dynode line by the short circuited end.

World Radio History

System Design Criteria for Space Television†

by

A. J. VITERBI, B.S., M.S.‡

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

Summary: A narrow band television system for relaying to earth images of the planets is described. The principal consideration is the necessity of communicating over extremely long ranges. Because of the resulting high noise environment, the channel bandwidth is severely restricted. Bandwidth compression is achieved by storing the video signal on magnetic tape or photographic film and transmitting at a reduced information rate.

An f.m.-p.m. telemetry system is utilized. The doppler-shifted carrier is recovered from the noise by a very narrow tracking filter or phase-locked loop. The derived carrier is mixed with the incoming signal to yield the noisy video information. The discriminator is also a phase-locked loop with sufficient bandwidth to track the information signal. The recording operation is reversed at the receiver to present the video image at periodic intervals, or a facsimile technique may be employed. The main interest of the paper is directed toward the theory and design of the very narrow band telemetry system and the unconventional discriminator required for ultra-long-range television communication.

1. Introduction

The initial probings of man into space require sensors capable of functioning in extraterrestial environments and communications systems capable of transmitting information over extremely long ranges. Obtaining visual images of the planets represents one of the more challenging early experiments.

To establish the requirements for the design of a space television system for a particular mission, it is first necessary to determine the signal levels and noise environment through which communications must be secured. If the space vehicle is at a range R from earth, the received signal power is given by

$$P_S = \frac{P_T G_T}{4\pi R^2} A_r \qquad \text{.........(1)}$$

where P_T is the transmitter power, G_T the transmitter antenna gain, and A_r the effective area of the receiving antenna. §

For a typical mission to Venus, the minimum transmission range is of the order of 25 million miles. If it is assumed that 10 watts of telemetry sideband power are available (over and above the carrier power required for vehicle tracking), a vehicle antenna can be beamed in the direction of the earth with a gain of 10 db, and a receiving antenna with an 85-ft.-diameter parabolic reflector and 60 per cent efficiency is utilized, the received power would be about $1{\cdot}6 \times 10^{-18}$ watts, or -148 dbm.

At the same time, the noise power per unit bandwidth at the receiver input is given by

$$\frac{P_N}{B} = kT\,(NF) \qquad \text{.........(2)}$$

where k is Boltzmann's constant, T the absolute temperature of the ambient (assumed 293°K), B the receiver bandwidth, and NF the effective noise figure of the receiving system.

With a noise figure of 6 db, the noise power per unit bandwidth is $1{\cdot}6 \times 10^{-20}$ watts/cycles/sec, or -168 dbm/cycles/sec.

It will prove convenient to define a spectral density (Φ) as the ratio of noise power per unit bandwidth to received signal

$$\Phi = \frac{P_N}{BP_S} \qquad \text{.........(3)}$$

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‡ Jet Propulsion Laboratory, California Institute of Technology, Pasadena 3, California, U.S.A.

U.D.C. No. 621.397:621.398:629.19

§A list of symbols used in this paper, together with their meaning, is given in the Appendix.

World Radio History

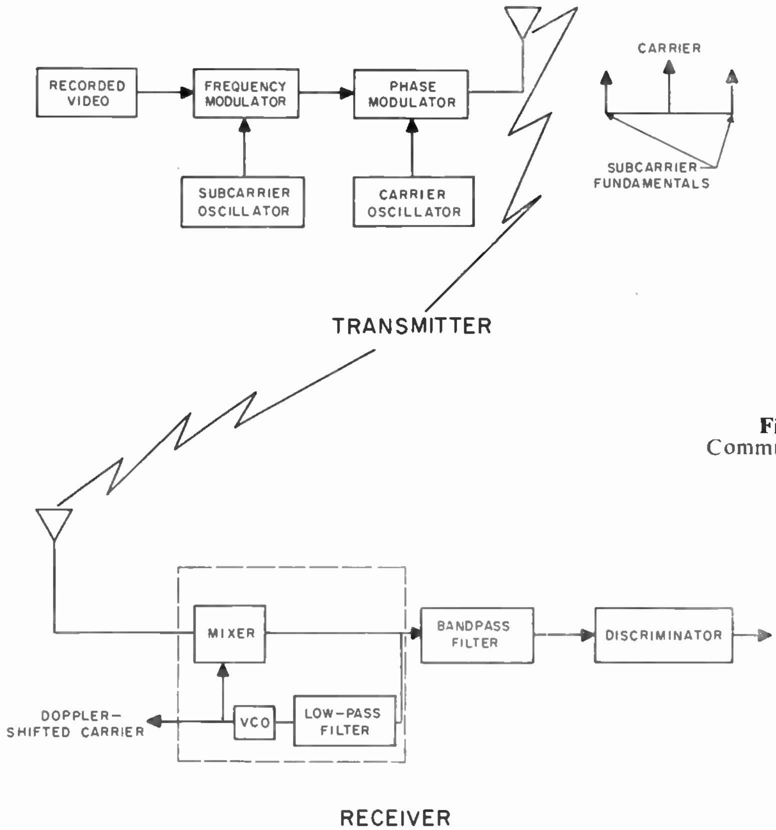

Fig. 1. F.M./P.M. Communication System.

For the hypothetical conditions discussed, the spectral density is about 10^{-2} rad^2/c/s (equivalent to a signal-to-noise ratio of 20 db in 1-c/s band).

An f.m.-p.m. communications system is envisaged for the dual purpose of tracking the space vehicle and telemetering the video information. As shown in Fig. 1, the recorded video signal is used to frequency-modulate a sub-carrier oscillator. The resulting signal is used to phase-modulate the transmitter carrier. The first detector of the receiver is a coherent tracking filter capable of recovering the doppler-shifted carrier from the noisy input. The carrier is then mixed with the input to detect the noisy video signal which is filtered by the discriminator.

A particular discriminator design is described which shows that if the quality requirement that eight grey levels be discernible is placed on the received image, the transmission bandwidth is limited to about 1 c/s. Bandwidth compression is achieved by storing the video information on magnetic tape or photographic film and transmitting it for an arbitrary length of time. On earth, the process is reversed and an image is flashed on the screen at periodic intervals. In the following Sections, the theory and design of the narrow-band discriminator and its effect on the over-all system design are discussed.

2. Discriminator Operation

The key to recovering narrow-band signals from high-noise environments lies in the operation of the phase-locked-loop discriminator (Fig. 2).[1, 2] Its function and design depend, of course, on the characteristics of the frequency-modulated signal which it must detect. Since the video signal is necessarily filtered before modulation, the cut-off frequency of the signal spectrum corresponds approximately to the minimum rise-time or maximum slope that the signal waveform can have at any given time. Hence, the design of the discriminator should be based on receiving a sinusoidal signal which is frequency-modulated by a ramp waveform of

World Radio History

the maximum slope expected in the presence of the highest-noise environment to be encountered. The discriminator may be viewed essentially as a servomechanism in which a frequency-variable oscillator is controlled by the frequency of the incoming signal. In the process of controlling the oscillator, a voltage proportional to frequency is produced, thus yielding the discriminator action. The real advantage of the loop over conventional detection methods lies in its noise-squelching properties, made possible by the narrow-band operation of the phase-locked-loop discriminator.

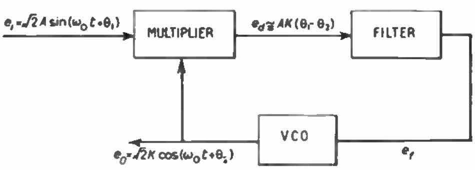

Fig. 2. Phase-locked loop.

The noise-free operation of the device is as shown in Fig. 2. The input signal is multiplied by the output of the controlled oscillator. The latter produces a phase shift of 90 deg, so that in quiescent operation, when the input signal is an unmodulated sine wave, the output is a cosine wave. The multiplier output e_d can be written as follows:

$$e_d = e_i e_o = 2AK \quad \sin\left[\omega_0 t + \theta_1(t)\right] \quad \cos\left[\omega_0 t + \theta_2(t)\right]$$
$$= AK \quad \sin\left[\theta_1(t) - \theta_2(t)\right] + \sin\left[2\omega_0 + \theta_1(t) + \theta_2(t)\right] \quad \text{.........(4)}$$

The low-pass filter of the loop discards the higher-frequency term. When the phase difference between input and output is small compared to 90 deg, the approximation, $\sin(\theta_1 - \theta_2) \cong (\theta_1 - \theta_2)$ is valid. Then the error voltage e_d becomes effectively:

$$e_d \cong AK\ [\theta_1\ (t) - \theta_2\ (t)] \quad \text{.........(5)}$$

and the multiplier is shown to be a phase detector.

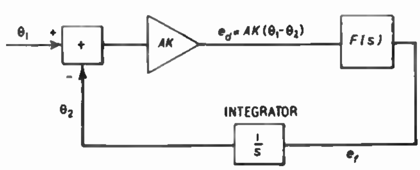

Fig. 3. Linear model of loop.

As is generally the case in the analysis of servomechanisms, it is convenient to replace the physical device by its linear model, which puts in evidence the pertinent quantity which is to be controlled. Equation (5) shows that the error voltage (e_d) is a constant multiple of the difference between input- and output-signal phases. Hence, in Fig. 3 the input and output signals are replaced by their phases, and an adder is substituted for the multiplier. Since the parameter of interest at the output of the voltage-controlled oscillator is phase and its input voltage is proportional to frequency, the oscillator must be replaced in the model by a simple integrator.

To analyse the operation of the discriminator with a noisy input by means of a simple linear model, certain additional assumptions must be made:

(1) The propagation of signal phase modulation and that of random noise through the network are independent; this is equivalent to stating that the superposition principle holds for linear systems.

(2) In order for the first assumption to be at least approximately fulfilled, as well as for reasons discussed elsewhere in this section, both the phase difference $(\theta_1 - \theta_2)$ and the output-phase jitter σ_v must at all times be small compared with 1 rad.

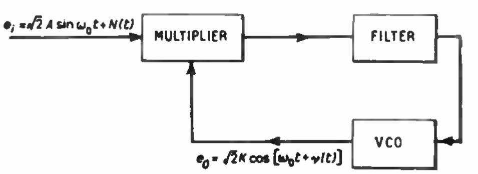

Fig. 4. Operation with additive noise and no modulation.

Based on assumption (1), the discriminator operation will be considered in the presence of noise and without signal modulation. The pertinent signals are shown in Fig. 4. The input signal is

$$e_i = \sqrt{2}\,A\ \sin\omega_0 t + N\ (t) \quad \text{.........(6)}$$

where $N(t)$ represents the additive noise.

World Radio History

The phase output voltage must be

$$\begin{aligned} e_o &= \sqrt{2}\,K \cos[\omega_0 t + v(t)] \\ &= \sqrt{2}\,K[\cos\omega_0 t \cos v(t) - \sin\omega_0 t \sin v(t)] \\ &\cong \sqrt{2}\,K[\cos\omega_0 t - v(t)\sin\omega_0 t] \qquad \ldots\ldots\ldots(7) \end{aligned}$$

where the phase noise jitter $v(t)$ is small compared to 1 rad, thus making the approximation valid.

The phase-difference voltage is the product of the two terms

$$e_d = e_i e_o \cong AK \sin 2\omega_0 t - AK\,v(t)[1 - \cos 2\omega_0 t] + \sqrt{2}\,KN(t)[\cos\omega_0 t - v(t)\sin\omega_0 t]$$

This expression may be simplified according to a further approximation based on the following observations: The loop filter will not pass the double frequency terms; $v(t)$ is a slowly varying function, at all times small compared to 1 rad, while $N(t)$ is wide-band white noise centered about a frequency ω_0. Then

$$e_d(t) \cong -AK\,v(t) + \sqrt{2}\,K \cos\omega_0 t\, N(t) \ldots(8)$$

The product $v(t)\,N(t)\sin\omega_0 t$ is considered negligible relative to $N(t)\cos\omega_0 t$ since $v(t) << 1$ rad. The first term of eqn. (8) is simply the low-frequency output-phase noise amplified by AK. The second term, the product of the input noise and the carrier, has a low-frequency component of flat power spectral density equal to $P_N K^2/B$, plus higher-frequency components centered about twice the carrier frequency (which may be neglected since these components will not pass through the loop filter). Then $e_d(t)$ is effectively AK times the difference between a low-frequency input-noise signal of power density $P_N/A^2B = \Phi$ and the phase jitter $v(t)$. Hence, the linear model of Fig. 3 is suited to the description of noise propagation through the loop. Linear-transform methods may be used to compute the output phase-noise power as well as the output signals.

The oscillator output phase and the frequency-controlling signal may be obtained from Fig. 3 in terms of the input phase:

$$\left.\begin{aligned} H(s) &= \frac{\theta_2(s)}{\theta_1(s)} = \frac{AKF(s)}{s + AKF(s)} \\ \frac{e_f(s)}{\theta_1(s)} &= s\,H(s) \end{aligned}\right\} \ldots(9)$$

Of greater interest than the phase-noise spectral density is the total noise or mean-square jitter σ_v^2, which is the integral of the output density.

Thus,

$$\sigma_v^2 = \Phi\int_{-\infty}^{\infty}\frac{|H(\omega)|^2}{2\pi}\,d\omega = 2\Phi\int_0^{\infty}\frac{|H(\omega)|^2}{2\pi}\,d\omega = 2\Phi B_L \qquad \ldots\ldots\ldots(10)$$

The integral $\frac{1}{2\pi}\int_0^{\infty} |H(\omega)|^2\, d\omega$ is defined as the loop-noise band-width B_L, since an ideal low-pass filter of bandwidth B_L c/s having a white-noise input of spectral density Φ will produce the same amount of noise as is present at the loop-phase output.

3. Phase-locked Loop Design

In the design of any closed-loop control system, whether noise is present or not, a restriction must be placed on the magnitude of the transient error. This is particularly important in the present case for two reasons. First, the linear model requires a small phase difference and a small phase-noise jitter; even more important in the actual operation is the fact that if the phase difference ever exceeds 90 deg, the multiplier ceases to function as a phase detector, and phase-lock is temporarily lost. At this point, the oscillator is no longer controlled, and discriminator action has ceased. The loop has automatic search properties, however, which permit the oscillator to achieve phase-lock even when it is initially uncontrolled, provided the incoming frequency lies within the pull-in range of the phase-locked loop.

When noise is present at the input, the output-phase jitter adds to the transient phase difference to produce an instantaneous error signal of $[\theta_1(t) - \theta_2(t) - v(t)]$.

The mean-square error is then

$$\lim_{T\to\infty}\frac{1}{2T}\left\{\int_{-T}^{T}[\theta_1(t)-\theta_2(t)]^2\,dt + \int_{-T}^{T} v^2(t)\,dt\right\} \ldots(11)$$

The peak error is

$$|\theta_1 - \theta_2|_{max} + |v|_{max}\,\text{rad} \qquad \ldots\ldots\ldots\ldots(12)$$

For a particular maximum signal-to-noise ratio, the loop filter can be so designed that it will keep either the mean-square error or the peak error at a minimum. An optimum linear system which minimizes the mean-square error may be synthesized using the Wiener integral

World Radio History

equation.[1] However, the peak error criterion is more compatible with the design procedure to be developed here.

On this basis, the form of the loop filter will now be derived. Since the modulating frequency is assumed never to exceed a maximum time rate of change, $\dot{f}$ rad/sec^2, the peak transient error which can occur is that caused by a frequency ramp of slope $\dot{f}$. This corresponds to a phase input $1/2\,\dot{f}t^2$ whose transform is $\dot{f}/s^3$. It is well known that a servo-loop can track an acceleration input with finite error only if the open-loop transfer function provides two integrations. One virtual integration is provided by the voltage controlled oscillator (see Fig. 3); thus, the loop filter must produce the other. For stability reasons, the filter must also have one zero in the finite left-half s-plane. The loop-filter transfer function may have any number of other poles and zeros in the left-half plane, but since they are not required, they will be omitted for the sake of the simplicity of filter. The form of the loop filter has thus been determined as

$$F(s)=\frac{1+as}{s} \quad \ldots\ldots\ldots\ldots\ldots(13)$$

where the assumption is made that if the filter amplifier has a gain other than unity, it is lumped into the multiplicative factor K. The phase transfer function

$$H(s)=\frac{\theta_2(s)}{\theta_1(s)}=\frac{AK(1+as)}{s^2+AK(as+1)} \ldots\ldots(14)$$

It is found experimentally that if the damping factor of the loop $(a/2)\sqrt{(AK)}$ is made equal to $1/\sqrt{2}=0{\cdot}707$, a satisfactory balance is achieved between the transient error and the output noise. The transfer function then becomes

$$H(s)=\frac{1+\sqrt{\dfrac{2}{AK}}\,s}{1+\sqrt{\dfrac{2}{AK}}\,s+\dfrac{s^2}{AK}} \ldots\ldots\ldots(15)$$

The filter transfer function is then

$$F(s)=\frac{1+\sqrt{\dfrac{2}{AK}}\,s}{s} \quad \ldots\ldots\ldots(16)$$

The loop bandwidth can be computed from eqn. (10) as

$$B_L=\frac{1}{2\pi}\int_0^\infty |H(\omega)|^2\,d\omega$$
$$=\frac{3}{4\sqrt{2}}\sqrt{AK}\ \text{cycles/sec} \quad \ldots\ldots\ldots(17)$$

The transfer functions can be expressed in terms of the loop bandwidth by substituting eqn. (17) into the previous expressions:

$$H(s)=\frac{1+\dfrac{3}{4B_L}s}{1+\dfrac{3}{4B_L}s+\dfrac{9}{32B_L^2}s^2} \quad \ldots\ldots(18)$$

$$F(s)=\frac{1+\dfrac{3}{4B_L}s}{s} \quad \ldots\ldots\ldots(19)$$

The transform of the phase error is found in terms of the transfer function as

$$[\theta_1(s)-\theta_2(s)]=\theta_1(s)[1-H(s)] \quad \ldots\ldots(20)$$

If a frequency ramp is applied at the input

$$[\theta_1(s)-\theta_2(s)]=\frac{\dot{f}}{s^3}\left(\frac{\dfrac{9}{32B_L^2}s^2}{1+\dfrac{3}{4B_L}s+\dfrac{9}{32B_L^2}s^2}\right) \quad (21)$$

If the inverse transform is taken, it is found that the peak phase error is only slightly higher than the steady-state error, since the damping factor is high enough to keep the oscillations quite low. The steady-state error is

$$E_{ss}=\left(\theta_{1ss}-\theta_{2ss}\right)=\frac{9\dot{f}}{32B_L^2}\ \text{rad} \quad \ldots\ldots(22)$$

So far only the question of phase lock has been considered. Of equal importance to the functioning of the discriminator is how closely the frequency-proportional output e_f follows the instantaneous frequency of the input. Since

$$\frac{e_f(s)}{\theta_1(s)}=sH(s)=\frac{s\left(1+\dfrac{3}{4B_L}s\right)}{1+\dfrac{3}{4B_L}s+\dfrac{9}{32B_L^2}s^2} \quad (23)$$

in the absence of noise the output is a filtered, replica of the derivative of the input phase, which in turn is the modulating wave-form. However, it is clear that the transfer function from input to frequency output does not approach zero for infinite frequencies and hence

World Radio History

does not produce filtering for wide-band noise. Removal of the finite numerator zero of eqn. (23) by means of a simple low-pass filter at the frequency output produces the necessary filtering. The resulting configuration appears in Fig. 5. The discriminator transfer function becomes

$$\frac{e_f'}{\theta_1} = \frac{s}{1+\dfrac{3s}{4B_L}+\dfrac{9}{32B_L^2}s^2} \qquad \ldots\ldots\ldots(24)$$

and the total output noise or mean-square-frequency jitter is

$$\sigma^2 = \frac{\Phi}{2\pi}\int_{-\infty}^{\infty}\left|\frac{e_f'}{\theta_1}(\omega)\right|^2 \mathrm{d}\omega = \frac{64}{27}\Phi B_L^3\left(\frac{\text{rad}}{\text{sec}}\right)^2 \quad \ldots(25)$$

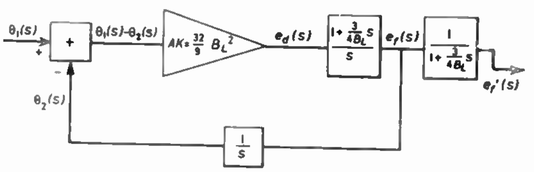

Fig. 5. Complete loop design.

All the noise terms have been related to the loop-noise band-width. The loop bandwidth will be a key parameter in the design. For a given maximum frequency slope, $\dot{f}$ rad/sec², and the greatest permissible steady-state phase error (which is approximately equal to the peak phase error), eqn. (22) yields the loop-bandwidth value which should be chosen.

4. Limiter Operation

Another component of the receiver which is an integral part of the discriminator is the bandpass limiter preceding the phase-locked loop. It has been shown[3] that if a noisy sinusoid is applied at the input of a bandpass limiter, the output signal-to-noise ratio is a constant multiple of the input ratio:

$$\left(\frac{P_S}{P_N}\right)_0 = \frac{\pi}{4}\left(\frac{P_S}{P_N}\right)_I \qquad \ldots\ldots\ldots(26)$$

It is also characteristic of the limiter that for all levels of input noise the total output power is constant:

$$(P_S+P_N)_0 = \text{constant}$$

If the input signal power is held fixed and only the noise level varies, the output-signal power, in the absence of noise, may be equated to the sum of signal-plus-noise power at the output when the input is noisy:

$$(P_S+P_N)_0 = [(P_S)_I]_{\text{noiseless}} = [(P_S)_0]_{\text{noiseless}} \quad \ldots(27)$$

where the gain of the limiter is such that in the absence of noise the output power equals the input-signal power. Combining eqns. (26) and (27),

$$\frac{(P_S)_0}{\left[(P_S)_0\right]_{\text{noiseless}}} = \frac{(P_S)_0}{(P_S+P_N)_0} = \frac{1}{1+\left(\dfrac{P_N}{P_S}\right)_0} = \frac{1}{1+\dfrac{4}{\pi}\left(\dfrac{P_N}{P_S}\right)_I} \qquad \ldots\ldots\ldots\ldots(28)$$

Following the notation of the previous Sections for the signal and noise power into the limiter,

$$(P_S)_I = [(P_S)_0]_{\text{noiseless}} = A^2$$
$$(P_N)_I = P_N$$

whence

$$\frac{(P_S)_0}{A^2} = \frac{1}{1+\dfrac{4}{\pi}\dfrac{P_N}{A^2}} = \alpha^2 \qquad \ldots\ldots(29)$$

where α is defined as the suppression factor.

Hence, if the noisy input to the discriminator passes through a bandpass limiter prior to the phase-locked loop, the limiter has the effect of reducing the signal power from A^2 to $\alpha^2 A^2$.

As a consequence of the presence of the limiter, the loop gain is reduced from AK to αAK; as the channel noise is increased (eqn. 29), the loop-noise bandwidth becomes narrower (eqn. 17), and vice versa. Of course, when the noise bandwidth narrows as the input noise increases, the transient error will also increase. This requires that the system design be based on the maximum-noise (minimum-loop-bandwidth) conditions. Thus, the loop bandwidth determined in the design procedure which follows is

$$B_{L_0} = \frac{3}{4\sqrt{2}}\sqrt{\alpha_0 AK}\ \text{cycles/sec} \ \ldots\ldots(30)$$

where the subscripts refer to threshold and α_0 is determined from eqn. (29) for the minimum signal-to-noise ratio expected.

5. System Design

Considering the ideas developed in the previous Sections, it is clear that the telemetry system design must be based on the phase-locked-loop capabilities. First of all, the worst noise conditions should be assumed; that is, the spectral density used should be that resulting from communication at the greatest range expected.

One of the prime considerations must be that the loop remain in lock most of the time so that a meaningful output will be obtained. Hence, a restriction on the peak error must be established. The peak phase error due to the signal modulation was shown to be approximately

$$(\theta_1-\theta_2)_{max} \cong (\theta_1-\theta_2)_{ss} = \frac{9\dot{f}}{32\,B_{L_0}^2}\,\text{rad} \quad \ldots(31)$$

where B_{L_0} is the narrowest loop-noise bandwidth expected which results from the threshold signal-to-noise conditions.

The threshold-phase noise power is

$$\sigma_i^2 = 2\Phi B_{L0}\,\text{rad}^2 \quad \ldots\ldots\ldots(32)$$

If the phase noise is normally distributed, the instantaneous noise will be less than $2\sigma_v$ for 95 per cent of the time. Hence, the restriction

$$\frac{9\dot{f}}{32\,B_{L_0}^2} + 2\sqrt{2}\,\Phi^{1/2}B_{L_0}^{1/2} \leqslant \frac{\pi}{2}\,\text{rad} \quad \ldots(33)$$

ensures that the loop will be in lock most of the time. A more conservative approach would substitute $3\sigma_v$ for $2\sigma_v$ and reduce the maximum allowable peak error to less than $\frac{1}{3}\pi$ rad.

The other performance criterion which must be put on a quantitative basis is the data accuracy at the output. This data accuracy, or frequency-output signal-to-noise ratio, is obviously proportional to $\triangle/\sigma_f$, where $\triangle$ is the total modulating-frequency deviation. The expression for the frequency jitter

$$\sigma_f = \frac{8}{3\sqrt{3}}\Phi^{1/2}B_{L_0}^{3/2}\,\text{rad} \quad \ldots\ldots\ldots(34)$$

was found in the previous Section. If the data-accuracy ratio $\triangle/\sigma_f$ is set at some desired value D,

$$\triangle = \frac{8}{3\sqrt{3}}D\Phi^{1/2}B_{L_0}^{3/2}\,\frac{\text{rad}}{\text{sec}} \quad \ldots\ldots\ldots(35)$$

The design restrictions on the problem are represented by eqns. (33) and (35):

(1) The threshold-loop bandwidth must be sufficiently narrow at the maximum communicating range that the noise jitter σ_v will be considerably less than $\frac{1}{2}\pi$ rad.

(2) However, the smaller B_{L_0} is made, the shallower is the slope of the modulation which can be tolerated if eqn. (33) is to be satisfied.

(3) Also, the total frequency deviation must be made sufficiently large for any given required data accuracy.

The maximum slope of the modulating signal $\dot{f}$, together with the frequency deviation, determines the magnitude of the most important property of the communication system: the maximum information rate or bandwidth achievable. The information rate may be defined as the inverse of the time required for a frequency ramp of slope $\dot{f}$ rad/sec² to vary over a frequency range of $\triangle$ rad/sec. Thus,

$$R_i \equiv \frac{\dot{f}}{\triangle}\,\frac{\text{rad}}{\text{sec}} \quad \ldots\ldots\ldots(36)$$

The information bandwidth is closely related to this:

$$B_i \equiv \frac{R_i}{2\pi} = \frac{\dot{f}}{2\pi\triangle}\,\text{cycles/sec} \quad \ldots\ldots\ldots(37)$$

It should be noted at this point that B_L and B are totally separate quantities: B_L is the noise bandwidth selected to maintain phase lock for a given channel signal-to-noise ratio, while B_i is the maximum data bandwidth which results from this choice of B_L and the accuracy requirements. The expression for information rate is readily arrived at by combining eqns. (33), (35), and (36).

$$R_i = \frac{\dot{f}}{\triangle} = \frac{\frac{32}{9}B_{L_0}^2\left(\frac{\pi}{2} - 2\sqrt{2}\Phi^{1/2}B_{L_0}^{1/2}\right)}{\frac{8}{3\sqrt{3}}D\,\Phi^{1/2}\,B_{L_0}^{3/2}} \quad \ldots(38)$$

$$= \frac{1}{D\sqrt{3}}\left(\frac{2\pi B_{L_0}^{1/2}}{\Phi^{1/2}} - 8\sqrt{2}\,B_{L_0}\right)\frac{\text{rad}}{\text{sec}}$$

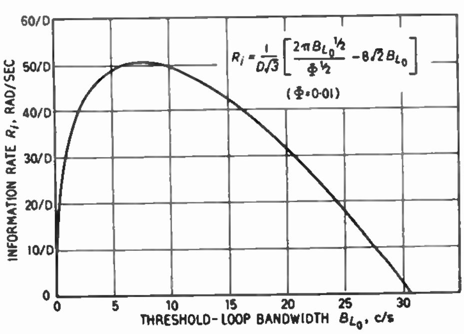

Fig. 6. Information rate as a function of threshold-loop bandwidth.

World Radio History

It is of interest to plot this expression for information rate as a function of threshold-loop-noise bandwidth for a given value of channel spectral noise density. This appears in Fig. 6 for $\Phi = 0{\cdot}01$, which corresponds to the maximum noise involved in the Venus television experiment to be discussed in the next Section. It is seen from this graph that the information rate achieves a peak for a loop-noise bandwidth of 7·7 c/s. This, then, is the optimum value of B_{L_0}. In general, it can be shown by setting the derivative $\partial R_i/\partial B_{L0}$ equal to zero that the maximum rate is achieved when

$$B_{L_0} = \frac{\pi^2}{128\Phi} \qquad \text{.........(39)}$$

for which value

$$R_{imax} = \frac{\pi^2}{16}\sqrt{\frac{2}{3}}\,\frac{1}{D\Phi} \cong \frac{1}{2D\Phi}\;\frac{\text{rad}}{\text{sec}} \qquad \text{......(40)}$$

Thus, an optimum design procedure has been developed for determining loop bandwidth and information rate. The resulting frequency deviation and modulating signal slopes are obtained from these two quantities.

To summarize, the procedure consists of the following steps:

(1) The noise spectral density normalized by the received signal power Φ at the projected communications threshold range is derived using eqn. (1).

(2) The accuracy required for the information to be telemetered ($D = \Delta/\sigma_f$) is specified.

(3) Using eqns. (39) and (40), the maximum information rate R_i and loop bandwidth B_{L0} at threshold are determined, or R_i is plotted as a function of B_{L_0} from eqn. (38).

(4) Total deviation Δ is determined from eqn. (35).

(5) Other quantities of interest may now be obtained. The maximum slope $\dot{f} = R_i\Delta$; the threshold loop gain $\alpha_0 AK$ is derived from B_{L_0}, using eqn. (30).

(6) The suppression factor at threshold α_0 is obtained from eqn. (29). The term P_N/A^2 in this equation is the product of Φ and the noise bandwidth of the predetection bandpass filter preceding the limiter (Fig. 1). For lesser noise levels, α can also be determined from eqn. (29).

6. Design Example

To return to the problem of transmitting a television image from the vicinity of Venus, it is possible to apply the principles that have been derived.

Following the design steps specified in the previous section:

(1) The normalized spectral density was found to be 0·01 rad²/cycles/sec.

(2) Approximately eight levels of grey are to be discriminated in the received photograph. This requires that $D \cong 8$.

(3) For $\Phi = 0{\cdot}01$, R_i is plotted against B_{L0} in Fig. 6. From the graph, or using eqn. (40), it is found that the maximum possible information rate (R_i) is 6·25 rad/sec, which corresponds to an information bandwidth of about 1 c/s. The threshold-loop bandwidth for this condition is 7·7 c/s.

(4) For this value of B_{L_0}, the total frequency deviation is

$$\Delta = 26{\cdot}3\,\frac{\text{rad}}{\text{sec}} = 4{\cdot}19 \text{ c/s.}$$

and the output-frequency jitter is 3·3 rad/sec = 0·52 c/s r.m.s.

(5) The maximum slope

$$\dot{f} = R_i\Delta = 164\,\frac{\text{rad}}{\text{sec}^2} = 26{\cdot}2\,\frac{\text{c/s}}{\text{sec}}$$

The threshold gain

$$\alpha_0\, AK = \frac{9}{32}\, B_{L_0}{}^2 = 16{\cdot}7$$

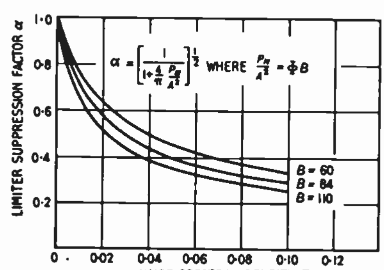

Fig. 7. Limiter suppression factor as a function of noise spectral density for I.R.I.G. (Inter-Range Instrumentation Group) Channels 1, 2 and 3.

(6) The suppression factor α is plotted in Fig. 7 as a function of Φ for predetection-noise bandwidths of 60, 84, and 110 c/s.

World Radio History

At threshold ($\Phi = 0{\cdot}01$), using a 60-c/s bandwidth, $\alpha_0 = 0{\cdot}34$. The loop gain in the absence of noise must then be

$$AK = \frac{16{\cdot}7}{0{\cdot}34} = 49{\cdot}2$$

One of the most important parameters of the television system is the time required to transmit the image. Assuming that a 200-line scan is used and that a resolution of 200 elements per line is desired, it will be necessary to transmit 40,000 separate picture elements. Since the maximum information rate was found to be 6·25 rad/sec, each element will require 0·16 sec to be transmitted. Thus, the time required to send all 40,000 elements of the image is about 1·85 hr.

7.Discriminator Parameters and Mechanization

For the design example under consideration it is of interest ot derive the parameters of the loop fi:ter and discuss the operation of the multiplier and voltage-controlled oscillator. The transfer function of the loop filter is given by eqn. (16) as

$$F(s) = \frac{1 + as}{s} \quad \text{where } a = \sqrt{\frac{2}{AK}}$$

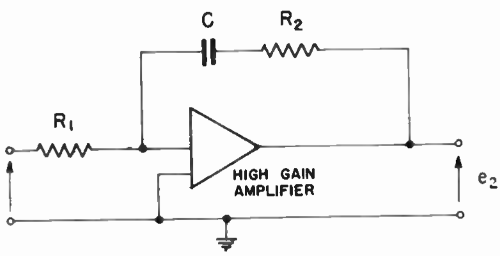

Fig. 8. Loop filter mechanization.

This can be closely approximated by an operational amplifier as shown schematically in Fig. 8. If the gain of the amplifier is very high, the voltage transfer ratio is approximately

$$\frac{e_2(s)}{e_1(s)} \cong \frac{R_2 + 1/Cs}{R_1} = \frac{1 + R_2Cs}{R_1\,Cs} \qquad \ldots(41)$$

$$\text{Then } R_2C = a = \sqrt{\frac{2}{AK}} \qquad \ldots\ldots\ldots(42)$$

while R_1C is a scale factor which may be lumped into the gain factor K.

To determine K, the multiplier and the VCO operation must be considered. The multiplier may be as simple as a diode or transistor switch. Its output for a phase difference of 1 radian between input and output is K_M volts. The VCO may be a reactance tube controlled oscillator whose output for a variation of 1 volt at its input is K_V c/s. The factor K is then given as the product

$$K = 2\pi\, K_M\, K_V\, \frac{1}{R_1C} \qquad \ldots\ldots\ldots\ldots(43)$$

where K_M has the dimensions $\dfrac{\text{volts}}{\text{rad}}$ and K_V the dimensions $\dfrac{\text{cycles/sec.}}{\text{volt}}$ The factor 2π is required so that K will have the dimensions rad./sec.

To return to the design example of the previous Section, AK was found to be 49·2. For a typical set of components

$$K_M = 14\frac{\text{volts}}{\text{rad}}$$

$$K_V = 5\,\frac{\text{c/s}}{\text{volt}}$$

The input to the phase-lock loop from the band-pass limiter shall be $A = 1$ volt r.m.s. Then $K = 49{\cdot}2\,\dfrac{\text{rad.}}{\text{sec}}$ If C is chosen to be 10 microfarads, eqn. (43) gives

$$49{\cdot}2 = K = \frac{2\pi(14)(5)}{R_1 10^{-5}}$$

whence $R_1 = 890\text{k}\,\Omega$

Also from eqn. (42)

$$R_2 = \frac{1}{C}\sqrt{\frac{2}{AK}} \simeq 20\text{k}\,\Omega$$

The transfer function of the low pass filter on the output of the loop has its denominator equal to the numerator of $F(s)$ (see Fig. 5).

$$\text{Thus} \qquad \frac{e_f'(s)}{e_f(s)} = \frac{1}{1 + R_2Cs} \qquad \ldots\ldots\ldots(44)$$

where the product of R_2 and C is 0·2, and this can be realized as a low-pass filter as shown in Fig. 9.

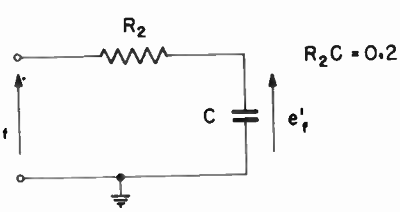

Fig. 9. Output filter mechanization.

World Radio History

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10. Appendix: List of Symbols

A	= r.m.s. signal voltage.
AK	= loop gain.
A_r	= effective antenna area.
B	= predetection receiver bandwidth.
B_i	= information bandwidth.
B_L	= loop bandwidth.
B_{L_0}	= threshold-loop bandwidth.
D	= data accuracy.
e_d	= error voltage.
e_f	= frequency-controlling voltage.
e_f'	= discriminator output.
e_i	= input signal.
e_o	= phase-output signal.
E_{ss}	= steady-state error.
f	= maximum slope of modulating signal.
$F(s)$, $F(\omega)$	= loop-filter transfer function.
G_T	= transmitter antenna gain.
$H(s)$, $H(\omega)$	= loop-phase transfer function.
K	= r.m.s. output voltage.
k	= Boltzmann's constant.
$N(t)$	= noise signal.
NF	= effective receiver noise figure.
P_N	= noise power in predetection bandwidth.
P_S	= received signal power = A^2.
P_T	= transmitter power.
R	= range from earth.
R_i	= information rate.
T	= absolute temperature of the ambient.
α	= limiter suppression factor.
α_0	= limiter suppression factor at threshold.
Δ	= total frequency deviation.
θ_1	= input-signal phase.
θ_2	= output-signal phase.
$v(t)$	= output-phase jitter.
σ_f	= r.m.s. discriminator output noise.
σ_v	= r.m.s. output-phase noise.
Φ	= noise spectral density.
ω_0	= modulation subcarrier frequency.

World Radio History

Switching Circuits using Bi-Directional Non-Linear Impedances†

by

T. B. TOMLINSON, PH.D., B.SC., ASSOCIATE MEMBER ‡

Summary : In a brief general review of logic circuitry, a case is developed for a bi-directional, non-linear switching element. In order to compare circuits using such elements with those using standard semiconductor diodes, the main features of diode gates are considered in detail. Particular attention is paid to design of p-n-p transistor driver stages and their dependence on the logic sequence. Circuits using a bidirectional "constant voltage" element are described, including a "two-decision" AND gate. Possible types of constant voltage element are discussed and experimental results are given for multi-electrode silicon carbide devices. An interesting feature of the latter is the non-linear behaviour of capacitance. An attempt is made to compare gating circuits employing constant voltage, bi-directional elements with standard diode gates from a performance/cost point of view.

A "majority-logic" gate is made possible by a "constant current," bidirectional element. If one input is used as a control, this gate can become a many function gate, thus a 3-input gate can be controlled to act as a 3-AND, an OR, or a "2 or more out of 3" logical unit without change of input or output connections. Some possible realisations of the constant current device are discusssed. A binary-octal decoder circuit and a simple binary full adder circuit are given as examples of the application of the non-linear elements and to illustrate their unusual features.

1. Introduction

The term "switching circuit" implies an on/off arrangement, i.e. a circuit containing a component which has two states, for example, the conducting and non-conducting states of a relay switch. Nowadays, the term is taken to embody electronic switching circuits and it is also used synonymously with "logical circuits" such as are used in the arithmetic and control units of a digital computer. It thus embraces a multitude of circuit combinations which operate on binary signals. For this reason, the author was caused to think of the wider implications where the two states were taken to be "signal on" and "signal off" rather than "current on" and "current off." This signal might be conveyed by a level of voltage or current or magnetic field or other suitable physical phenomenon. Thus, in the more general aspect of switching one can include "steering networks" which cause signals to follow appropriate paths through a network.

Where complex networks are involved it is generally considered to be more economic to perform the bulk of the logic in passive elements and diodes have been the most popular device for this purpose. When the signal level falls to a value too low for precise operation of subsequent logic circuits it is amplified to restore it to the appropriate level. Now the most economic way of using such amplifiers is to operate them in the on/off mode. This has the advantage of reducing power dissipation in the amplifying device, whether valve or transistor, and also of generating output signals which alternate between two distinct levels. As the amplifiers are to work in conjunction with the passive logical elements, it appears at first sight that there would be a natural combination when the latter are also simple on/off devices. However, in certain of the logical circuits, e.g. a diode AND gate, the amplifier is "on" when the following switching element is "off" and vice versa. There is consequently a case for looking into the possibilities of switching elements other than simple on/off types especially where there is a possibility of utilising a much cheaper device. In order to perform signal steering operations the devices must possess some non-linear property,

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‡ Formerly with Computer Developments Ltd., Kenton, Middlesex; now of Racal Engineering Ltd., Bracknell, Berkshire.

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World Radio History

but they could be symmetrical in the positive and negative quadrants as in the characteristic shown in Fig. 1.

In the final analysis the preferred system of design will be a compromise between performance (speed of operation, etc.), reliability and cost of the whole system. Therefore, any assessment of a new logical element must take into account the driver circuits and subsequent amplifier circuits and, as will become apparent, the impedance levels are particularly important. In this paper an attempt is made to compare standard diode logic circuits with possible new circuits using bi-directional non-linear elements. In the main, the comparison is restricted to direct coupled (d.c.) logic, the signals being represented by steady state voltages on two levels, one representing a binary 1 and the other a binary 0. After a brief outline of the general approach to the design of complex switching systems, the basic diode circuits to perform AND and OR operations are reviewed in conjunction with typical transistor driver circuits. This enables one to set up a basis of comparison for circuits employing an alternative type of passive device. In the second part of the paper, circuits using one of these alternatives, a bi-directional non-linear device, are described and there follows an assessment of the (modified) requirements of driver and subsequent amplifier stages. Some remarks on reliability, possible cost, etc. are included. In the third part, other types of bi-directional devices are considered and some suggestions are made for novel logical circuits which make use of them.

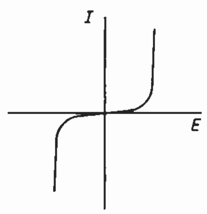

Fig. 1. Symmetrical characteristic of a bi-directional switching element.

1.1. *Logical Circuit Design—General Approach*

A network using only the three basic operations AND, OR and INHIBIT can produce any desired logical function of a combination of input signals. The amount of equipment required might well be reduced if other operations such as "EQUIVALENT TO" or "ANY 2 OUT OF 3" were made available. However, in very complex systems, it is found to be more practical and economic to design a few universal "building blocks." These follow a few simple interconnection rules so that the logical design may be effected quite separately from the electronic circuit design, thereby considerably reducing the difficulties of the overall design. The "building blocks" may be arranged to use a minimum of different types of devices and printed circuitry and other mass production techniques may be employed to reduce costs; moreover, the number of replacement packages required for servicing is much reduced. For the same reasons it is uneconomic to employ too many stages of logic in cascade without buffer amplifying stages. For instance, in 4-stage logic the impedance level at the fourth stage would be widely different from that at the first and the choice of value for a particular load resistor would depend in a complicated manner on the values of other load resistors. A single change in the logic would then lead to extensive re-design of the electronic circuitry.

No additional logical function is achieved by the sequence AND-AND or OR-OR, consequently the general pattern in any branch of the network is—AND-OR-AND-OR—with INVERTER stages inserted at appropriate points. Amplifiers must be included in the chain at suitable intervals to restore the original amplitude but there is no automatic choice for their position. In two-stage logic the sequence could be AND-OR-AMPLIFIER—or alternatively OR-AND-AMPLIFIER. The choice will depend to a great extent on the form of the signal pulses on which logic is to be performed. These will be in the form of binary signals supplied by flip-flop circuits, registers, counters, etc. In the case of flip-flops and most register circuits, two output pulses are available which are logically complementary; let these be A and $\overline{A}$ where $\overline{A}$ = NOT A. According to Boolean algebra

$$\overline{A.B.C.D.} = \overline{A}+\overline{B}+\overline{C}+\overline{D}+\ldots$$

This means that from a logical point of view there is no difference between starting the

World Radio History

logical net with AND gates fed by inputs A, B, C. etc., and starting the net with OR gates fed by $\overline{A}$, $\overline{B}$, $\overline{C}$, etc. A following section on transistor driver stages shows that the sequence AND-OR-AMPLIFIER is most economically achieved with negative logic, i.e. when the more negative value of the input potential is the unique level, representing the signal present condition. Correspondingly, the sequence OR-AND-AMPLIFIER is best achieved with positive logic where the positive value of the input pulse is the unique level. Where "barred" inputs are automatically available the choice of logic sequence is an open one and depends on secondary considerations.

If positive pulses only are available at the inputs to the net, one should endeavour to arrange the logic to go OR-AND-AMPLIFIER. On the other hand, the minimal system from pure logic point of view might call for more AND gates than OR gates, in which case the above arrangement would call for a liberal number of INHIBIT and/or INVERT stages which might more than cancel the saving on the OR-AND gate pairs. In these circumstances one would revert to the AND-OR-AMPLIFIER sequence which requires less amplifier stages (one per OR gate). If negative pulses only are initially available the reverse argument applies.

Once the cost of the various "building blocks" has been assessed, the costs of the several alternative arrangements may be compared and the minimal design, i.e. the design with a given performance at lowest cost, may be chosen. As the cost of semi-conductor diodes is much less than that of transistors, one endeavours to drive as many diode gates as possible from a given transistor driver and also to have as many depths or stages of logic as possible between amplifier stages; the number is not likely to exceed 2 in fast logic such as is required in digital computer systems.

2. Diode Gates

2.1. *Positive logic* AND *gate*

The more positive value of input potential (E_1 of the pulse shown in Fig. 2(a)) represents the unique level and this is taken to be a "1" say. The lower potential E_0 is the non-active condition and represents a "0" say. A conventional diode AND gate takes the form of Fig. 2(b) where $E_2 > E_1$. In the figure a 3-gate is shown but more inputs could be used. If any of the inputs A, B, C is at the lower potential E_0, the corresponding diode d_A, d_B, d_C will be in the conducting state, thereby clamping the output terminal to a potential close to, but slightly greater than E_0. Only when all inputs A AND B AND C are at the higher potential E_1 will the output rise to this value, and so indicate a "1." Note that any input source in the active condition is isolated from all other input sources because its diode is in the non-conducting state —this is a valuable feature of diode gates.

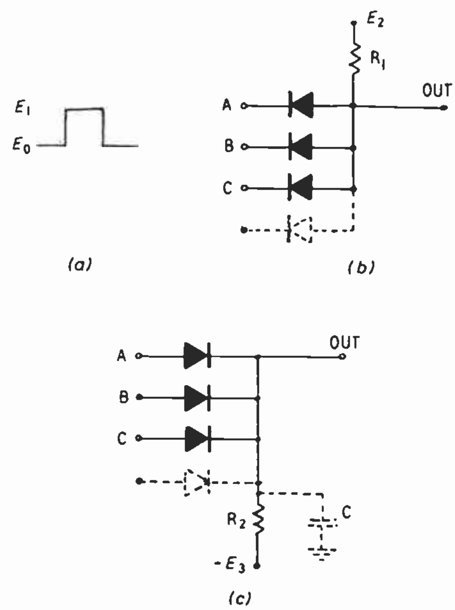

Fig. 2. Positive logic gates: (a) switching levels, (b) AND gate, (c) OR gate.

There will be a standing current in R_1 of $i_1 = (E_2 - E_0)/R_1$ when the gate is closed and in the worst case, i.e. when all but one of the inputs is at E_1, this current i_1 must flow into the one source still at potential E_0. Hence, all input sources A, B, C . . . must be capable of absorbing current i_1 without significant change of terminal voltage—this requires sources of low output impedance. Moreover, it is in the interests of economy that these sources are able to supply a number (n) of AND gates in parallel to avoid need for buffer amplifiers, and the magnitude of the current which has to be absorbed is thereby raised to ni_1.

World Radio History

There will be an optimum design maximum for n, above which the need for extremely low impedance makes the cost of driver stages prohibitively high.

2.2. *Positive logic* OR *gate*

The circuit is that of Fig. 2(c) for a 3-input gate; in the absence of input signals the diodes are all conducting since the load resistor is returned to a negative supply voltage $-E_3$. Consequently, if any one or more of inputs A, B, C rises to E_1, the output also rises to the higher potential E_1. Thus the output is a "1" if A OR B OR C is a "1"; this corresponds to the inclusive OR.

The maximum value for load resistance R_2 will be determined by the required switching speed and the magnitude of the shunt capacitance C; this capacitance is made up of the input capacitance of the following stage, the capacitance contributed by other diodes in the gate and the stray capacitance of the wiring. On a "1," the output can rise quickly to E_1, since C is charged up rapidly from the input source(s), which can be designed to have a low output impedance. If all the input voltages fall to E_0 all the diodes become cut off and C will have to discharge towards potential—E_3 via load R_2. When the output potential has fallen to E_0, the diodes conduct again. In practice both E_1 and E_0 might be small, e.g. 0V and -6V respectively, and $-E_3$ might be relatively large, e.g. -24V: the current through R_2 is then almost constant at E_3/R_2 and the rate of fall of the output voltage will be approximately constant at E_3/CR_2 volts/sec.

A convenient value is chosen for E_3 and then R_2 is determined. If the OR gate is to be followed by an AND gate, other considerations apply and, as will be explained later, R_2 will have to have a much lower value.

2.3. *Decoupling gates*

Figure 3 shows a transistor driver stage feeding a number of AND gates and OR gates in parallel (positive logic). When the transistor is off the collector potential will vary according to the number of AND and OR gates connected to it; each different combination would require a special variant on the load resistor values, R_1 and R_2, to preserve the same unique level of the output pulse. This is undesirable, leading to a multiplicity of slightly different units: the use of a standard block will certainly be preferred for complex systems and in order to accomplish this aim, one is prepared to include "redundant gates." If the general logic followed the sequence AND-OR-AMPLIFIER, each parallel OR gate would be decoupled by a "redundant AND" as shown in Fig. 4.

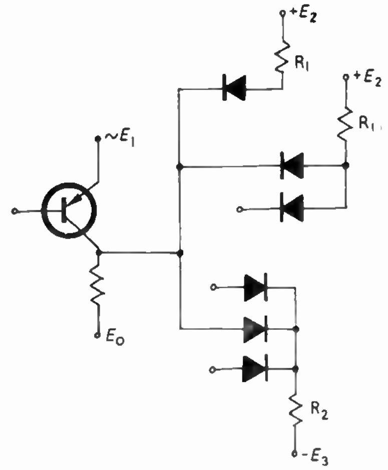

Fig. 3. Transistor driver connected to AND and OR gates in parallel.

For the converse sequence a redundant OR would be used before each parallel AND gate. Fortunately, this requirement to drive an OR as well as an AND gate in parallel from the same source does not occur frequently in practice and

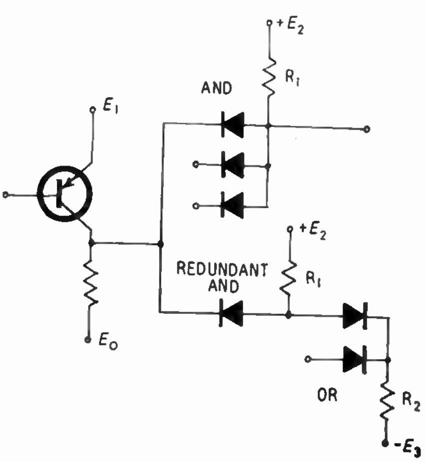

Fig. 4. Parallel connected OR gate decoupled by redundant AND.

World Radio History

the number of occasions on which it does so can be kept to a minimum by variation of the logical design. For instance, in a digital computer the number of such occasions can be kept below 5 per cent.

2.4. The AND-OR *sequence*

In a complex switching arrangement, the logical functions to be performed will require combinations of AND and OR circuits and it will be necessary to use the output of the one as the input of the other. Consider the "two level" system of an AND circuit followed by a single OR circuit (Fig. 5). Even if perfect diodes were available, and if the load connected to the output terminal were to draw no current from the circuit, the potential at point P cannot, in any circumstances, rise above $-E_3 + \dfrac{(E_2+E_3)R_2}{R_1+R_2}$

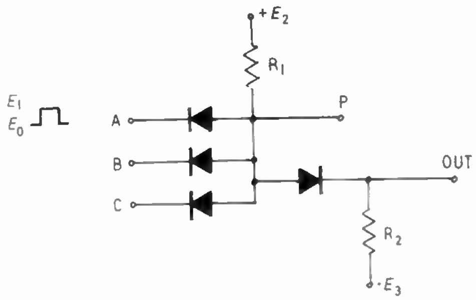

Fig. 5. Circuit for AND-OR sequence.

If this potential is less than E_1 the circuit cannot give its full output. Hence, we must arrange that

$$E_1 \leqslant -E_3 + \frac{(E_2+E_3)R_2}{R_1+R_2}$$

$$\text{i.e. } R_1 \leqslant R_2 \frac{E_2-E_1}{E_3+E_1} \qquad \ldots\ldots\ldots(1)$$

Now, as we have seen, R_2 is limited by capacitance conditions; therefore R_1 is also limited in value.

If the AND gate were required to supply an input to each of n similar OR gates, the equivalent value of R_2 would be R_2/n and R_1 would have to be reduced further by factor n. To avoid multiplicity of design, it is common practice to restrict the number of OR gates to one only. Any additional OR gates are preceded by redundant AND gates thus permitting the design of a single "package" in the sequence, AND-OR-AMPLIFIER.

2.5. The OR-AND *sequence*

This is the two-level system of one of the inputs of the AND circuit of Fig. 2(b) connected directly on to the output terminal of the OR circuit of Fig. 2(c). In this arrangement, the relative values of R_2, R_1 and $-E_3$, $+E_2$, must be such that the potential at the output point may fall to the lower value E_0 of the input signal; the following expression applies

$$R_2 \leqslant R_1 \frac{E_3+E_0}{E_2-E_0} \qquad \ldots\ldots\ldots(2)$$

The argument now follows the same pattern as that for the AND-OR sequence; the response time will be limited by stray capacitance across R_1 and for a given E_2 this sets a limit on R_1 which further imposes a limit on R_2 according to expression (2) above; hence, the maximum number of OR gates which may be driven in parallel from a common source is decided.

Once again it is good design, on economic grounds, to restrict the number of AND gates fed from an OR gate to one only.

2.6. Transistor driver stage

For the purpose of this paper it can be assumed that the signal source will be a *pnp* transistor operating in the common emitter connection, the output appearing at the collector. The transistor could be the output stage of a two-stage buffer amplifier or one of a pair in a flip-flop circuit. The transistor should be driven over its whole characteristic so that the output swings between the collector supply voltage (E_0) and the emitter potential ($\sim E_1$): this gives a predetermined output pulse amplitude. However, it must not be driven into the saturated condition, i.e. the condition in which the base-collector junction becomes forward-biased. If this occurs an excessive number of minority carriers are introduced into the base-collector region and this junction cannot become reverse-biased until these are removed. This hole storage effect makes it impossible to switch off the transistor immediately on raising the base potential, a true delay of as much as several microseconds may occur before this is accomplished and this will limit the maximum p.r.f. of switching. Various methods have been devised to

World Radio History

eliminate this effect by an extra clamping diode; one such method is illustrated in Fig. 6 (which depicts a two-stage amplifier providing barred outputs, described later in more detail). The clamping diode D_s is chosen to conduct at a lower voltage drop than the collector-base diode and is also provided with a small forward bias because of the volts-drop across R_b. Once D_s conducts, degenerative feedback to the base occurs and the current which would otherwise flow in the transistor from collector to base is by-passed via D_s. The value of R_b must be kept to a minimum so that the collector voltage is clamped within a few tenths of a volt of the saturation value, in which case the internal dissipation is only slightly greater than it would be if saturation were allowed to occur.

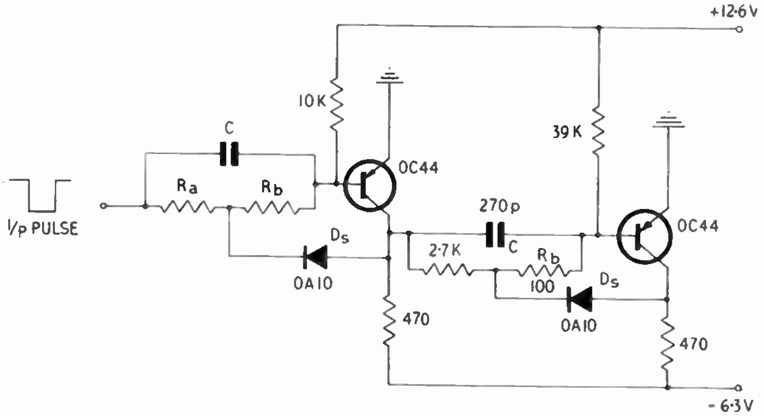

Fig. 6. Two-stage amplifier providing two complementary outputs.

2.7. *Transistor driving an* OR *gate (positive logic)*

The maximum current is required to flow in the OR gate for positive change of voltage and this conveniently coincides with the transistor "on." However, when all inputs to an OR gate are "0's," current still flows into the gate circuit because load R_2 is returned to a negative potential $-E_3$. If a transistor source in the "off" condition is connected to an OR gate having n_1 inputs for which all the other inputs are also at "0," then the current i_2 which flows in R_2 must flow via the collector resistors, in parallel, of all the n_1 transistor stages feeding that gate. The current through the collector resistor of each driver stage is therefore i_2/n_1 and, in the very worst case, $n_1 = 1$ (a redundant OR). If the transistor were to feed n_2 OR gates in parallel, the sum of all these currents would cause a significant change of voltage level at the collector which is likely to interfere with the working of other gates for which this driver is a common source; the output signal from the OR gates will also be increased in amplitude. To make the effect small it would be necessary for the collector resistors to have a value $R_L \ll R_2\, n_1/n_2$. Consequently, n_2 may have to be limited in number and R_L made small. However, R_L cannot be reduced excessively compared with R_2, otherwise the current out of the transistor, when switched on, will flow mainly in this load resistor and not into the gate—this would require a "larger" transistor which would almost certainly turn out to be either slower in operation or more expensive. These disadvantages may be eliminated by the use of a single clamping diode d_p per OR gate as shown in Fig. 7. This diode is returned to a supply voltage some 2 volts or so less negative than the collector supply voltage. When all the inputs A, B . . . to the gate are at -8V diode d_p conducts and clamps the potential at P to -6V. The output voltage swing is therefore held constant: also input diodes d_A, d_B . . . are biased off, consequently no current is required to flow in the collector resistors dur-

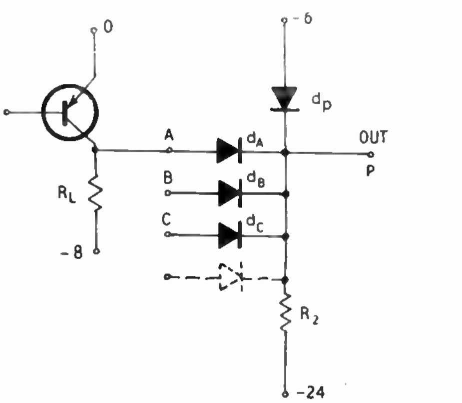

Fig. 7. Positive logic OR gate with single clamping diode.

ing their "off" time; the value of these resistors may therefore be quite high, restricted only by capacitance effects. The reverse currents of the gate diodes are usually small enough to be of secondary importance. The inclusion of d_p thus enables the transistor drivers to be used more efficiently, standardizes the output and enables more OR gates to be fed from a common source.

If, in the circuit of Fig. 7, the transistor is the only one of the input drivers in the "on" condition, it must supply current, $24V/R_2$ into this gate, plus current $8V/R_L$ into the collector resistor R_L. Now the "standing current" in the gate is 18 volts/R_2 and this is a measure of the current available to hold closed a following AND gate and amplifier, hence one might define a "current loss factor" for the ratio of transistor current to standing gate current. In the present example:

current loss factor=

$$\frac{24/R_2+8/R_L}{18/R_2} \cong 1{\cdot}3+0{\cdot}5\frac{R_2}{R_L}$$

The factor 1·3 could be made to tend towards unity by increasing the gate supply voltage from -24V to more and more negative values. The magnitude of the term $0{\cdot}5\,R_2/R_L$ will in the event depend on the rise time specified, since this governs the values of both R_2 and R_L. For slow speed operations it will be small: an OR-AND-AMPLIFIER sequence to operate at 1Mc/s p.r.f. would have an overall loss factor in the region of 3·0.

In order to drive n_2 OR gates in parallel the transistor must be capable of supplying a maximum current $n_2 \cdot \frac{24V}{R_2} + \frac{8V}{R_L}$.

2.8. *Transistor driving an* AND *gate (positive logic)*

Consider next the case of an AND gate driven from a *pnp* transistor driver circuit (Fig. 8)—here the problem is worse. When the driver is switched on, (a "1") the rise of collector voltage turns off the gate diode, thus the driver is in a condition to supply current, but is not called upon to do so. Yet, when all the other inputs to the gate are "1's," the standing gate current must be capable of being "absorbed" into the one driver at "0" in order to keep the gate closed. As this driver is in the "off" condition the current must leak away through the collector resistor. The result is an increase of potential at the collector, reducing the signal swing available from this source; this is likely to interfere with the working of other gates for which this one driver is a common source. As explained above, the collector resistance cannot be reduced excessively to prevent this effect. This disadvantage can be eliminated by the use of an additional diode per source (d_x of Fig. 8 shown dotted). This diode is returned to a supply E_0, and the collector load R_L is returned to a more negative potential E_4. Potential E_4 and load R_L are so chosen that in the absence of d_x and with the transistor off, the collector potential would fall below E_0, hence d_x conducts and the output signal swing is clamped near to E_0. To ensure that the diode will always conduct when the transistor is "off," the current through R_L which is $\frac{E_0-E_4}{R_L}$ must exceed the standing gate current $\frac{E_2-E_0}{R_1}$ by an adequate margin to allow for tolerances (say 1·5 times). When the driver is required to operate n_3 such AND gates, the current through R_L must be capable of rising to $n_3.\frac{(E_2-E_0)}{R_1}$ to cater for the worst condition, i.e. when the one driver source is called upon to hold closed all n_3 gates. This will inevitably lead to a design maximum for n_3 otherwise the cost of the driver circuit becomes uneconomic.

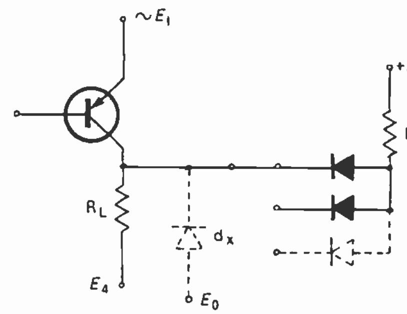

Fig. 8. Positive logic AND gate driven by *pnp* transistor.

When the transistor is switched on the collector voltage must rise to E_1. To do so it must supply current $\frac{E_1-E_4}{R_L}$ into R_L. But, from

World Radio History

above $\frac{E_0 - E_4}{R_L} \geqslant n_3 . \frac{E_2 - E_0}{R_1}$, hence the current from the transistor must exceed or be equal to $n_3 . \frac{E_1 - E_4}{E_0 - E_4}$ times the standing gate current. The "current loss factor" is therefore $n_3 . \frac{E_1 - E_4}{E_0 - E_4}$. With typical values, $E_0 = -6\text{V}$, $E_1 = 0$, $E_4 = -18\text{V}$, $n_3 = 5$, the current loss factor is 7·5. This pinpoints the advantage of making E_4 much more negative than E_0; if E_4 had been -9V, the current loss factor would have been 15.

2.9. *Negative logic gates*

The more negative value of input potential (E_0 of the pulse in Fig. 9(a) represents the unique level and to be consistent with previous sections this is taken to be a "1." The conventional diode AND gate is that of Fig. 9(b), where $-E_3$ is more negative than E_0. Note that this circuit is similar to the OR circuit for positive logic. Only when all the inputs fall to E_0 can the output fall to this value. If any of the inputs A, B, C . . . ., remains at potential E_1, the corresponding diode will remain conducting. The current through R_1 required to maintain the output at E_1 must therefore be supplied by the one source when that source potential is at the upper value E_1.

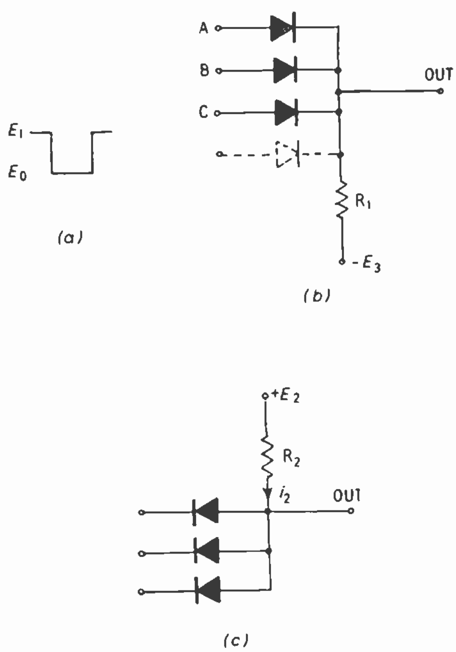

Fig. 9. Negative logic gates: (a) switching levels, (b) AND gate, (c) OR gate.

The OR gate for negative logic is the same circuit as for the positive logic AND gate; the action is basically similar to that of the positive logic OR gate.

2.10. *Transistor driver for negative logic* AND *gate*

Just as for positive logic, there will be restrictions on the value of resistors R_1 and R_2 but there is one significant difference: the signal required to hold the AND gate closed is a positive pulse, and this occurs when the transistor is "on"—it is therefore in the right condition to supply the current necessary to keep the gate "closed." However, if a transistor source in the "off" condition is connected to an AND gate for which all the other inputs are also at "1," then the current i_1 which flows in R_1 must flow via all the transistor collector resistors in parallel. If the one driver stage were to feed several AND gates in parallel it would require that the current in that collector resistor would have to rise to a value of order i_1, possibly greater, under the worst conditions (all gates open). This state of affairs corresponds exactly to that pertaining to the positive logic OR gate for all inputs at "0." The same remedy, i.e. the inclusion of a single clamping diode d_p per gate, as in Fig. 7, surmounts the difficulty.

This requirement of one extra diode per gate compares favourably with that of one extra diode per input source necessary for the positive logic AND gate; current loss factor is similar to that of the positive logic OR gate and is therefore several times better than for the positive logic AND gate. Furthermore, the current drain into the clamping voltage supply is only equal in magnitude to the standing current in the gate resistor whereas for positive logic AND gates, the current drain at the clamping point on each transistor collector must be n_3 times the standing gate current where n_3 is the maximum number of gates to be driven by any one transistor source. The reduction of total current consumption from the supply voltages is an economic factor not to be ignored. This high current drain into the clamp diode at the collector becomes a greater embarrassment at high p.r.f.

World Radio History

Semi-conductor diodes are subject to carrier storage but this can be made negligible for the diodes used in the gates themselves by keeping currents sufficiently low (e.g. < 5mA for 1 Mc/s p.r.f.). However, the larger forward current requirement of the collector clamp diode can lead to pronounced carrier storage and a special (more expensive) low storage diode such as the Mullard OA10 must be employed in this position at high p.r.f. Here again, the negative logic AND gate has the advantage over its positive logic counterpart.

2.11. *Transistor driver for negative logic* OR *gate*

As might be expected, the driving conditions here are similar to those for the positive logic AND gate. The standing current i_2 in the OR gate load resistor must drain away down one source collector load resistor when that source is the only one to supply a "1." Where a source is feeding n_2 such OR gates, it must be capable of absorbing current n_2i_2. Once again, the conditions are inappropriate in that the source current is required to be a maximum when the transistor is off. To prevent the need for very low transistor collector resistors R_L with aforementioned disadvantages, it will be necessary to use one clamping diode per driver source. This diode conducts when the collector potential falls to E_0, the transistor load R_L being returned to a more negative supply voltage, as in the circuit of Fig. 8 previously described.

2.12. *Provisional conclusions*

On the basis of the foregoing discussion one would conclude:

(a) With positive logic it is more economical to drive OR gates from *pnp* transistor sources: the logical design should therefore follow the sequence OR-AND-AMPLIFIER. At this stage it is advisable to follow the AND gate by an amplifier to restore the signal pulse to its original shape and amplitude.

(b) With negative logic the more economical arrangement is to drive AND gates from *pnp* transistor sources: the logic design sequence would then be AND-OR-AMPLIFIER.

These conclusions are not necessarily valid in all practical conditions because there are yet other factors which may have to be taken into account, depending upon the application for which the equipment is designed. This is likely to influence amplifier design and the following paragraph is a brief outline of two different approaches to the problem.

As a result of the general branching out of the logic network the signal becomes attenuated as it proceeds through each depth of logic; as we have seen, impedance levels must be made higher, and consequently the effects of capacitance are increased as is also the possibility of "cross-talk." An amplifier is therefore included in the chain after the appropriate number of cascaded stages. It is possible to use only one stage of amplification, for instance a common-emitter transistor stage; this leads to a reversal of signal polarity and the next logic cycle would have to be reversed. This arrangement is likely to be inconvenient as it requires the provision of a new range of standard blocks; also numerous single stage inverters may have to be provided at various points in the network to bring the incoming signal to the appropriate polarity. It is generally preferred to use a minimum of standard blocks: one arrangement is to use an amplifier consisting of two stages of inversion so that the output pulse is of the same polarity as the input and is also restored to the standard amplitude. The second (output) stage needs to be a high current stage in order to drive the following gates but the first stage could be a low current stage; d.c. coupling between the two stages makes it possible to operate the second stage at the correct d.c. level, restoring the output voltages to the standard values. This combination might be preferred where positive logic is used and where the mark-space ratio of the signal pulse is small; in these circumstances there could be some saving of power. The sort of application where this factor is important enough to influence design is in electronic telephone exchanges where the mark-space ratio can be very small. In other applications it may be preferred to design both stages of the amplifier to be capable of supplying sufficient current to operate the design maximum number of gates. It is then possible to take off "barred" signal pulses from the first stage whenever this is required; there is then no need for isolated inverter stages and the logic can follow a different pattern utilizing "barred" values wherever this leads to overall economy of gating components. Such a system is frequently

World Radio History

used in digital computer networks. A typical two-stage amplifier of this type for operation at 1 Mc/s p.r.f. is shown in Fig. 6. Resistor R_a must be chosen to provide the correct range of base potential when the input terminal is connected to the preceding gate. Capacitors C serve an important role in speeding up the switching time. Both rise and fall time are improved by using the transistors at low effective current gain (this implies large initial changes of base current and this is assisted by capacitors C) and by using load resistance $R_L \ll \frac{1}{2\pi f\alpha C_c}$ where $f\alpha$ is the cut-off frequency and C_c is the collector capacitance†.

3. Non-linear Resistance Gates

3.1. *Constant voltage* AND *gates*

An AND gate has been devised which makes use of non-linear resistance elements which have characteristics of the type illustrated by Fig. 1. The increase of voltage required to produce a given increase of current gets smaller and smaller at higher current levels; hence, for the purpose of reference, the elements will be called "constant voltage (CV) elements" and gates using such elements will be called CV gates. A typical characteristic might follow the power law

$$I = kE^m$$

and, as will be shown, the value of m should be high, for instance not less than 4·0 over the useful part. The basic 3-input AND gate is shown in Fig. 10, in which the elements a, b, c,

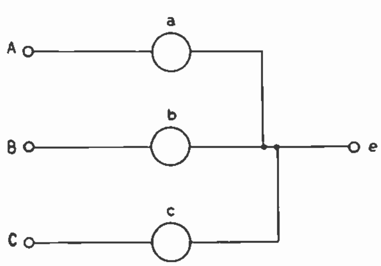

Fig. 10. Basic 3 input CV AND gate.

† J. L. Moss, "Large-signal transient response of junction transistors," *Proc. Inst. Radio Engrs,* **42**, pp. 1773-1784, December 1954.

J. W. Easley, "The effect of collector capacity on the transient response of junction transistors," *Trans. Inst. Radio Engrs (Electron Devices),* **ED-4**, pp. 6-14 January 1957.

are similar. The gate is assumed to operate from similar positive pulses, of amplitude E. Because of the bi-directional properties of the elements, it is immaterial from their point of view whether positive or negative logic is used. The input pulse sources must have an impedance low compared with that of the elements a, b, c, when voltage E is applied across them: in the first instance, the source impedances are taken to be negligibly small.

If voltage E is applied to one input only, say A, the current through A divides equally between elements b and c and flows into the low impedance sources at B and C. Thus

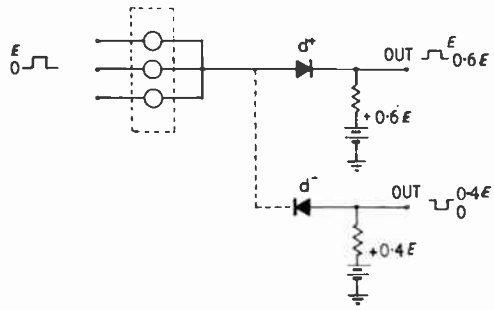

Fig. 11. Output circuit for CV AND gate.

the current through element a has to be twice that through elements b and c but, because of the non-linear nature of the resistance, it is only necessary for the voltage across a to exceed that across elements b and c by a small margin to provide this double value current. Therefore the (unloaded) output voltage e is fairly close to 0·5 E and for a fourth power law characteristic the particular value of e would be 0·46E. By symmetry, it follows that, if two input voltages are applied, say at A and B, the output voltage e will have the value (E − 0·46E) i.e. 0·54E. When E is applied to all three elements, the unloaded output voltage is E, and for all inputs at zero, the output is, of course, zero.

If now the output is connected to a high impedance load R via a diode (d^+ of Fig. 11) biased to say 0·6E, there is a positive output voltage with a maximum value 0·4E only when all three inputs are applied, corresponding to a 3-AND gate. The gate may be modified to give *double selection* properties by means of a second diode d–, biased to $0{\cdot}4E$. This provides a second output voltage which falls from $0{\cdot}4E$ to zero when all the inputs fall to zero. This is equivalent to a second 3-AND gate preceded by 3 INVERTER stages.

At first sight, the single output circuit appears to use three non-linear elements plus one diode as compared with three diodes in the conventional diode gate. In actual fact, the three non-linear elements can be combined into one, with three areas of electrode printed, sprayed, or otherwise deposited on one face of a thin disc

Fig. 12. Configuration for composite three non-linear element disc.

and a common electrode on the reverse face (Fig. 12). Provided that the spacing between separate electrodes is some 4 to 5 times the thickness of the disc, the cross-currents between A, B, C are negligible because of the fourth-power relationship. The relative economies of non-linear resistance and diode gates are compared more closely in a later Section.

3.2. AND *gates with more than 3 inputs*

The simplest method of deriving the output voltage e is by the graphical method illustrated by Fig. 13. The characteristic (1) of any one element is drawn on a log-log scale assuming a fourth power relation. From this is deduced a

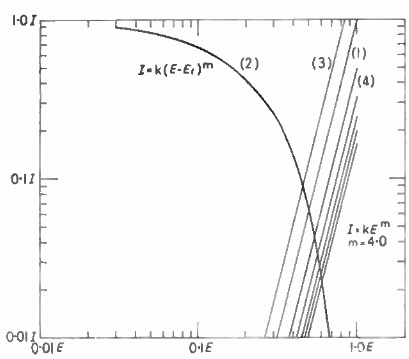

Fig. 13. Graphical derivation of output voltage of non-linear AND gate.

load line (2) for any element by plotting the respective current values I_1 of characteristic (1) against voltage $E_2 = E - E_1$ where E is the magnitude of the input pulse. If input E is applied to element A only, the combined characteristic (3) of elements B and C is parallel to (1) but has twice the current values for each applied voltage. The intersection of (2) and (3) gives the output voltage e at the junction—this is $0{\cdot}46E$. To find e for input E applied to two elements one could draw a new load line similar to (2) but with doubled current values; it is much simpler to draw a new linear characteristic (4), parallel to (1), with halved current values. The intersection of (2) and (4) gives the value $e = 0{\cdot}54E$ for two applied pulses.

The method is extended as follows:—by plotting further characteristics parallel to (1) with current level equal to one-third that of (1), one deduces that, for a 4-gate with three inputs

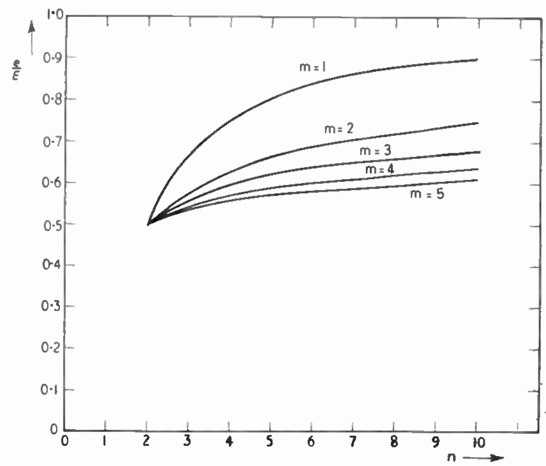

Fig. 14. Output of CV AND gate for signals applied to all but one of n inputs, plotted for different power indices.

applied, output $e = 0{\cdot}57E$. Now, in any n-gate, voltage e approaches most nearly the full output E when $(n-1)$ inputs are applied. This case is the one requiring maximum discrimination in the diode limiter. Results for a 5-gate, 6-gate, etc., are derived and plotted in Fig. 14. In similar manner graphical solutions may be obtained for $(n-1)$ inputs applied to an n-gate for third, fifth, etc. power law materials. In Fig. 14 these results are also plotted for values of n from 2 to 10, and for power law indices from 1 to 5. A gate with a large number of

World Radio History

inputs is possible if the discrimination ratio E/e is large enough: to give ample margin for tolerances of voltage a maximum bias of $-0{\cdot}65E$ might be agreed upon. In the ideal conditions assumed, this makes possible a 10-gate for fourth power material.

3.3. *Tolerances*

The essential reason for using materials with a high power law is to maintain the output voltage e as near as possible to $0{\cdot}5E$ when all but one of the input voltages are applied. It follows that a wide tolerance of device resistance (for a given voltage) can be accepted. For instance, suppose that one of the elements had a resistance level 50 per cent. of the standard. This will have the maximum effect (and equivalent to two elements in parallel) when this element is included amongst the $(n-1)$ which have inputs applied. Clearly this is equivalent to applying the same input to n input terminals of an $(n+1)$ gate in which all components are standard. Reference to Fig. 14 shows that there is only a small increase of e and the gate is still workable. If one of the resistances is 100 per cent. above standard, the output e will be a maximum when input voltages are applied to all elements with the exception of this one. The n gate then behaves similarly to one containing standard level components with inputs applied to $2(n-1)$ elements, i.e. equivalent to a $(2n-1)$ gate. Hence, a bias level set to be suitable for a $(2n-1)$ gate would cater for as wide a tolerance margin as -50 per cent. to $+100$ per cent. on any one element of an n-gate.

3.4. *Non-linear resistance material*

The most obvious material currently available is silicon carbide which is at present marketed as single elements under the trade names Morganohm, Metrosil, Atmite, etc. In this form the elements are used as surge limiters and the uniformity and stability of the parameters have not been of particular importance. More recently there has been interest in the materials as a means of producing, in conjunction with two linear resistors, a square law characteristic for analogue computers and in surge limiters for use with transistors. The stability is quite adequate for the present application.

The silicon carbide is porous and so it is necessary to avoid effects of humidity by impregnation. The resistance varies with temperature by $-0{\cdot}25$ per cent. to $-0{\cdot}5$ per cent. per °C at constant voltage in the range 20°C to 60°C: the effects of this variation are not unduly important since it does not alter values of e, but only alters the currents drawn from the pulse sources.

In order to use non-linear devices for gating in conjunction with transistor driver stages, a current in the range 1 to 10 mA at 6 volts would be indicated. It is desirable that the index of non-linearity should remain high down to a voltage of 2 volts or less. With material currently available in this country the index falls considerably at this voltage level unless a very thin disc is used. Reduction of thickness has the disadvantage of reducing mechanical strength (the material is very brittle) and thereby introducing manufacturing difficulty but, more important, it leads to increased capacitance. The characteristic of a typical disc of British manufacture having a thickness of 0·05 in and with sprayed brass electrodes of diameter ¼ in. is shown in Fig. 15(a). Notice

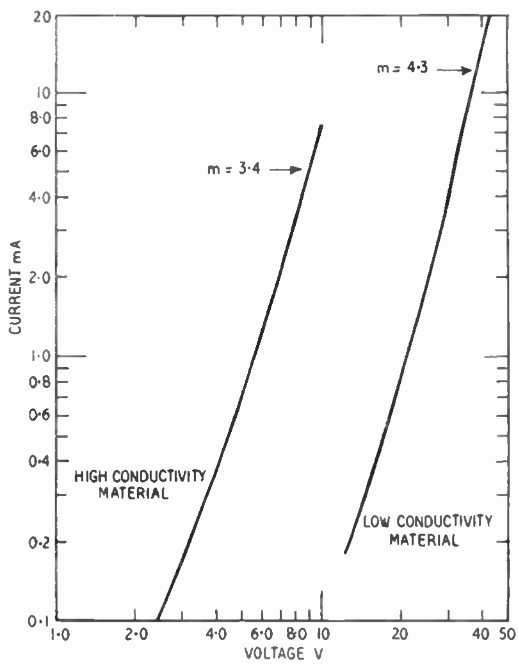

Fig. 15. (a) Characteristics of typical silicon carbide elements.

that this element requires a voltage pulse of 30-40 volts amplitude for satisfactory application in the AND circuit. It could be used with thermionic valve driver circuits but would not be successful with present day low voltage tran-

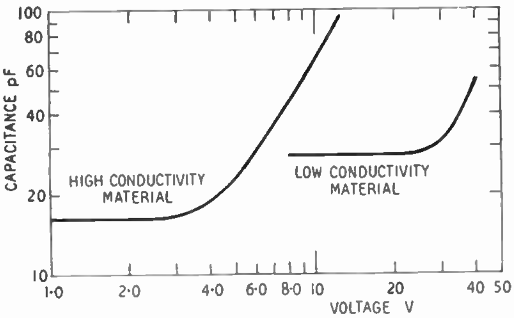

Fig. 15. (b) Capacitance/voltage variation of silicon carbide discs.

sistors. The non-linear index m has a value 4·3 in the higher voltage region. In the U.S.A. many millions of silicon carbide discs have been installed as amplitude equalizers in long-distance telephone equipment. For a comparable value of m, the material used in these discs has a higher conductivity for a given field strength than the British product. Thus Fig. 15(a) includes the characteristic of a disc of thickness 0·03 in., with electrodes of only 3/32 in. diameter. This element may be used with low voltage transistors and provides adequate current for gating purposes.

The greatest difficulty encountered with silicon carbide discs has been the lack of uniformity of properties over the surface of a disc. For both low and high conductivity materials, six similar area brass electrodes were sprayed on and currents were compared for a given voltage. It was quite common to find a ratio as high as four times between the highest and lowest currents. This is too high to allow practical use of the device but a more strict control of materials in production should reduce this ratio. An initial surge overload treatment given to each individual element of a composite disc might help in this respect.

A preliminary examination has been made of two possible alternatives. The first was compressed, powdered, 0·1Ωcm n-type silicon. The current density was more than adequate for a device operating at 10 V but the power law index m was in the region of 1·5 only. Oxidation of the powder in wet nitrogen at 1200°C led to a small improvement in m but its value was still less than 2·0. A more promising material was found to be cadmium sulphide powder, activated by gallium as a donor impurity. Compressed pellets of this material (0·8cm^2, 0·4mm thick) passed a current of 5mA at 6V. The non-linear index m exceeded 4·0, down to as low a voltage as 1·0V. Powders of lower conductivity yielded pellets showing values of m up to 7·0. Unfortunately, attempts to prepare pellets in a binder such as Araldite, nitrocellulose, etc., led to an immediate fall in m and also in the conductivity. Also reproducibility from batch to batch was poor, some samples showing an irreversible change after passing currents of 10mA/cm^2. There are apparently many problems to be solved before an alternative to silicon carbide is available.

3.5. *Effects of capacitance*

Consider as an example a 3-AND gate. Figure 16 includes the capacitances of the non-linear material, and these are first assumed to be of constant value C. The shape of the output

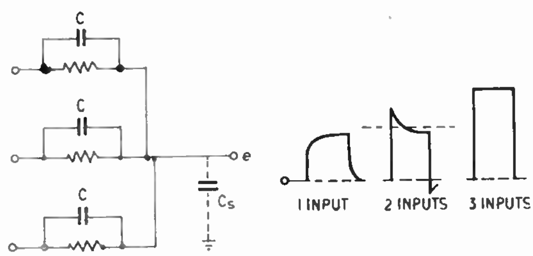

Fig. 16. Effects of capacitance for 3-AND gate.

pulses e for an applied perfect square pulse are illustrated. For two inputs the leading edge rises to 0·67E and a small narrow pulse (above the dotted line) could well get through the limiter circuit which follows. The effect would be much worse in a 4-gate with three inputs applied. One way to remove this "breakthrough" would be to introduce shunt capacitance C_s (possibly strays) but this might cause the transmitted pulse for all inputs applied to have a "slowed up" leading edge. The latter effect might be reduced by using higher conductivity material (if available).

World Radio History

In actual fact, it was found that in the silicon carbide discs the effective capacitance increases markedly at higher voltages. This variation of capacitance was measured on a Q-meter at 5 Mc/s for the lower conductivity silicon carbide samples and the graph of Fig. 15(b) is that for the same disc whose current variation is shown in characteristic (a). This non-linear behaviour of capacitance suggests that AND gates with four, and even six, inputs might be used, even when fast rise times are essential. In preliminary experiments, pulses at 1 Mc/s repetition rate with rise and fall times of 50 mμsec have been passed through a 3-AND gating circuit. The open circuit output pulse amplitude e for 40V input pulses applied to two inputs was in the region of $0{\cdot}55E$ and was almost perfectly flat-topped. With the diode and load circuit there was the usual difficulty that a small differentiated pulse got through to the load circuit via the diode capacitance; a current pulse of approximately 1 mA was available into the load. The absence of overshoot on the leading and trailing edges indicated that the variation of capacitance under pulse conditions was somewhat greater than that implied by the characteristic of Fig. 15(b). This is probably true since the representation of such a complex material by a parallel combination of a resistance and a capacitance is greatly oversimplified. An equivalent circuit such as that of Fig. 17 is a more likely representation in which both capacitance and resistance are closely interdependent. The capacitance across each internal "barrier" will be virtually eliminated once the field is high enough to cause conduction over the barrier and the overall capacitance is thereby increased. This conception would indicate that the *variation* in effective resistance and effective capacitance would be fairly closely correlated. Since the effective resistance is supposed to be concentrated almost entirely at these barriers then it will be much lower at higher frequencies.

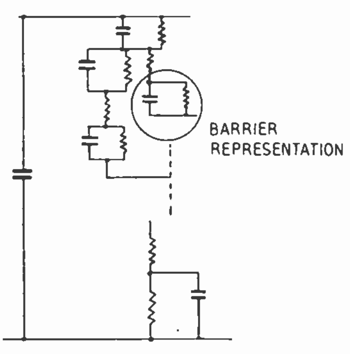

Fig. 17. Equivalent circuit of silicon carbide disc.

The capacitance variation of the device made from the higher conductivity material could not be measured by the simple resonance method because of the extra heavy damping of the tuned circuit. It was decided to test this specimen by a pulse method. A 100 kc/s square wave voltage of 0 - 12V peak/peak. rise time < 40 mμsec and no discernable overshoot was applied from a source of extra low internal impedance to the disc in series with a resistor R. The voltage waveform across R was examined by an oscilloscope having amplifiers capable of resolving a rise time of 12 mμsec. The waveform across R showed the expected overshoot; a capacitance C (including 10 pF input capacitance of the oscilloscope) was now placed across R such that the overshoot just disappeared, and the output voltage then measured. If the ratio of output voltage to input voltage is r, then the capacitance C of the disc is $Cr/(1-r)$. Using various values of input voltage and resistance R, values for C were obtained as a function of applied voltage.

The results are plotted in Fig. 15(b): the initial value of 16 pF is not excessive and provided that the diode load resistance R and capacitance C_s are small this type of disc gives satisfactory operation up to 1 Mc/s p.r.f. for a four-input gate. It is suggested that in practice a standard article might be produced from this material, for instance a four-gate. When less than this number of inputs is required, the appropriate number of electrodes are simply left disconnected—by this means a minimum price could be achieved.

3.6. OR *gates*

One of the main requirements of an OR gate is that it should transmit any one (or more) input pulses to the output terminal with the minimum disturbance to circuits connected to the other input terminals. The "disengagement" of the other input pulse sources by any one output pulse is readily achieved with diodes. Non-linear resistance devices may be used to direct current into low impedance load circuits (e.g., input impedance of a transistor amplifier) and at the same time have negligible effect on the other input pulse sources.

Consider the arrangement of Fig. 18. If a, b, c are fourth-power resistance elements and if R is such that e is small compared with the amplitude E of the one applied input pulse (say $e = 0{\cdot}1E$) then to a first approximation $i' = i/10^4$. This ratio gives adequate discrimination

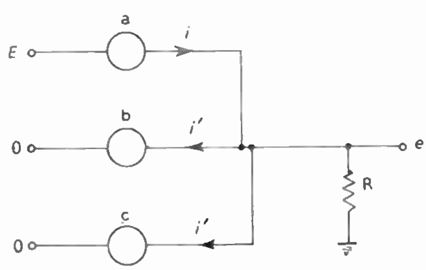

Fig. 18. Return currents in a 3-input non-linear OR gate.

between wanted and unwanted pulse currents. In a large and complex arrangement of OR gates the circuit of Fig. 18 can be expanded into a matrix arrangement, the necessary electrode areas being printed or otherwise deposited as rows and columns on opposite sides of a thin sheet of material. The spacing of the electrodes must be 4 or 5 times the thickness of the sheet in order to make "cross coupling" negligible. For the same reason all output points must be terminated in low impedance circuits: these could be the input impedances of following buffer amplifiers. A large mesh of this type might be used, with considerable economy, in a coder for computers or to give a fixed translation in a telephone exchange system.

3.7. *The* OR-AND-AMPLIFIER *sequence*

The OR gate using CV devices suffers from severe voltage attenuation, the output is therefore insufficient to drive a subsequent AND gate.

It would be possible to drive constant voltage AND gates from transistor OR gates but these would have to be high current transistor circuits to provide the necessary current. This would be an expensive arrangement because all the gates would be working at high current. The AND-OR-AMPLIFIER sequence is to be preferred.

3.8. *The* AND-OR-AMPLIFIER *sequence*

The constant voltage OR gate is not particularly suitable for direct connection after an AND gate because of the low output current and voltage available from it. Furthermore, the non-linear AND gate already includes a diode limiter, and in slow speed circuits this same diode may also fulfil the purpose of the diode in a standard OR gate. Figure 19 shows two non-linear AND gates whose outputs feed into a diode OR gate. The diodes are normally biased off, therefore the trailing edge of the output pulse may be slow due to capacitance C_s. To reduce this effect the load resistance R_2 would have to be small. This is possible without upsetting the AND gate action since the diode prevents current being drawn from the gate except under the appropriate input conditions. However, low load resistance means low output voltage, hence the following transistor amplifier would have to have a high voltage gain first stage followed by high current gain second stage to restore

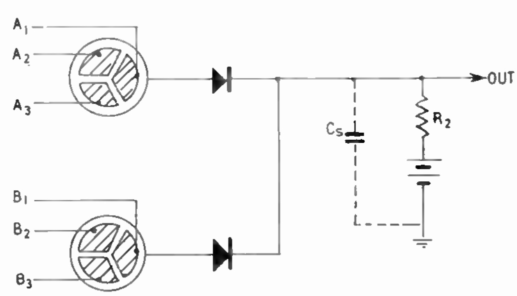

Fig. 19. Non-linear AND gates feeding diode OR gate.

polarity, to correct the d.c. level and to provide the necessary output current. A better arrangement would be to have (biased) transistor OR gates followed by single-stage high-current amplifiers. The replacement of the diode by a transistor would increase the cost but the performance would be improved. The capacitance of the silicon carbide discs is here an advantage in that it assists in the provision of a large rapid change of base current when the AND gate opens.

It is worth noting that with either diode or transistor OR gates, the non-linear AND gate can feed into more than one OR gate without change

World Radio History

of design parameters. This removes a restriction on the logic compared with the parallel situation for diode gates and advantage should be taken of this wherever possible.

3.9. *Transistor driver for constant voltage* AND *gate*

As a result of the bidirectional properties of the elements, the maximum current required to be drawn from any transistor source when it is the only one "on" is exactly equal to that required to be absorbed by the source when it is the only one not "on." The driver circuits must therefore be of low impedance both when "on" and when "off," hence a diode clamp must be provided at the collector of each transistor as in the circuits of Figs. 8 and 20. There is no relative advantage between positive logic and negative logic: the following argument is developed for positive logic only but the deductions apply equally well to negative logic. For simplicity, assume that pulses with voltage levels 0V (the unique level) and -10V are to be applied to the AND gate, that the elements have power law characteristic $i = kE^4$ and that no gate has more than six inputs. If this is so, Fig. 17 shows that the voltage e at the junction of the elements varies between the limits of -6V (for 1 input) and -4V (for 5 inputs). The maximum current through any one element is therefore $k(6)^4$ and the minimum value is $k(4)^4$. Each source must, in the worst possible circumstances, be able to absorb current $n_3k(6)^4$ where n_3 is maximum number of AND gates supplied in parallel. Let the diode clamp circuit be that of Fig. 20. For this circuit to absorb the required current

$$\frac{20}{R_L} \geqslant n_3k(6)^4$$

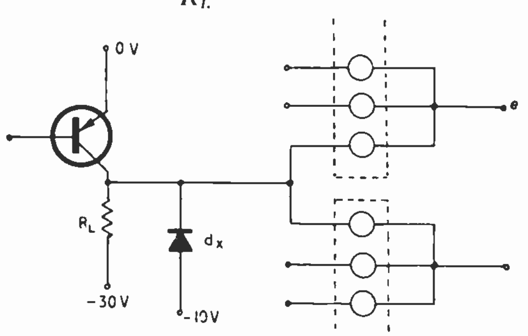

Fig. 20. Transistor driver circuit with diode clamp.

When the transistor is switched "on" it must first provide enough collector current to raise the collector voltage to 0V (approx.), therefore the current into the load R_L will be $30/R_L$. From the above, this is $1{\cdot}5\ n_3k(6)^4$. The source must also provide an additional maximum current $n_3k(6)^4$ into the gate to cater for the worst case i.e. when all the other inputs to all the n_3 gates supplied are -10V. The total required maximum current is therefore $2{\cdot}5n_3k(6)^4$. When all inputs to a gate are at "1," the diode limiter conducts; if we assume a diode bias voltage $-3{\cdot}5$V and that the load circuit is low impedance, then the voltage across each element of the gate is $3{\cdot}5$V. The output current is therefore $n_3k(3{\cdot}5)^4$, and the current loss factor is

$$\frac{2{\cdot}5n_3k(6)^4}{n_1k(3{\cdot}5)^4} \text{ which is approx. } \frac{22n_3}{n_1} \quad . \text{ If } n_3 = 5$$

as for the diode gates and if n_1 is put at the minimum value 2 for a 2-input gate, then the current loss factor is 55. The following amplifier would have to have a gain in excess of this figure. This current loss factor appears large compared with the factor $7{\cdot}5$ calculated for the positive logic diode AND gate but it must be remembered that the latter has a further loss in the following OR gate. Thus the standing current in R_2 of the AND-OR sequence of Fig. 5 would not exceed about one-third of the standing current in R_1 using practical values for E_2, $-E_3$: the overall loss factor is therefore in the region 20 or more.

It is worth noting at this juncture that the "worst possible condition" referred to in the above arguments was introduced mainly for comparison purposes. In practice the worst condition might occur only once or twice in a complex net. To design the standard building block around this condition would be uneconomic. The preferred procedure would be to design the building block for the next worst condition which occurred fairly frequently in the logic and so obtain a cheaper unit. The odd cases where worse conditions obtain would be dealt with by change of logic or by employing additional buffer amplifiers. The cost of the few extra amplifier stages and/or gates would be more than compensated by the overall economy effected on the building blocks in a truly minimal cost system.

World Radio History

4. Comparative Summary of Non-linear AND Gate and Diode AND Gate

Non-linear AND	Diode AND
Economy of components. The cost of a standard multiple silicon carbide element has been estimated to be approximately the same as that of a single diode.	
Requires diode clamps, one per driver source.	Negative logic AND-OR / Positive logic OR-AND } sequences require only one clamp diode *per* AND *gate*; also supply current consumption less. For inverse sequences this advantage is lost.
Higher current loss factor will impose restriction on p.r.f. or make transistor OR gate necessary.	
With extra diode can give 2-decision outputs.	
Permits termination of AND gate by more than one OR gate without change of design parameters.	
The economy which can accrue from these properties can only be assessed after complete minimal logic design of network to perform specified requirement but it might be substantial.	If "barred" inputs available these advantages mainly cancelled but cost of two-stage amplifier greater since both stages must be high current stages and an extra clamp diode is required for the first stage.

The possible saving of components due to the two decision property of the non-linear AND gate is exemplified by the schematic circuit of Fig. 21. This represents simply the basic units

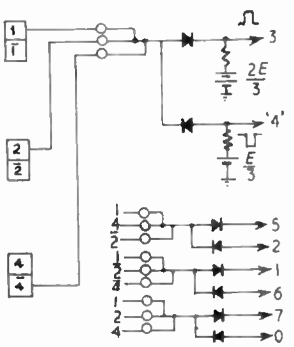

Fig. 21. Binary-octal decoder using non-linear gates.

of a binary-octal decoder fed from flip-flops. Thus, inputs 1, 2, $\bar{4}$ give positive output "3," and negative output "4," etc. Outputs 1, 3, 5, 7 are positive, outputs 0, 2, 4, 6 are negative.

With complementary input connections these polarities are reversed.

Those instances where one or other type of gate is markedly superior are not immediately evident in the general sense; only when the whole system has been subjected to a complete minimal design from both logic and engineering point of view will a final decision be made. Certain cases will be obvious from the beginning, for instance if "barred" inputs are available, positive logic OR-AND-AMPLIFIER or negative logic AND-OR-AMPLIFIER sequences using diodes are unlikely to be surpassed. In other circumstances the extra facilities peculiar to the non-linear AND gate may give it an economic advantage. The foregoing study, although mainly theoretical, gives a positive indication that further research is warranted on non-linear impedance materials and related devices.

5. Constant Current Gates

Reflections on the analogous results which might derive from a characteristic, non-linear in the opposite sense, such as that of Fig. 22,

World Radio History

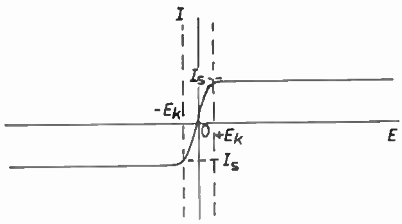

Fig. 22. Alternative (constant current) non-linear characteristic for switching element.

led to some interesting logical functions. This time the non-linearity is of the type known as current saturation and the devices will be designated constant current (CI) devices. Assume that the characteristic is perfectly symmetrical, saturates at current I_s, and has "knee" voltage E_k as illustrated.

5.1. *The basic "2 or more" out of 3 gate*

The circuit is that of Fig. 23(a) in which all elements have the characteristic of Fig. 22 and the applied voltage is E where $E \gg E_k$.

If input E is applied to A and if B, C are held at zero, then element (a) is saturated and the current divides between elements b and c. To provide this half-current through b and c, only a small potential difference need exist across them, hence the output voltage e is in the region of $0{\cdot}5E_k$ and is small compared with E. If input voltage E is applied at both A and B, then element (c) must operate in the saturated condition to take both input currents, whereas (a) and (b) are non-saturated. Consequently the output e rises to a value almost equal to E. For three inputs applied, the output is of course E. To summarize—the unloaded output voltage e is

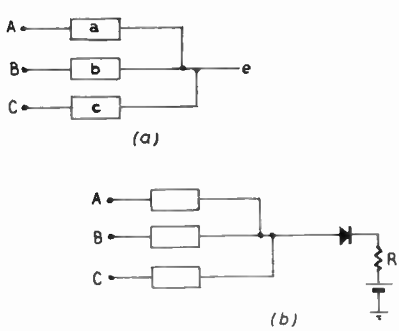

Fig. 23. "2 or more out of 3" gate.

very small for one input raised to E and the others zero; the output is E approx. for two or three inputs applied. The circuit could work directly into a "voltage operated" device such as a vacuum valve or cold-cathode gas-filled tube. With well saturated characteristics no discrimination circuit should be necessary. The saturation current must be reasonably equal in all three elements; the limit would seem to be ± 30 per cent. for the 3-input gate. To work into a "current operated" device such as a transistor, a discriminator circuit is necessary. In Fig. 23(b) a diode discriminator is used; the bias voltage need only be small if the gate elements have good saturation at low voltage; a semi-conductor diode may require no bias (if E_k is small enough).

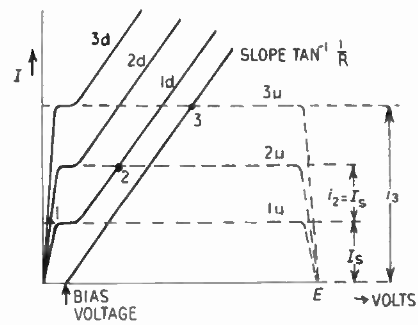

Fig. 24. Operating points for gate of Fig. 23.

In Fig. 24 the operating points for the various input combinations are:—

Point 1, 1 up, 2 down, output current zero
Point 2, 2 up, 1 down, output current i_2
Point 3, 3 up, output current i_3

Current flows in the load only when two or more inputs are applied: the diode may be used as part of an OR gate. Provided that the load resistance R has a value less than E/I_s, the output current i_2 is independent of the magnitude of R and will be equal to the saturation current I_s. Similarly i_3 will be $3I_s$ if $R \ll E_s/3I_s$. The ratio i_3/i_2 is therefore 3. It would be possible to include a second biased diode and load circuit which only passed current for the condition "3 up," i.e. the circuit becomes a dual selection network, a 3-AND gate plus a "2 or more out of 3" gate.

Much the same result may be achieved by

World Radio History

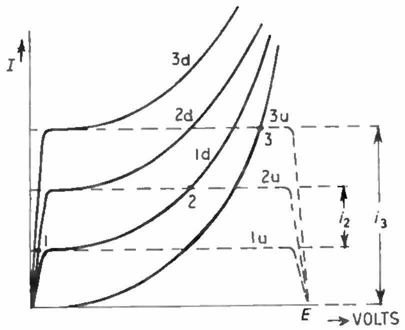

Fig. 25. Operating points for gate of Fig. 23 with the diode replaced by a CV element.

replacing the diode and bias circuit by a CV element. The characteristics of Fig. 25 explain the action.

5.2. *Controlled gates*

The three inputs of Fig. 23(b) might consist of two "signal" inputs A and B and a control voltage C which also has possible values E and 0. Consider the output voltage e under the various conditions remembering that $e \cong E$ for 2 or more out of three inputs applied.

$$C = 0$$

Only if both A and B are equal to E will the necessary two inputs be present; the circuit therefore behaves as a 2-AND gate.

$$C = E$$

Since one input (C) is already at voltage E only one more is required to fulfil the two out of three condition. Therefore if either A or B or both are at V an output is obtained. The circuit behaves as an inclusive OR gate.

The circuit may therefore be modified from a 2-AND gate to an OR gate simply by changing the control voltage at C from 0 to E; a low impedance source is necessary.

This basic arrangement introduces what is thought to be a new concept, that of a computer or other logic system in which signals are not only applied by programme to different logic networks but one in which the action of the logic circuits themselves may be modified by selection of appropriate control voltages. A "3 or more out of 5" gate would have three signal inputs A, B, D, and two control inputs C_1, C_2 (these could be combined into a single control C using a single constant current element of saturation current $2I_s$). This gate would be an OR gate for $C = E$, a 3-AND gate for $C = 0$, and for the two control inputs open-circuit the gate becomes a "2 or more out of 3" gate.

An AND/OR circuit of this type could feed directly into an OR/AND circuit, also of this type, provided that the saturation current value I_{s2} of the second gate elements were less than the corresponding value I_{s1} for the first gate elements. To allow maximum tolerance, it is suggested that $I_{s2} = \frac{1}{2} I_{s1}$. After the second gate, and before any divergent branch points, a buffer amplifier is recommended.

5.3. *A binary digit adder*

This is included as an example of the use of controlled CI gates; it is a very simple arrangement compared with that using diode gates. In Fig. 26 the "2 or more out of 3" gate gives the carry directly; the "3 or more out of 5" gate is normally operating as an OR gate with control voltage E applied at C: if there is an output carry voltage, the control voltage at C becomes zero, and the "5 gate" changes to a 3-AND gate so that the sum digit appears only under the correct input condition. With a suitable short delay before the "sum" digits are read off, the circuit could be used in a parallel arithmetic unit—the carry propagation should be quite rapid (only "1 depth" of gating per unit). The circuit has been tried out experimentally using symmetrical transistors as the constant current devices, and was found to operate satisfactorily on square wave input signals up to a p.r.f. of 100 kc/s. Above this frequency the response time of the symmetrical transistors is too slow because of the high impedance in the base circuit with the result that unwanted short pulses appear in the output.

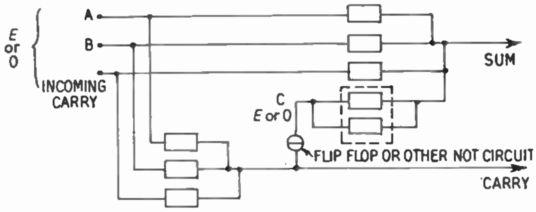

Fig. 26. Binary digit adder using CI gates.

5.4. *Materials for constant current devices*

The phenomenon of current saturation occurs in numerous physical devices. It is encountered in a primary photo-conductor but the current level is far too low for the present application. The saturation of emission from a "bright emitter" is well known but the technique must be considered to be outmoded in this context. A semi-conductor element can be made with two opposed junctions so as to obtain a symmetrical characteristic, each sector being similar to the reverse characteristic of a germanium diode. Elements to give adequate current at or near room temperature would have to be fabricated from material with a low energy gap. The variation of saturation current with temperature would be an embarrassment, particularly as a consequence of dissipation with the elements themselves. With wider gap materials including germanium it would be necessary to raise the temperature. In the latter event, any amplifier in the enclosure would have to employ silicon transistors. If all the logic elements were included in the same volume, temperature control is practically possible but unattractive.

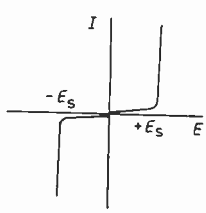

Fig. 27. Characteristic of double Zener diode element.

Another method of obtaining the characteristic involves what is known as the "pinch effect" such as occurs in the field effect transistor. Current flows within a thin semi-conducting sheet and a junction is formed parallel to the conducting path. A bias voltage on the junction controls the effective thickness of the conducting path and so modulates the current which can pass. A device based on this effect is described in the literature†; using silicon, pinch-off voltages from 5V to 15V and saturation currents in the range 1mA to 20mA have been obtained. This pinch-off voltage is rather high for the present application but would be less with germanium as the basic material. Provided that such devices can be manufactured in quantity at reasonable cost there is no doubt a whole series of applications for them in logic circuits such as those outlined above.

† R. M. Warner, Jr., W. H. Jackson, E. I. Doucette, and H. A. Stone, Jr., "A semiconductor current limiter." *Proc. Inst. Radio Engrs*, **47**, pp. 44-56, January 1959.

6. The Dual of the Preceding CI Gates using CV Elements

The element could be two Zener diodes in series opposition or, preferably, a device with two Zener junctions on opposite sides of the same material. Such devices are available in the U.S.A. where they are known as "double-anode" diodes. They are simpler to produce than normal Zener diodes since the diffusion technique automatically produces a junction on both sides of the semi-conductor wafer: multiple elements on the same wafer are also possible. The characteristic is as shown in Fig. 27; the "break" is extremely sharp and the breakdown voltage can be as low as 3V or so.

The dual circuit of the "2 out of 3" gate is that of Fig. 28; the elements are arranged in a closed loop which includes the load resistor R. The sources are constant current sources which supply a train of current pulses to each element. They could be transistors with constant base current (diode limiter at base) such that the transistor does not quite saturate during the relevant part of the signal. These current sources cannot have a common terminal for more than two

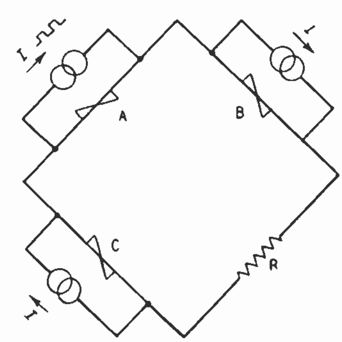

Fig. 28. "2 out of 3" gate using CV elements.

World Radio History

inputs, hence they will have to be transformer coupled and the system is only applicable to dynamic circuitry, e.g. SEAC type logic.

In Fig. 28, if current I flows at A but zero current at B and C, the e.m.f. in the mesh is $E_{s\ peak}$ where E_s is the breakdown voltage. This e.m.f. is insufficient to break down elements B and C in series and hence current i in the load is zero. If current pulses are applied at both A and B, the e.m.f. in the mesh is $2E_s$, which is sufficient to break down the remaining non-energized element C: output voltage E_s appears across the load R. Similarly, current flows in R for all three inputs applied and $3E_s$ appears at the output. To get a useful output current R should be small: a small R is also necessary to reduce the effects of the Zener diode capacitance. This capacitance varies considerably with reverse bias voltage and may be as large as 200 pF when one junction is at a small reverse bias in the single Zener diodes at present available. These are capable of passing a current considerably in excess of that required for gating purposes; a smaller double junction device is desirable and this would have a smaller capacitance.

To operate the circuit as a controlled gate the "control potential" in this case would be a train of pulses; clock pulses could be used and these would retime the output waveforms. If one of the current sources is omitted the gate automatically becomes a 2-AND gate.

In most dynamic systems the transformer is already present in the amplifiers, re-shapers, etc.; the only extra requirement may be additional secondary windings. The double Zener diodes would automatically "square off" the waveform of the transformer output voltage and would make redundant some of the diodes already used in dynamic circuitry to prevent "overshoot," etc. When one current only is attempting to drive two devices the other two diodes are non-conducting, but the signal may pass through the internal impedances of the current generators so producing a load current; for this reason, constant-current generators of high internal impedance are necessary. When these are coupled to the circuit by means of transformers current will flow into the load via the secondary inductance. To prevent a significant rise of load current during a pulse period T, it is necessary for the time constant $2L_s/R$ to be long compared with T; therefore L_s must be made as large as possible to operate at the specified p.r.f. and the value of R is restricted. In practice R can contain the base-emitter circuit of a transistor: when the gate is "open" the current always flows in the same direction during the pulse period.

7. Conclusions

A theoretical examination has been made of the possibility of using bi-directional non-linear switching elements in logic circuits. This examination has revealed some unusual properties which are not available from standard diode gating circuitry. If full use is made of these properties, and if composite multi-electrode elements are made available, then these bi-directional elements might well be used with economy in many complex logic networks.

Preliminary experiments have been carried out which support the theoretical assessment of the circuits. Results of measurements on silicon carbide discs show these to be basically suitable for use in constant voltage gates but that present day material is not sufficiently uniform to make multi-electrode devices a practical proposition. Results of early attempts to find an alternative non-linear material are also reported.

It has been shown that constant current gates would lend themselves readily to a controlled logic system where the logical functions of the gates may be changed without alteration of either input or output connections.

The above assessment has necessarily been limited since materials and devices of requisite performance, tolerance, etc., are not immediately available. Nevertheless, it can be concluded that the bi-directional, non-linear switching circuits are not without promise and that further examination of the ideas and continued investigation into suitable materials may prove to be well worth while.

World Radio History

Radio Engineering Overseas . . .

The following abstracts are taken from European and Commonwealth journals received in the Library of the Institution. Members who wish to borrow any of these journals should apply to the Librarian, stating full bibliographical details, i.e. title, author, journal and date, of the paper required. All papers are in the language of the country of origin of the journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

THE "THYRATRON TRANSISTOR

Some of the applications for the storing and switching transistor which has been developed by the Federal German Post Office and which was described in a paper last year in the *Brit.I.R.E. Journal*, have now been summarized by the inventors. The thyratron-like input characteristic permits its use as an electronic switch or for waveform generation, and examples of applications in telecommunication engineering are discussed in addition to monostable, bistable and unstable flip-flop circuits. The storing and switching transistor may be used in these cases as a circuit element for through-connections, for the control of gating circuits, for storing information, or for the generation of periodic waveforms.

"Applications for the storing and switching transistor." W. von Munch and H. Salow. *Nachrichtentechnische Zeitschrift*, **13**, pp. 301-310, June 1959.

ELECTRONIC PAPER INSPECTION

Companies which manufacture and sell high quality paper, such as bond paper or book paper, must be very careful that defects such as small dirt spots or shiny marks in the paper to not reach the customer. At present the paper is checked by visual inspection one sheet at a time. A considerable number of measurements have been made by the National Research Council of Canada to determine what characteristics of paper at a flaw are different from those of good paper, in order to determine which characteristic is most suitable to be used by an automatic detector to give a fairly reliable indication of the presence of the flaw. The method of detection should be suitable for checking paper travelling at several hundred feet per minute without marking or damaging the paper. On the basis of these tests an electronic flaw detector has been built. The detector measures variations in the amount of light reflected from the paper by means of photomultipliers. Lenses focus the light reflected from a small area of the paper up into the photomultipliers. Rotating mirrors in the path of the light beam swing the beam so that the spot being looked at travels across the paper. The size and shape of the spot is defined by an aperture in the optical system. Variations in the light reflected from the paper produce electrical pulses from the photomultipliers, which are used to actuate the sorting mechanism. Mercury vapour lamps illuminate the paper where it is scanned. The present model can check paper at up to 300 feet per minute, but could be modified to operate at a higher speed.

"The development of an electronic detector of flaws in paper." M. P. MacMartin and N. L. Kusters. *The Engineering Journal of Canada*, **42**, pp. 77-85, June 1959.

TRAVELLING-WAVE TUBE AMPLIFIERS

After preliminary remarks on travelling-wave tubes in radio links, a German paper discusses the most important parameters of travelling-wave tubes which influence the gain, including the different means for focusing the electron beam. The possible upper limits for the gain are investigated, as far as noise and oscillation by internal reflections are concerned. A calculation of the reflections of a non-uniform active helix line with a localized attenuation shows the influence of these reflections on gain, delay and impedance fluctuations. Finally some characteristic measurements are given which confirm the validity of the theoretical investigations.

"The limits of gain in travelling-wave tubes for radio links." W. Klein. *Archiv der Elektrischen Übertragung*, **13**, pp. 273-286, July 1959.

INTELLIGIBILITY OF SPEECH

The intelligibility of syllables is greatly reduced during transmission of speech when strong noise voltages are present at the same time. Intelligibility can be improved when a high quality microphone is used and when the volume is gradually increased with frequency to, for example, 20 db at 3,400 c/s, in comparison with 0 db at 300 c/s. It is shown in a German paper that the improvement depends on the noise volume at the receiving end and on the degree of volume increase. In this case, the overall volume of the speech is reduced so that the volume is not changed by the increase at the higher audio frequencies. The results of the required measurements are described.

"Tests relating to the improvement of the intelligibility of speech in the presence of noise." O. Brosze, K. O. Schmidt and A. Schmoldt. *Nachrichtentechnische Zeitschrift*, **12**, pp. 297-300, June 1959.

World Radio History