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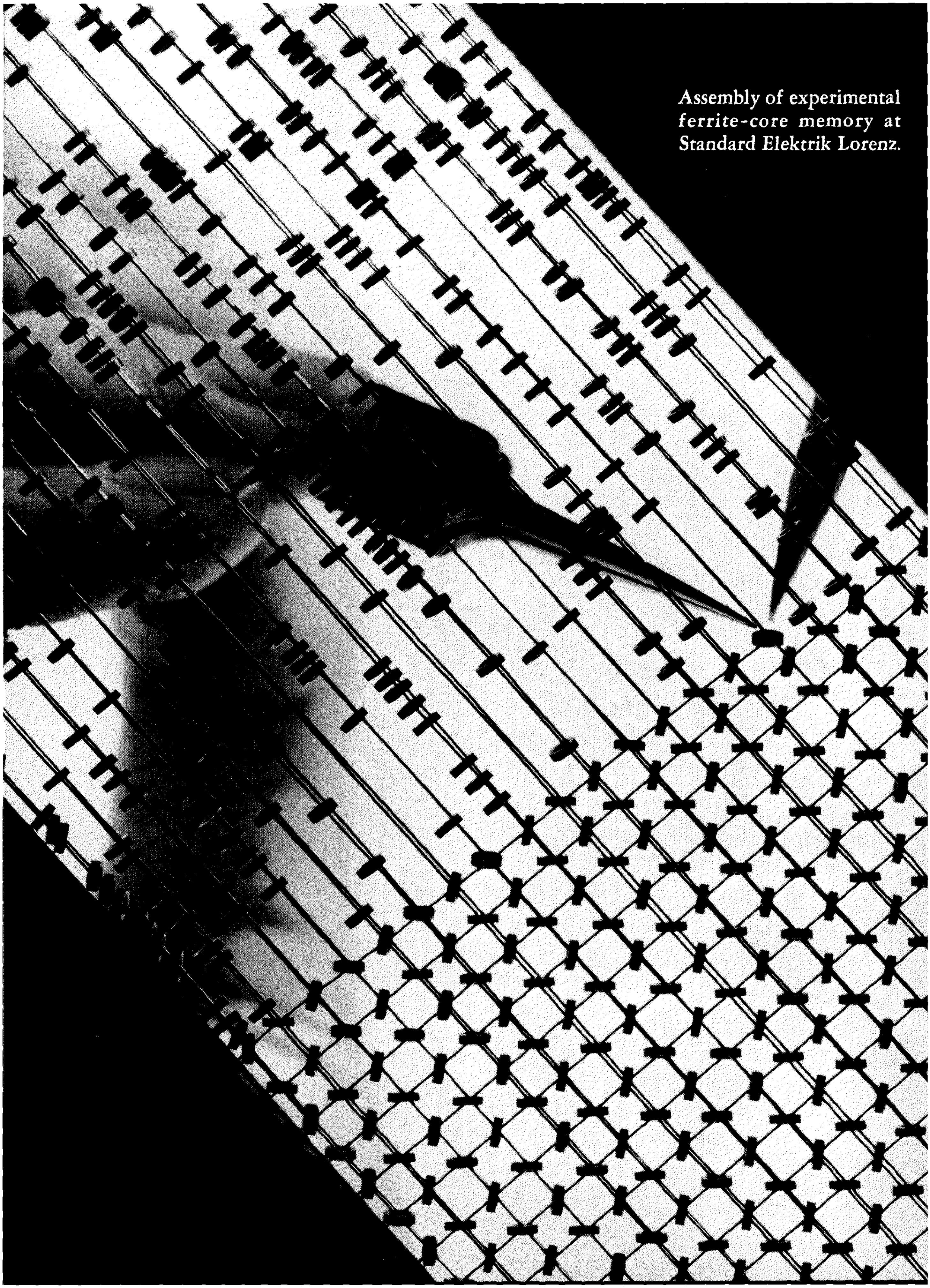

Assembly of experimental ferrite-core memory at Standard Elektrik Lorenz.

Coaxial-Cable Systems: Past and Future*

By A. W. MONTGOMERY

Standard Telephones and Cables Limited; London, England

SEEMINGLY, it is desirable to review some of the history of coaxial-cable systems if the trends of development that can be foreseen are to be correctly understood, since these spring from the developments of the past. Emphasis placed on certain characteristics of coaxial-cable systems has perhaps led to an unnecessarily restricted view of their possibilities. This augurs well for their future, since there remain many potential uses that invite examination.

A coaxial tube, it might be thought, is the obvious means by which telecommunication can be effected by cable. After all, a single wire in air with earth return has many of the features of a coaxial tube. Mathematically, all that is necessary to complete the tube is to wrap the earth around the wire at some suitable distance from it. This in fact was done to a reasonable approximation in an early use of coaxial cable for long-distance submarine telegraphy. In this, the sea, separated from the wire by an insulating layer, became what might today be called the return conductor.

1. History

The people who planned and laid the first transatlantic telegraph cable had tremendous faith, and when eventually this was justified by success, considerable study was made of the properties of coaxial tubes. This was in the middle and later 1800's. A considerable body of mathematical and other literature on the subject developed, of which the greater part is available and useful today.

Since the cable was a success for telegraph transmission, why then did it fail for speech? The answer, known to everyone today, is that speech requires a band of frequencies considerably wider than that needed for elementary telegraphy, and the attenuation of the cable rises very rapidly with frequency. It should be noted also that any bandwidth limitation is dependent on the rate of increase of attenuation with frequency, and not on any other essential characteristics of the theoretical design of the cable itself. This indicated that if amplification became available new possibilities would arise. However, before the invention of amplifying devices, it had been found that circuits could not be satisfactorily obtained from several wires in air along the same route and with earth return because of the resulting interference among them. Quite early, therefore, it became necessary to pair each wire with another to act as a return conductor.

As cable transmission became important, therefore, the same technique of using two wires per circuit naturally and correctly was adopted, and a very considerable art developed in the design of cables to attain smaller and smaller values of attenuation per unit distance and more and more freedom from interference among circuits.

From the theoretical work that had been going on, it was well known, however, that a return conductor completely surrounding a single wire and insulated from it would at high enough frequencies act as a sufficient shield to prevent interference between one tube and another. At this time (early 1920's) a cable system made in this way would have been entirely uneconomic and, as we know, an invention was awaited that would change the whole position. However, short lengths of coaxial cable were used for such purposes as carrying high-frequency currents to the antennae at the top of radio masts.

When the vacuum tube became readily available towards the end of the first World War, its usefulness was extremely great, but it suffered from the difficulty that its amplification–frequency characteristic showed considerable nonlinearity. It was eminently suitable for amplifying speech on existing types of cable pair, even though these required considerable attention to keep interference among circuits within bounds. And because the amplifiers could obviously amplify at frequencies considerably higher than those required by direct speech, attempts continued to be made to get more than one speech channel from each pair in a cable. (There had, of

* Based on a paper presented at the Christopher Columbus Celebrations in Genoa, Italy, on October 10, 1957.

course, been successful attempts before amplification became available to do this by means of such devices as phantom circuits, but these are no part of this theme.) The attempts had some success and attention was concentrated on packing channels as closely together as possible, consistent with satisfactory transmission characteristics, which by now could be specified with some exactitude.

The long-awaited invention that improved the linearity of amplification–frequency characteristics to an extraordinary degree, that is to say, the invention of negative feedback, immediately led engineers to consider means of getting more channels per pair, and so perhaps concentrated interest in only one of several promising directions. Attention also returned to the now-practicable use of coaxial tubes, which was attractive because of the potentially very large numbers of channels obtainable from them. It was apparent that a considerable amount of equipment would be necessary to locate an individual channel in its special place in the frequency spectrum. Terminal equipments therefore were from the beginning expected to be somewhat costly. The cost could be decreased by reducing to the minimum the amount of equipment required for each individual channel, and by making maximum use of equipment common to many channels.

These various concepts suggested that coaxial systems were most suited to routes over which large numbers of channels were required for long distances. The logic of this, however, obscured the more important truth that in fact the coaxial tube with its amplifiers was a means for obtaining a large bandwidth at a low cost, and that its economical use was not dependent on its use for large groups of channels and over long distances. What is clear, however, from the historical sketch given so far is how the earlier belief developed; although it should have been apparent even then that any coaxial-cable system has as its chief merit the provision of a wide frequency band at low cost.

2. *What is Bandwidth?*

It is desirable to remind ourselves of what we mean by a wide available band, and what are the causes that set any limit to the frequencies that can be used. The lower limit of frequency might be fixed by the point at which the loss of screening effect of the return conductor becomes important: but in fact some frequency higher than this (such as 60 kilocycles per second) is selected to ease the solution of equalisation problems. The higher limit may be set in practice by the effects of small departures from continuous cylindrical symmetry in manufacture or installation, which result in small impedance irregularities and consequent reflection effects. With careful design and manufacture, these can be kept under control to almost any desired extent; pulse testing methods give exact and extremely useful information both for design and inspection purposes. Figure 1 shows the results of a typical measurement. Although the variations from the mean value are very small, yet each can be assigned to a specific cause at a definite place in the cable should it be important to do so. Figure 2 shows an assembly of test equipment for the purpose.

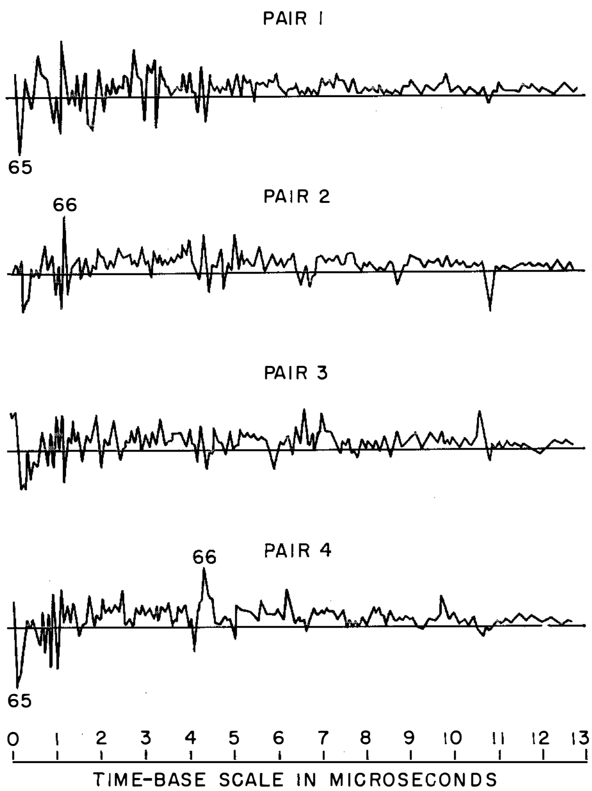

Figure 1—Oscillograms of reflections of a test pulse applied to an installed length of coaxial cable. The magnitudes of the largest reflections are stated in decibels below the incident pulse level. The incident pulse width was 0·1 microsecond.

In the end, however, the upper limit is fixed by economic considerations, compromise being effected by giving appropriate weight to factors such as required use of the circuits, possible designs of cable, spacing of repeaters, and design of the system as a whole to take care of known relationships between noise and amplification, and similar properties.

Again emphasising that a coaxial system is essentially an inexpensive one, nevertheless means were, and always will be, sought to reduce cost still further. A considerable saving was made by transmitting along the cable itself the power needed to operate at least some of the repeaters. Since the low frequencies were not in any case available for telecommunication transmission, this did not restrict the band. Accordingly the power losses that occurred along a cable of particular dimensions and the maximum voltage that could be impressed on the cable became factors in the design. Another series of factors was based on the maximum power output obtainable from a repeater, and in consequence the number of individual channels that could be obtained, making due allowance for the fact that all channels would not be in use at all times, nor would the powers in each channel be identical from moment to moment. There are many such sets of conflicting requirements, as there are in all telecommunication system designs, and these few are mentioned merely to illustrate the fact that with such a variety of factors to be taken into account undoubtedly many equally attractive solutions could be reached.

The solution by compromise is frequent in telecommunication problems and fortunately the chaotic conditions that could have resulted have been avoided by agreement among telephone administrations to recommend essential features for general adoption. The point to emphasise is that, through the medium of the Comité Consultatif International Télégraphique et Téléphonique and the wisdom of its members, it has been possible to select certain important features, possibly

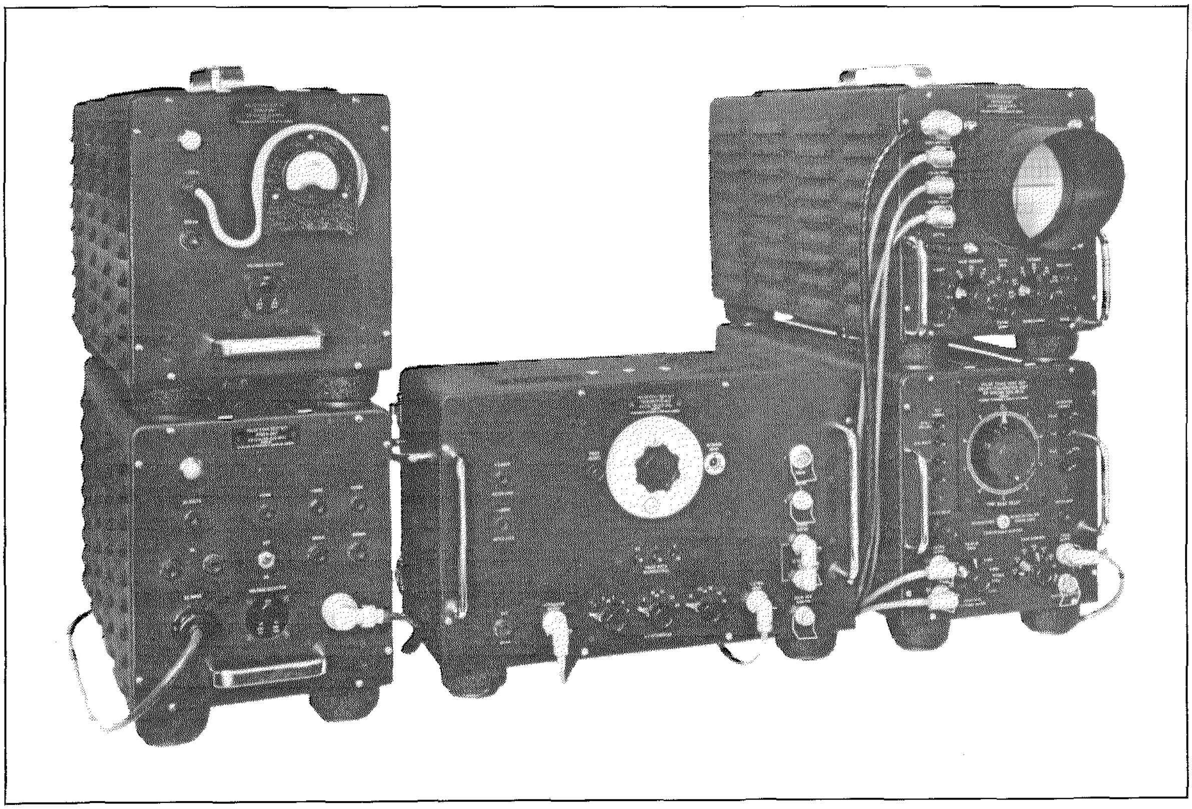

Figure 2—Assembly of equipment for making pulse tests on coaxial cable.

taken from a variety of types of system presented to them, to recommend these for general use, and so to enable world-wide communication to be effected most economically today and tomorrow. It obviously becomes increasingly important as invention makes communication more and more easy over greater and greater distances that requirements continue to be established wisely. They may not always be attainable immediately in their entirety, perhaps because of the cost of changing existing things, but they may eventually be met and maintained throughout the world.

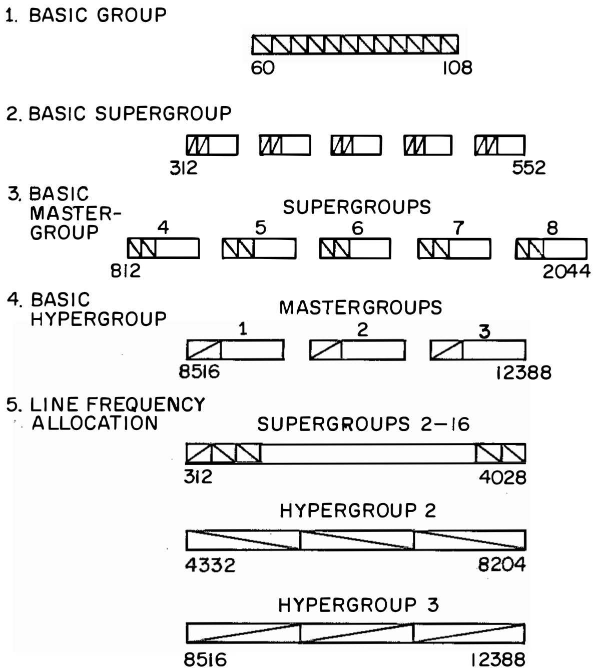

Figure 3—Wide-band coaxial telephone system adopted by Comité Consultatif International Télégraphique et Téléphonique. The line frequency allocation consists of hypergroups 2 and 3 and either hypergroup 1 or supergroups 2 through 16 as shown. Frequency limits are indicated in kilocycles per second.

3. *Increasing Frequency*

The extension of the transmission band into higher frequencies is already particularly important today, and not least in the field of coaxial-cable systems. For there are many ways in which progress is being made and bandwidth increased by one means or another. The continual increase in the usefulness of the bandwidth available from 0·375-inch (9·5-millimetre) coaxial cable has led to discussions by the Comité Consultatif International Télégraphique et Téléphonique concernin the use of frequencies up to about 12 megacycles per second for telephone and television systems, and provisional recommendations have been made. Figure 3 reminds us of the proposed frequency allocation, and illustrates well that it has been possible to build on the earlier recommendations as demands for channels have increased.

Equipment design becomes increasingly difficult as the frequency band increases, but remains within the bounds of economic achievement. Figures 4 and 5 show a typical amplifier already available to cover the 12-megacycle-per-second band. Despite the rapid growth of demands for channels, the 12-megacycle-per-second system should be adequate for some time to come. This is expected to be the case even when the transmission of colour-television signals is taken into account, since its bandwidth requirement is commonly expected to be within the limits of this system.

4. *Increasing Attenuation*

There is also another totally different approach to the quest for circuits. From the beginning, because the use of coaxial systems was thought of in terms of large groups of channels, there has been reluctance to part off from the cable small groups of channels at intermediate points. This could always have been done economically, but insistence on getting the maximum number of channels from a line has always led to some misgivings about this application. And as the demand for channels increases and pressure is brought to bear to obtain more and more channels from the available bandwidth, there develops even greater reluctance to attempt to drop small numbers of channels at intermediate points.

The question of the use of a coaxial-cable system for a relatively small number of channels becomes increasingly interesting. When considered in the past, as it was from time to time, the difficulty always proved to be that of providing an adequate number of repeaters of very small cost. Of course, the cost of a repeater must always include the cost of the building that houses it and

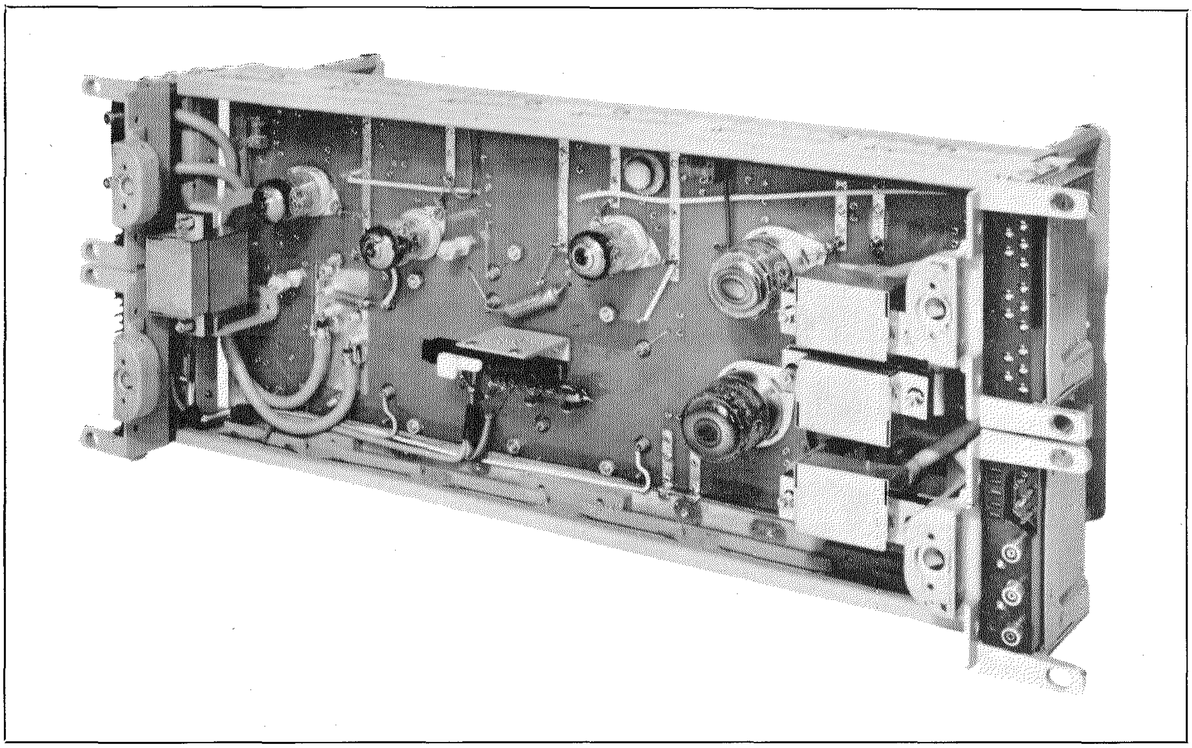

Figure 4—Amplifier for the 12-megacycle-per-second band that accommodates three hypergroups.

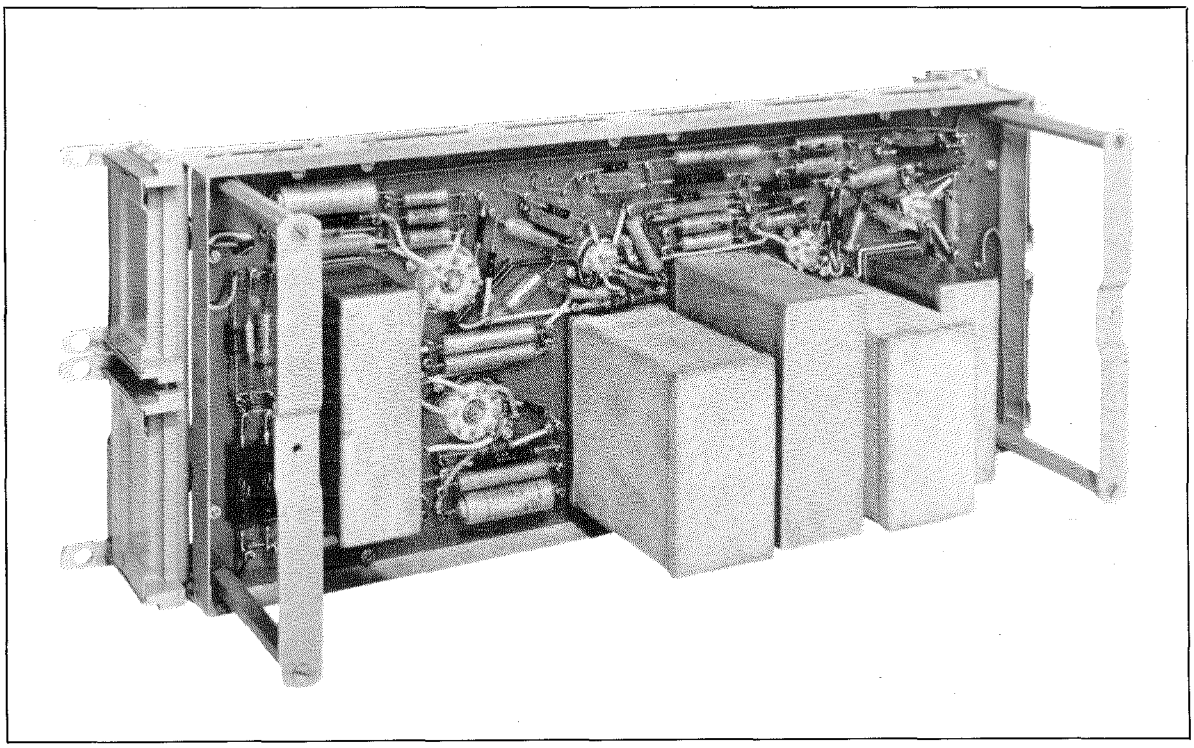

Figure 5—View of reverse side of amplifier shown in Figure 4.

of the power that must be supplied to it. Unattended repeaters, whose operating power is transmitted along the cable, have necessitated such total voltages as have limited reduction in size of the cable or, viewed differently, have imposed a limitation in the numbers of them that can be operated in tandem. An economic difficulty has always been that the 0·375-inch- (9·5-milli-

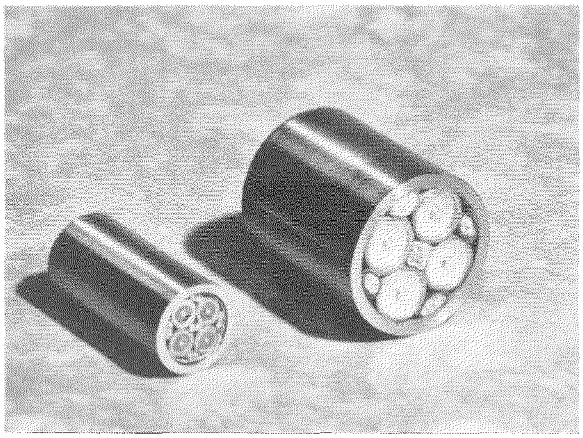

Figure 6—Cross-sectional view of 0·375- and 0·163-inch (9·5- and 4·1-millimetre) tubes in cables.

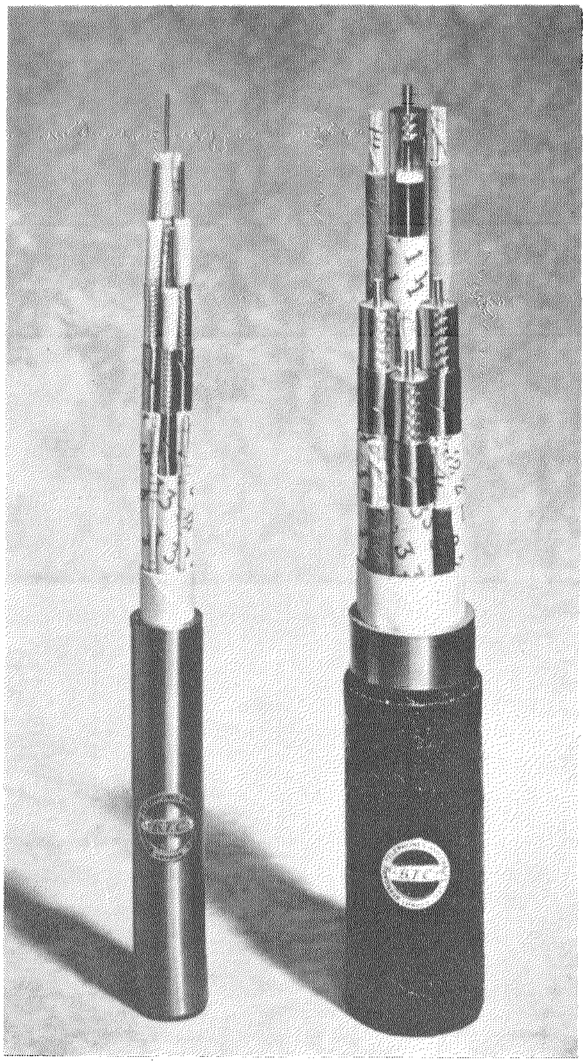

Figure 7—Stripped view of the 0·375- and 0·163-inch (9·5- and 4·1-millimetre) cables.

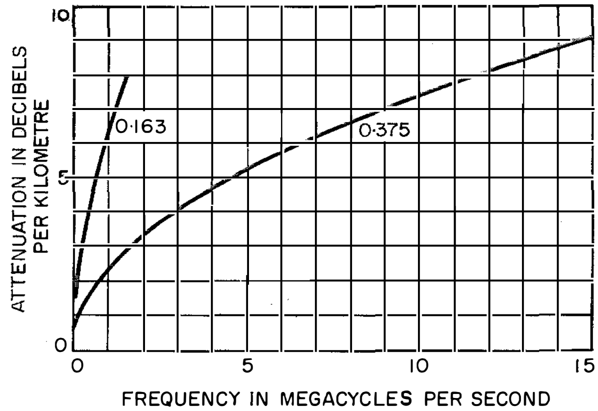

Figure 8—Comparison of the nominal attenuation at 10 degrees centigrade of the 0·163- and 0·375-inch (9·5- and 4·1-millimetre) coaxial tubes are a function of frequency.

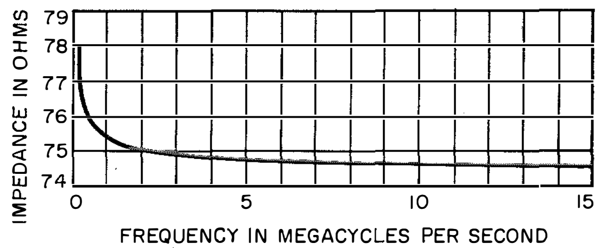

Figure 9—Nominal impedance of 0·375-inch (9·5-millimetre) tube versus frequency.

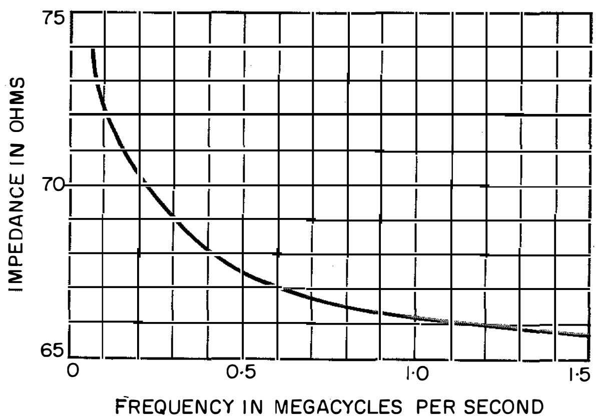

Figure 10—Nominal impedance of 0·163-inch (4·1-millimetre) tube plotted against frequency.

metre-) cable system is essentially economic even for quite short distances, and it has been correspondingly difficult to design a system that would show substantial economies if limited to even shorter distances.

The position changed considerably with the invention of the transistor, chiefly because it is essentially a low-voltage device. As the production of reliable transistors becomes assured, it is reasonable to contemplate their use in stable equipment. Because the transistor requires only moderate voltage and consumes little power two things change when comparison is made with valve-operated devices. These are the size of cable and the repeater spacing, both of which can be reduced without restricting in any way bandwidth availability. The function to be performed by these smaller cables would obviously be that for which the existing cables, however illogically, were considered not to be suitable. It would be reasonable to design initially for a limited number of circuits to be used over short distances. Cable attenuation could therefore be allowed to increase faster than the amplification obtainable by the insertion of additional repeaters.

From these considerations a system has been developed that is designed to provide up to 300 two-way speech channels, in general of Comité Consultatif International Télégraphique et Téléphonique quality, over a pair of coaxial tubes each of 0·163-inch (4·1-millimetre) diameter. The frequency band used lies between 60 and 1300 kilocycles per second and the repeater spacing has been fixed for this number of channels at approximately 12 000 feet (3·66 kilometres), this distance being a multiple of the accepted loading coil spacing of 6000 feet (1·83 kilometres). It is naturally not compulsory to use so many channels as this. Figures 6 and 7 show one version of the new 0·163-inch (4·1-millimetre) cable and a 0·375-inch (9.5-millimetre) cable for comparison. Figures 8, 9, and 10 show attenuation and impedance curves. Figures 11, 12, and 13 show the repeater and illustrate methods of connection and installation.

The transmission system itself is based on existing recommendations of the Comité Consultatif International Télégraphique et Téléphonique in that the basic 12-channel group in the frequency band from 60 to 108 kilocycles per second is used and supergroups are formed in the frequency band from 312 to 552 kilocycles per second in the

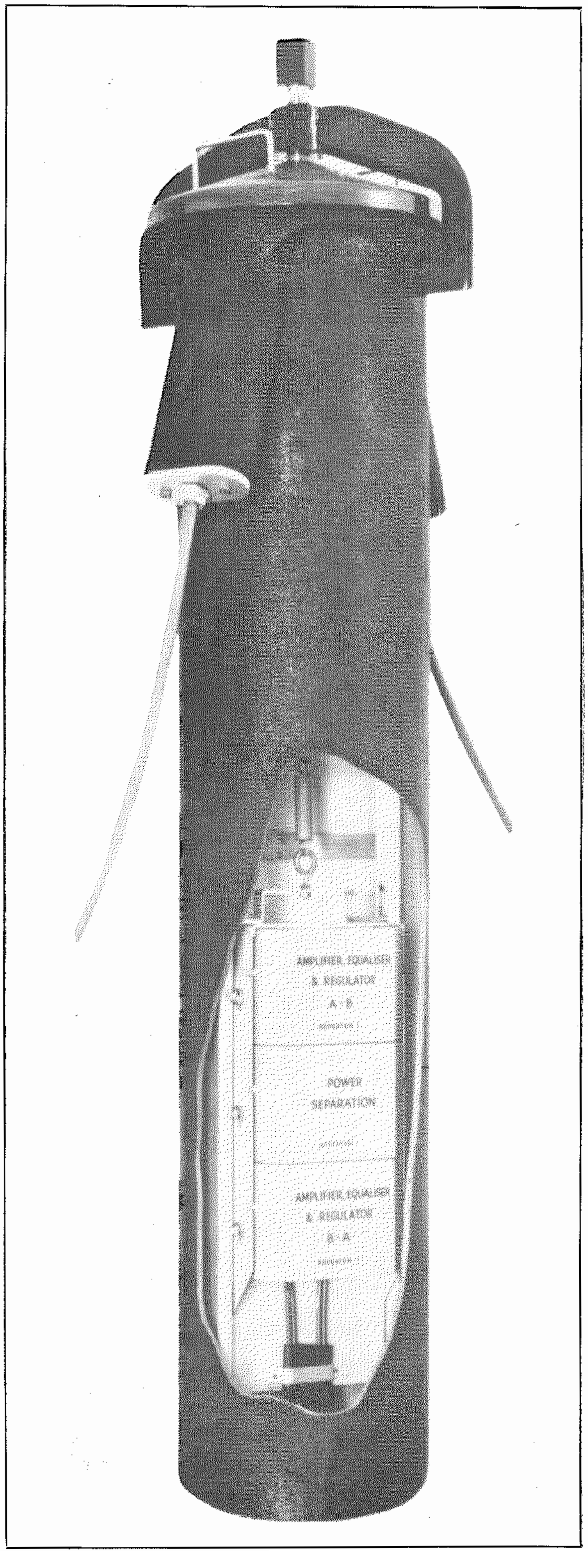

Figure 11—Two transistor repeaters and the power separation circuit for the repeaters are in a sealed housing.

Figure 12—Repeater housing projecting from the ground.

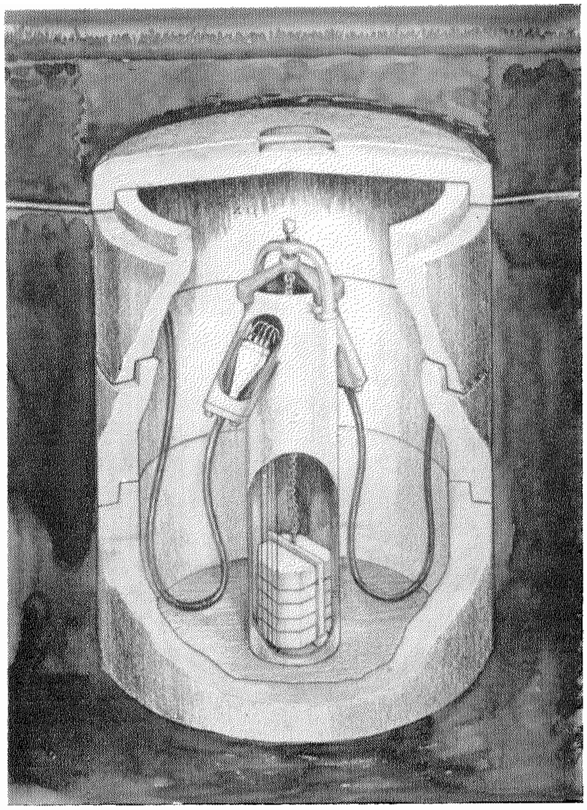

Figure 13—Cutaway view of repeater in manhole.

usual manner. Five of these supergroups may be used and will permit interconnection between this system and other approved systems by channel, group, or supergroup as desired. Equally, of course, programme channels can be obtained in a normal manner by combining 3 or 4 adjacent telephone channels.

5. *Future Requirements*

These two coaxial systems offer between them very great flexibility, which may well be very helpful in solving problems that arise with the increasing use of long-distance dialling by subscribers. Groups of channels of a variety of sizes and operating economically over a variety of distances become immediately available and accessible and can be rearranged very quickly to meet circuit requirements, whether caused by the growth of traffic or by the provision of additional exchanges of present or future types.

The future of coaxial-cable systems appears now to point in two main directions; *one*, in the use of higher and higher maximum frequencies for increasingly greater distances; and *two*, in the use of higher-attenuation cables providing wide frequency bands over shorter distances. Figure 14 is interesting in that it shows on the same graph the attenuation characteristics of the two cables mentioned, the frequency scales being different

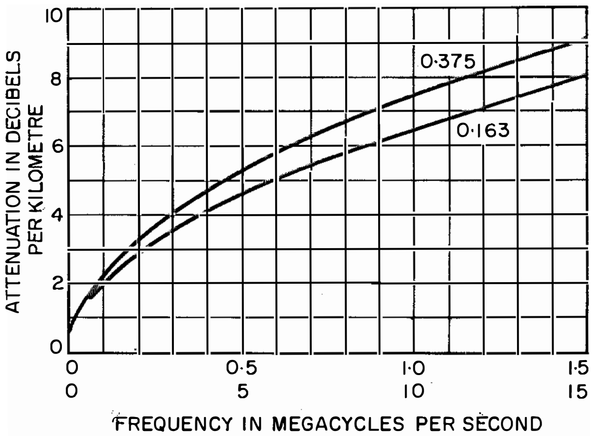

Figure 14—Nominal attenuation at 10 degrees centigrade for the two sizes of coaxial tubes. The higher-frequency scale is for the larger cable.

for the two curves. They illustrate once more that to speak of a wide band of frequencies is to speak of something that is adjustable as necessary.

However, reverting to the effects of history as outlined above, we may consider once more whether we are in fact not falling into our own trap by still regarding the coaxial-cable system as a means for providing a large number of channels closely packed in the frequency spectrum. Consequently, we should enquire whether it would not be interesting to look again at the economic use of the available wide bands. Because of the low voltage and power requirements of the transistor it may be reasonable to contemplate the use of repeaters at even closer intervals to make available even wider bands. Pulse systems, notably pulse-code-modulation systems, might then offer themselves as suitable for efficient economic use of the band, since one can often pay for simplicity of apparatus in bandwidth; and fortunately the transistor is eminently suitable for use in regenerative repeaters, which are reasonably simple. It remains to be seen however, whether the historical approach, which results in a demand for more and more channels from a given bandwidth, would allow a pulse-code modulation system to retain its simplicity, and correspondingly its economic justification.

6. Conclusions

From all of the above, one may conclude that the prospects for coaxial-cable systems are bright on account of the potentially available great bandwidths and the essential long- and short-term stability of the cable and its associated equipment. This means that whatever the type of transmission system evolved in the future or the demands made on any of them for increased maximum frequencies, there should be no difficulty in providing the appropriate transmission path. Also, the essential smoothness of the cable characteristics permits additions to equipment to be made that, at relatively small cost, will allow coaxial cables already installed to provide wider and wider frequency bands.

Perhaps the future prospects, too, are even brighter, since data and other information-transmission systems belonging to arts now in their infancy, will make increasing claim for very-stable circuits of as yet unknown bandwidth. Coaxial-cable systems with suitable maintenance are eminently adaptable to this type of requirement.

This paper has dealt only with main principles, which have been illustrated in part from existing practical embodiments of coaxial-cable systems. Having started with the deep-sea coaxial cable, it may be appropriate to terminate it in the same manner. As most excellently described previously by Sir Gordon Radley and Dr. M. J. Kelly, the use of amplification has already freed some more of the frequency band imprisoned in the submarine cable; and so the story feeds back to its beginning.

Some Modern Developments in Telegraph Transmission Equipment

BY W. F. S. CHITTLEBURGH

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EXPANSION of telegraph networks about 1930 was generally such that, because of the greatly increasing cost of telegraph operators, it became necessary to avoid manual retransmission. Administrations then had to decide between using some form of tape relay or introducing numerous separate point-to-point channels. Economic and other factors favoured the point-to-point arrangement, and at that time there was a universal changeover from time-sharing multiplex to frequency-division multiplex channels. The main trend in recent years has been towards longer point-to-point networks and the introduction of automatic switching of teleprinters. This has had two main effects on present-day voice-frequency telegraph design: firstly, a demand for the use of larger numbers of telegraph links in tandem, 6 to 8 in some cases; and secondly that the system be capable of resting for long periods in an idle condition, corresponding in amplitude modulation to a no-tone condition, and yet respond to the first incoming pulse with small distortion. It will be obvious that if 6 or 8 links are to be connected in tandem the distortion per link has to be kept low.

1. Theoretical Considerations

1.1 Telegraph Distortion

Telegraph distortion[1] can conveniently be divided into three components; characteristic, bias, and irregular or fortuitous distortion.

Methods of computing these various components into the probable distortion of several links in tandem are not yet fully established.[2] However both theory and experiment indicate that the characteristic distortion of n links in tandem tends to be n times that of one link, whilst irregular distortion is proportional to $n^{1/2}$. Though bias is algebraically additive it is due to random degrees of maladjustment, and the probable overall bias is again $n^{1/2}$ times the probable bias of one link.

The design of modern voice-frequency telegraph systems has been to keep the total distortion per link low, paying particular attention to the characteristic distortion, even to the extent of allowing interference from adjacent channels to increase somewhat; this involves the careful design of the amplitude–frequency and phase–frequency characteristics of the channels. The requirement of small distortion of the first mark signal after a long space condition has had its reaction on the design of the automatic level-adjusting means for amplitude-modulated systems.

1.2 Channel Filters

In achieving low characteristic distortion, a fundamental concept of channel filters has changed. A square-shaped discrimination characteristic, flat within the pass band, is no longer a neccessity. In fact it is the phase characteristic within the pass band that is more important; it should be kept reasonably linear. This concept applies equally to all forms of telegraph systems where the optimum ratio of signalling to bandwidth is required.

1.3 Amplitude Modulation

It is interesting to review some of the present methods that are, and can be, employed to convey telegraph signals through a channel using alternating currents restricted to a narrow band of frequencies.

Firstly with amplitude modulation the requirement that has arisen within recent years for the system to be capable of resting in the no-tone condition for long periods, and yet respond accurately to the first incoming pulse, has had its reaction on the automatic level-adjusting means. Most narrow-band systems relied, and

[1] Comité Consultatif International Télégraphique, "Draft List of Essential Telegraph Terms"; 1955.

[2] Comité Consultatif International Télégraphique et Téléphonique, Questions 8/9 and 9/9, Study Group 8; 1957–1960.

still do in some cases, on using a long time constant to "memorise" the incoming amplitude and to provide automatic gain and bias control. When tone was sent to line between messages it was only necessary to remember for the longest spacing period in a message, that is, 100 milliseconds or so. With very-long spacing periods this method of working became inadequate, and the next step was to put a delay in the signal path so that there was time to measure the incoming amplitude, set the automatic bias control, and then detect the signal. By this means quick-acting automatic level control was obtained that was capable of operating precisely on the first incoming signal after a long space.

Amplitude-modulated systems have been in use now for a sufficiently long period on both national and international networks that almost all the requirements[3] for such a system have been laid down. It is to be noted that although these requirements for amplitude-modulated equipment have changed little within the past few years, the actual performance of equipment being developed has improved to the extent that maintenance of these systems has been considerably eased.

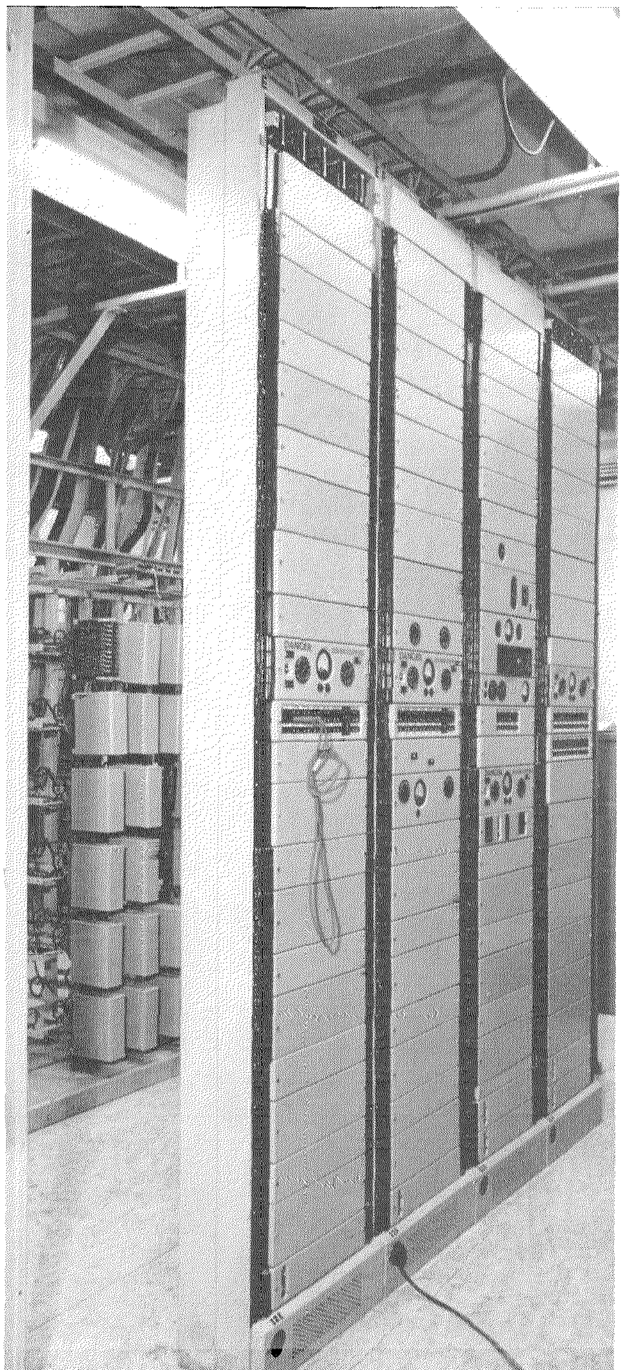
1.4 Frequency-Shift Modulation

Though frequency-shift-modulated systems have been developed for several years now their use generally has been somewhat limited and the Comité Consultatif International Télégraphique et Téléphonique, taking advantage of these facts, is at present attempting to get agreement on the requirements for such systems when used internationally.[3,4] These systems possess the advantage of being very insensitive to changes of amplitude, but they are generally quite sensitive to frequency instability.

Frequency-shift modulation was for a long time considered only as a wide-band method, and only in relatively recent years has it been realised that it can give a ratio of working speed to bandwidth comparable to amplitude modulation. Narrow-band frequency-shift modulation, as used today, actually has an amplitude variation during the frequency transitions, and this is a vital part in the process for obtaining the full ratio of signalling speed to bandwidth.

It is well established now that for the same speed-to-bandwidth ratio a frequency-shift-modulated channel can have at least a 6-decibel signal-to-noise advantage over an amplitude-modulated channel. However, the fact that tone is always present in the frequency-shift-modulated case involves the penalty of increased power loading of the transmission media, and

Figure 1—View of Oban terminal station of the transatlantic telephone cable showing some of the frequency-shift-modulated voice-frequency telegraph equipments used between Great Britain and Canada.



[3] Comité Consultatif International Télégraphique et Téléphonique, Documents of VIII Plenary Assembly; 1956.

[4] Comité Consultatif International Télégraphique et Téléphonique, Question 16/9, Study Group 8; 1957–1960.

the 6-decibel advantage is reduced somewhat. The Comité Consultatif International Télégraphique et Téléphonique has agreed that for international working of a 24-channel system over a telephone circuit the level per channel of a frequency-shift-modulated telegraph system should be reduced by 1·5 decibels relative to the corresponding level of an amplitude-modulated system. Nevertheless it has also been agreed that on certain types of telephone circuits where the extra power loading is not important, for example, audio circuits and open-wire carrier circuits et cetera, the same level may be transmitted for frequency-shift-modulated as for amplitude-modulated channels, so that the full 6-decibel advantage may be maintained.[3]

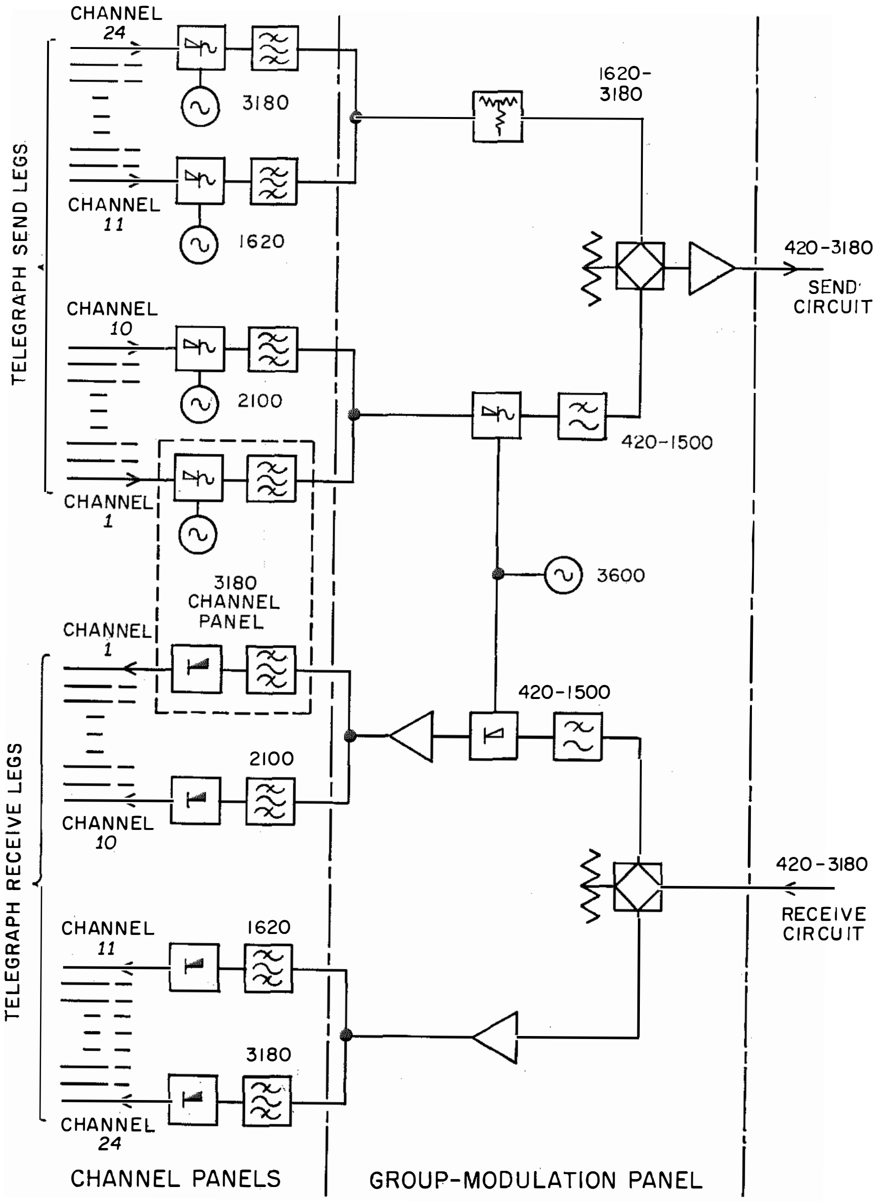


Figure 2—Group modulation arrangements in the *TA4* amplitude-modulated voice-frequency telegraph system. The frequency designations are in cycles per second.

Other advantages of frequency-shift modulation are that the level range over which the frequency-shift-modulated channels will operate is very wide, and that sudden changes of level cause only small increases of distortion, and these only when the change of level occurs at, or near, a signal instant. Unlike an amplitude-modulated channel, should the tone of a channel disappear, due to a fault condition, the output of the detector is left with no absolute control, and the channel becomes sensitive to noise. In this respect the Comité Consultatif International Télégraphique et Téléphonique has recommended that the equipment should be so designed that when such a fault occurs, the output of the receiver will be held in a start condition, and its response to noise prevented.[3]

1.5 Phase Modulation

Phase-modulated systems are extremely limited both in design and extent of application; as such they are, at the moment, of minor importance. These systems generally are of the phase-reversal type, the phase of a constant frequency being reversed at each transition from mark to space and vice versa. The disadvantage is that there is no reference condition to indicate whether a mark or a space is being sent. This means that should a transition be inserted or lost due to noise or other mischance the information is completely reversed from then onwards. Until now no solution, for start–stop operation, has been proved, though there is every hope that this may be done in the near future.[5]

There is a phase-modulated system[6] being marketed that is of interest. The individual voice-frequency channels are of a diplex nature, it being necessary to detect 4 different conditions of phase per channel. The system uses rhythmic transmission, and additional equipment is provided to convert, where necessary, the signals from arhythmic to rhythmic.

2. *System Features*

2.1 Sending Equipment for Amplitude-Modulated Systems

The older amplitude-modulated systems were designed, especially for large installations, to use multifrequency motor-generators as a source for the various channel tones, though in small installations individual oscillators for each channel could be provided. Though the multifrequency generator is still used, the tendency with modern systems is to provide individual oscillators for all channels, thus providing considerable flexibility, enabling small groups of channels to be assembled initially and to be increased as and when the need for more arises.

There is a technical reason also for the departure from multifrequency generators, especially from those that produce only the frequencies for channels 1 to 18. There is always a tendency when the lowest-frequency channels are directly modulated for the telegraph distortion of these channels to be somewhat greater than that on the higher-frequency channels; this is due to harmonic components of the modulating waveform interacting with the carrier frequency of a channel; it is known as carrier beat. There are methods whereby this interaction can largely be eliminated; in the *TA2* system, tones were modulated directly by means of a carefully designed double-balanced modulator, the direct-current input waveform on the receive leg being slightly filtered to remove the high-order frequency components. In the *TA4* system a simple modulator is used but only the tones of channels 11 to 24 are produced by direct modulation. Channels 1 to 10 are produced by group modulation of a second group of frequencies corresponding to channels 24 to 15. As will be seen in Figure 2, a similar group-demodulation process is performed at the receiving equipment. An advantage gained by this group modulation is that the coils and capacitors required in the channel equipment have convenient values and can be of small physical dimensions, enabling all channels to be made of the same minimum size.

2.2 Receiving Equipment for Amplitude-Modulated Systems

The design of the detector circuit of the early equipment has been described previously,[7] but it may be of interest to indicate briefly the circuits used for *TA2* and *TA4* equipments. In both of these the alternating-current amplifiers are linear over the operating range and are of constant gain; automatic bias control is used to compensate for changes of input level. The *TA2* equipment shown in Figure 3 utilizes a bias circuit having a quick charge and a very-long discharge time. The signal path includes a delay network, enabling the bias circuit to be charged to its correct voltage by the time the signal is applied to the output. This long discharge time on the bias circuit, though enabling the equipment to be designed and operated with very low characteristic distortion, is relatively inefficient when the mean incoming level suddenly drops;

[5] H. T. Prior et al, British Patent 693 704; September 8, 1950.

[6] E. T. Heald and R. G. Clabaugh, "Predicted Wave-Signalling Phase-Shift Telegraph System," *Communications and Electronics*, number 31, pages 316–319; July 1957.

[7] J. A. H. Lloyd, W. N. Roseway, V. J. Terry, and A. W. Montgomery, "New Voice Frequency Telegraph System," *Electrical Communication*, volume 10, pages 184–199; April, 1932.

the output signals suffer from a high degree of distortion until the bias circuit has discharged to its new correct voltage. Under normal conditions this system gives exceedingly good performance; its characteristic distortion for mixed messages is only about 3 per cent for 50-baud signals and 8 per cent for 80-baud signals.

In an endeavour to avoid the long time constant and at the same time to effect considerable simplification the *TA4* circuit[8] was developed. In this, as shown in Figure 4, the biasing circuit has two biases derived from the received signal; one for biasing the space–mark transitions, and the other for biasing the mark–space transitions. The circuit is so designed that, of the two biases, only that voltage that is the greater will be effective, each bias being independent of the other. The discharge time is made just sufficient to maintain the correct bias voltage for the instant of a mark–space transition, but not so long that it interferes with the new bias for the next space–mark transition. Thus every signal transition derives an individual bias, and very nearly instantaneous response of the equipment to sudden changes of level is achieved. Naturally if a change of level occurs at or near a signal instant, the modulation products of the level change will interact with those of the normal signal transition and may produce distortion, but this is liable to occur with any form of modulation when operating over narrow-band channels. The *TA4* system, because of its particular detector circuit, has been designed primarily

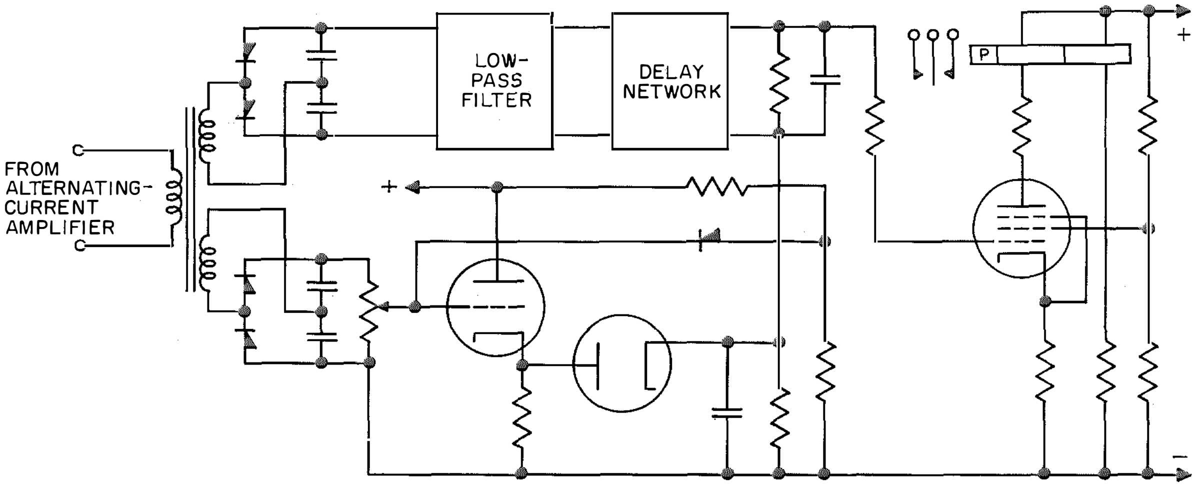


Figure 3—Biasing circuit of *TA2* amplitude-modulation detector.

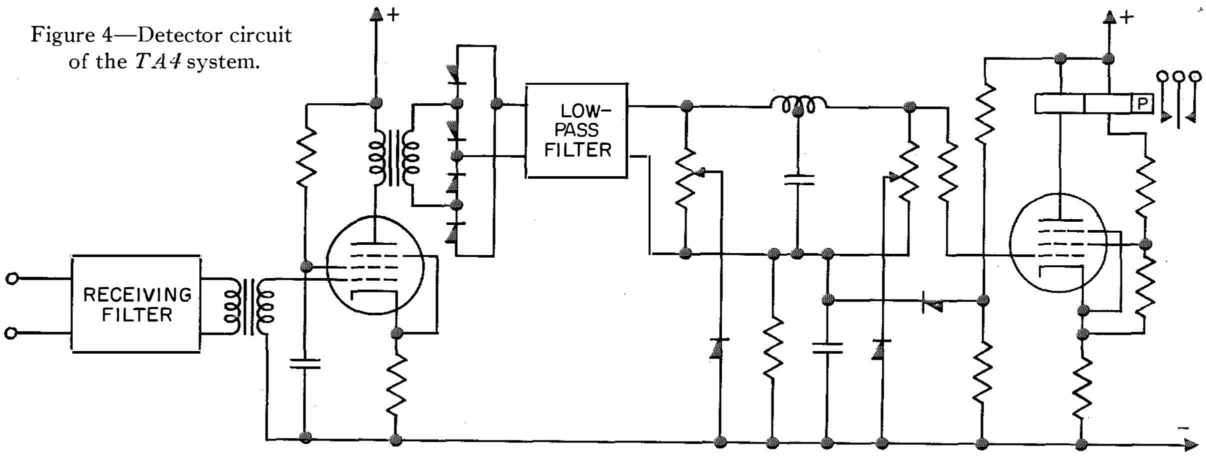


Figure 4—Detector circuit of the *TA4* system.

[8] H. T. Prior, British Patent 693 769; September, 1950.

for 50 bauds, or lower, speed of working, especially if several such links are connected in tandem; but where only one link is contemplated higher speed may be employed. The characteristic distortion of this system at 50 bauds is about 3 per cent, and even with signals predistorted by amounts up to 35 percent the increase of distortion will not be more than the above characteristic distortion value.

2.3 Frequency-Shift-Modulated Systems

When the frequency-shift-modulated system *TF1* was developed[9] it was thought that its major application would be for extremely noisy line circuits or for radio working. To a large extent because of the proposed radio application, the channel equipment was engineered as two separate panels as may be seen at the left of Figure 5. The upper panel mounts the oscillator and modulator; the larger, the detector. The use of frequency-shift-modulated systems is now becoming more widespread for line working whilst for radio working there are further requirements, so the latest frequency-shift-modulated channels have been engineered with both send and receive circuits on the same panel.

To compensate for effects of variation of the ambient temperature, the *TF1* system has individually adjusted compensation circuits. It was arranged that the mean frequency of the oscillator in the modulator remain constant with change of temperature; but the compensation in the detector circuit did not attempt to correct the tuning of the discriminator coils, compensating instead the bias distortion caused by any drift of tuning due to temperature. Another refinement was the provision of a 300-cycle pilot channel to correct for any frequency drift that might have occurred between the modulation and demodulation carrier frequencies of the vehicle telephone channel. A frequency error from this source causes equal bias to the signals of all channels, thus, by detecting the drift over the pilot channel, a common compensating bias was applied to all channels.

The advent of improved materials for resonant units has made it possible to provide circuits for the *TF3* system that do not require special temperature compensation. Pilot equipment again can be provided, in this case at either 300 or 3300 cycles. As with the amplitude-modulated equipment *TA4*, to enable all channels to be of the same small physical size, channels 1 to 6 are derived by group modulation. Channels 19 to

[9] W. F. S. Chittleburgh, D. Green, and A. W. Heywood, "Frequency-Modulated Voice-Frequency Telegraph System," *Post Office Electrical Engineers Journal*, volume 50, pages 69–75; July, 1957.

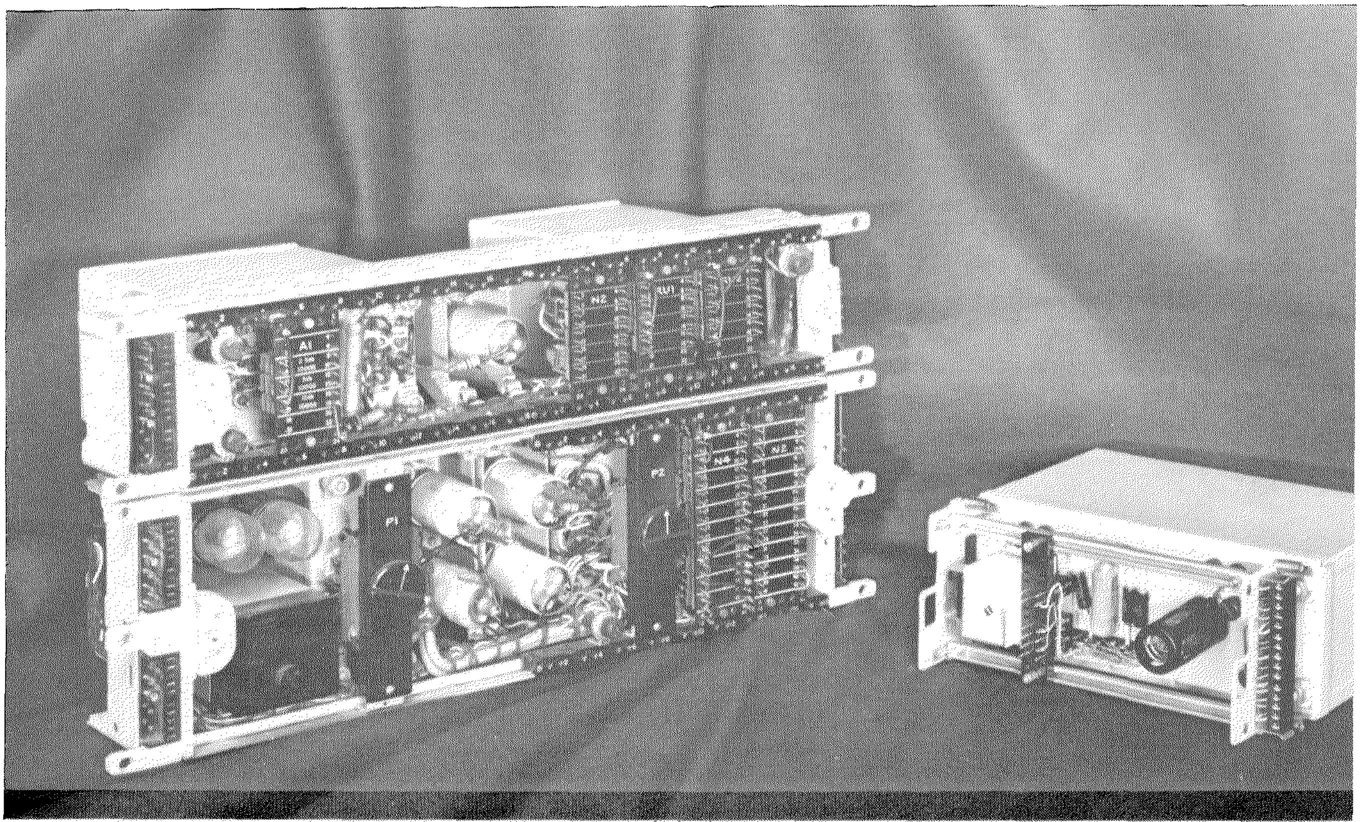


Figure 5—Equipment for *TF1* frequency-shift-modulated voice-frequency telegraph for one duplex channel end. At the left the lower unit is the detector and the upper panel accommodates the oscillator and modulator. At the right is the equivalent equipment in the *TF3* design.

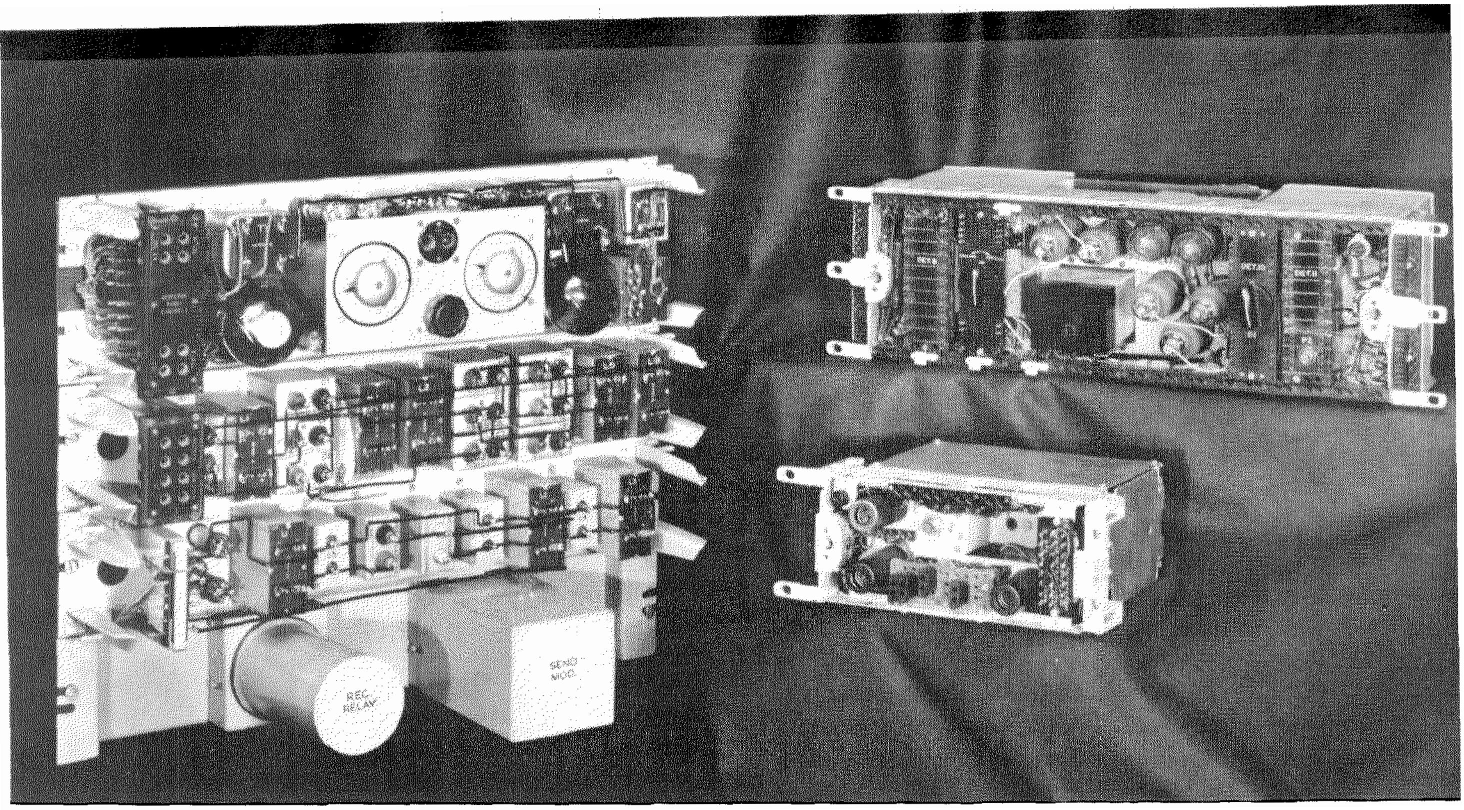


Figure 6—Comparison of three designs of amplitude-modulated equipment for one duplex channel end of a voice-frequency telegraph system.

24 are also derived by a similar group-modulation process; the reason is to ensure the high standard of performance of the equipment over its full temperature range. If produced by direct modulation the stability of the resonant units of channels 19 to 24 would approach their permissible limit. It has therefore been considered advisable not to use direct modulation of these channels for the main applications of this system, but only for special cases where a slightly lower standard of performance can be accepted.

The high standard of performance achieved by the *TF1* system has been maintained in the development of the *TF3* system, and there are considerable advantages with this latest system. Because transistors have been used to replace hard valves the power consumption has been reduced to only 1 watt for a single channel. The use of low-voltage supplies, together with the small powers required in the transistor circuits, enables smaller components generally to be used.

2.4 Mechanical Features

Modern techniques in the mechanical construction of panels, rack sides, and components in conjunction with improved circuits have enabled the cost and size of voice-frequency telegraph systems to be reduced by a most marked extent. A comparison between various amplitude-modulated equipments is shown in Figure 6. The equipment to the left was required for one channel end incorporating a static modulator, send filter, receive filter, detector, and receive relay for a system[7] produced about 1936. Such a system of 18 channels required 5 bays occupying a volume of 10 feet, 6 inches (3·2 metres) by 1 foot, 8·5 inches (0·52 metre) by 1 foot, 3 inches (0·38 metre). On one of these bays was mounted a multifrequency motor–generator, from which were obtained the various channel tones, and which could serve up to 180 channels if required.

The equipment in the upper right corner of Figure 6 is again a complete channel end. In this case an oscillator is provided on the panel to generate the particular channel tone, though the tone could be derived, as previously, from a multifrequency generator. This type of equipment is known as *TA2*, and was designed about 1949. It has become a standard system of the British Post Office, to whom it is known as Type IV equipment. Twenty-four such channels can be mounted on two rack sides of design known as new equipment practice (NEP).[10,11] These two rack sides when fitted back to back occupy the same space as one double-sided bay of the older design. The small panel to the lower right of Figure 6 is a complete channel end of a *TA4* system; 24 such channels mount on only a single new-equipment-practice rack side occupying a volume of 9 feet (2·74 metre) by 1 foot, 8·5 inches (0·52 metre) by 8·5 inches (0·22 metre).

The general reduction in size of these equipments was brought about by the availability of smaller coils, capacitors, and valves, and improved metallic rectifiers. The mechanical technique at the time of the design of the *TA2* equipment was to leave coils and capacitors individually unprotected from atmospheric conditions, but to put them, together with other components requiring protection, inside common hermetically sealed cans. These cans with their associated valves and other components were mounted on panel frameworks designed to slide into die-cast guides mounted on a light rigid steel rack side framework. The further reduction in size of the *TA4* equipment was by simplified and improved circuit design together with a new mechanical design that employed even smaller coils and capacitors, with resonant units individually sealed in cans of simple but efficient design. The type of valves used helped the general reduction; type *B7G* replaced the octal valves. The simplified circuitry required only three valves, one of these being that of the channel oscillator.

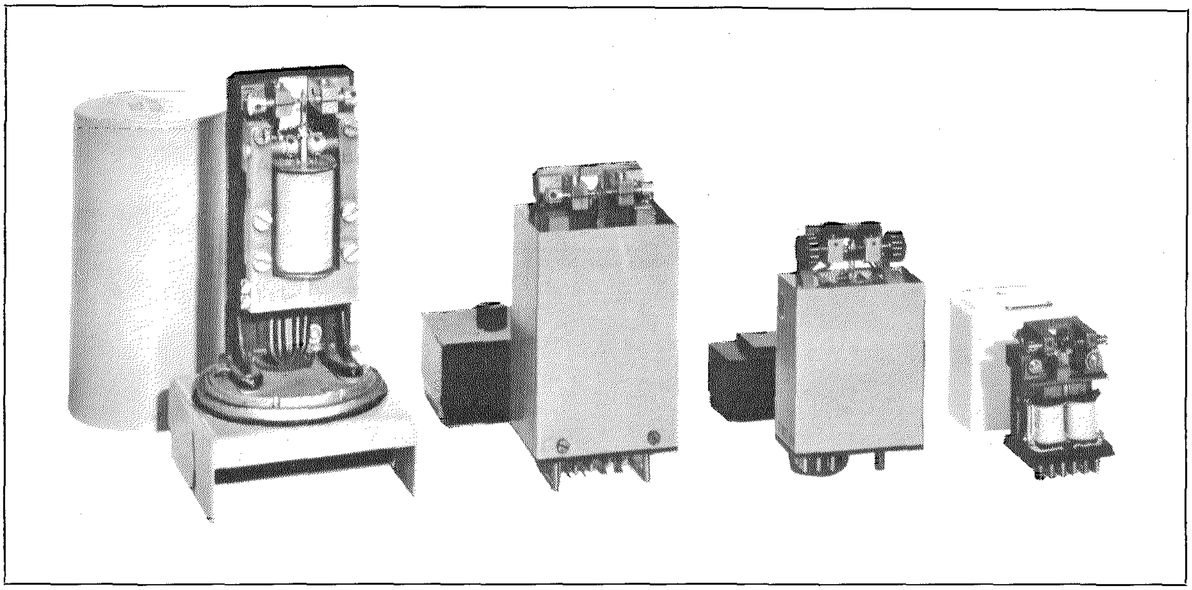


Figure 7—Four progressive designs of telegraph relays.

Frequency-shift-modulated equipments are compared in Figure 5, and just as marked reduction is shown as for the amplitude-modulated equipments. The two panels to the left of the picture constitute one channel end of the *TF1* system. These panels occupy five units of new-equipment-practice rack side, whilst the panel to to the right is for one channel end of the *TF3* system; it occupies only one unit of rack side. This reduction is not so much due to simplified circuit design as to the use of individually sealed resonant units and coils, to improved mechanical construction, and above all to the use of transistors in the place of hard valves.

2.5 Telegraph Relays

One important component in all telegraph systems is the telegraph relay, used to repeat the

[10] F. Fairley, R. J. M. Andrews, and A. C. Delamare, "Improved Equipment Practice Reduces Size of Telephone Transmission Systems," *Electrical Communication*, volume 27, pages 21–38; March, 1950.

[11] E. T. C. Harris and C. J. Spratt, "Improved Form of Mechanical Construction for Transmission Equipment—51-Type Construction," *Post Office Electrical Engineers Journal*, volume 51, pages 197–201; October, 1958.

detected alternating-current signals and to isolate the detector and amplifier circuits from the direct-current receive leg circuits. These relays also have progressed, as is shown in Figure 7. The order from left to right is the *4121* relay used on amplitude-modulated systems up till 1949, the *4148* relay for the *TA2* and *TF1* systems, the *4192* relay on *TA4*, and lastly the *4199* relay for the *TF3* equipment.

It is doubtful whether a further reduction in size of telegraph relays below that of the *4199* will be of much use in practice, since there is always the maintenance problem to be considered. If the physical dimensions of the relay become too small, it will require a highly skilled mechanic to maintain the relay, unless of course a sufficiently robust device is developed that needs negligible maintenance and is sufficiently cheap that at the end of its useful life it, or at least the worn part, could be discarded and easily replaced by a new one.

A modern competitor for the replacement of the electromechanical relay is, of course, the transistor. At the present stage of development and cost, these devices are not really a commercial proposition as a direct replacement for the conventional telegraph relay for the majority of its applications. However, the time can be foreseen when this replacement will occur.

3. *Miscellaneous Systems and Equipment*

3.1 High-Frequency Radio Telegraphy

Multichannel voice-frequency telegraph equipments of the *TF1* type have for some time been used on high-frequency radio circuits. To combat the effects of selective fading special diversity combining panels were provided, such that two telegraph channels, each carrying the same information but received either in space diversity or frequency diversity, could be combined to give one common output. Though these equipments have given satisfactory service, it has been found desirable for the frequency deviation of each telegraph channel to be greater than the existing ±30 cycles per second. Systems have now been developed from the *TF3* equipment with 340- and 170-cycle-per-second spacing of channels, the deviations being ±85 and ±42·5 cycles per second respectively.

The design of the diversity combining equipment is based on the principle of square-law combination.[12] It can be shown theoretically and practically that if two or more channels carrying the same telegraph information are combined in such a way that each contributes to a common output in proportion to the square of its relative received amplitude, then a distinct signal-to-noise advantage can be obtained. The 340-cycle-per-second system using dual diversity is recommended for start–stop telegraph signals at speeds up to 75 bauds and for synchronous transmission up to 100 bauds.

3.2 Submarine Telegraphy

There has been a rapid increase in the number of submarine-cable telephone systems in recent years, and this has resulted in a large increase in the number of international voice-frequency telegraph connections. The voice-frequency telegraphs so used are of the standard land-line type, and a notable case is the transatlantic telephone cable between Great Britain and America. One of the speech circuits to Canada is used continuously for 24 channels of frequency-shift-modulated voice-frequency telegraph equipment of the *TF1* design. These channels were planned initially to operate at 50 bauds and possibly 75 bauds; in fact, some are being used at speeds in excess of 80 bauds.

The vast majority of submarine telegraph circuits still use, and will continue to do so for a very long time, direct-current transmission. The efficiency of existing circuits has been improved by the insertion of a submersible amplifier[13] or, more accurately, part of the receiving equipment, at a point in the sea sufficiently remote from interference picked up in the shallow shore end of the cable. The improvement thus obtainable, although reduced by the fact that a repeatered cable may no longer be operated duplex, is appreciable, and it is customary to introduce such a repeater at each end of the cable with switched by-pass arrangements to permit operation in either direction according to traffic requirements.

[12] H. T. Prior, British Patent 709 793; June 17, 1952.
[13] C. H. Cramer, "Submerged Repeaters for Long Submarine Telegraph Cables," *Western Union Technical Review*, volume 5, pages 81–91; July, 1951.

3.3 High-Speed Data Transmission

As the demand for high-speed data transmission grows it is expected that these basic designs of frequency-shift-modulated systems will be extrapolated to provide any reasonable speed of transmission required over normal telephone circuits, whether the mode of transmission required is a parallel or series mode. Such a system has been developed purely for experimental purposes; this was designed to transmit a message at an aggregate speed of 1750 bauds in a parallel mode over seven 250-baud voice-frequency channels, each of 400-cycle-per-second bandwidth. Its application was of a particular type, and built into the telegraph system were the master speed controls and means to maintain synchronism from the actual message signals being transmitted between the two terminals.

3.4 Start–Stop Regenerative Repeaters

Despite the more widespread use of frequency-shift-modulated telegraph systems with their improved performance, it is still often necessary, to enable long telegraph circuits made up of several links in tandem to function satisfactorily, for the signals to be regenerated and repeated at some point in the circuit. The necessity for regenerators is even more marked on radio-telegraph circuits. Such a regenerator[14] has been developed using transistors and is illustrated in Figure 8. This is an interesting example of the practical application of transistors to digital circuits, and shows how a large amount of this type of circuitry can be mounted within a small space. No coils have been used in the design of this regenerator, the basic bricks being transistor binary circuits together with rectifier gating circuits. The regenerator with all the features found necessary for radio working; that is, adjustable false-start rejection, automatic insertion of a missing stop element, and accurate retransmission of a long space condition, is mounted within a unit 7 inches (18 centimetres) by 3·5 inches (9 centimetres) by 8 inches (20 centimetres). Employing one control oscillator for each 12 regenerators, 36 regenerators and their power supplies can be mounted on a 9-foot (2·74-metre) new-equipment-practice rack side.

[14] W. F. S. Chittleburgh et al, British Patent Application 20 160/56; June 29, 1956.

4. Conclusions

The demand for telegraph transmission and the development of such equipment have, in the past, been continuous. Although telegraphic traffic as such has tended to decline in recent times, a new demand has grown up. This is data transmission, which needs the same, or similar equipment. Signalling speeds are being requested that require for transmission only the narrow bandwidth of a standard voice-frequency telegraph channel or that may demand the full facilities of a complete telephone channel. These higher speeds call for new systems but, in most cases, the basic design problem is similar to that for voice-frequency telegraphy. The fundamentals of design that were evolved in the past, can be and are now being applied to these new data-transmission systems.

5. Acknowledgement

Thanks are extended to my many colleagues who have helped in the preparation of this article.

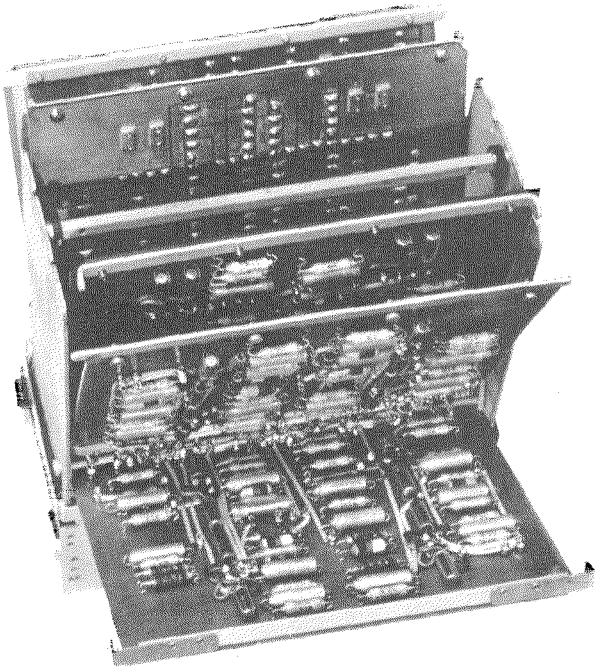


Figure 8—Start–stop-telegraph regenerative repeater using transistors.

Page Printer LO-15-B

By J. AUGUSTIN

Standard Elektrik Lorenz AG; Stuttgart, Germany

ACCEPTANCE of the model-*LO-15* page printer in Germany is attested by the fact that one of every two teleprinters in that country is of this type. Recently, improvements have been made in its design and are incorporated in the *LO-15-B*. The new teleprinter provides key-type controls for the answer-back unit, an all-blank repeats operation, a transmitter–distributor, and a reperforator. In addition, the operating noise has been substantially reduced by minimizing all openings in the housing through which sound generated in the interior of the case might escape.

1. General

The new printer is equipped with a monitor that automatically stops its operation if there is no paper left on the roll. It will resume operation only if paper has been inserted. This prevents loss of messages that could result with unattended operation in answering a call at a time when the machine was out of paper.

The direct-current power needed for operation is provided from the alternating-current mains, which may deliver 110, 127, or 220 volts. The power unit is mounted within the teleprinter housing.

The page printer may be installed in the conventional wooden console or in a new table-type metallic housing. A built-in fluorescent lamp provides uniform glare-free illumination of the printed material. Provision is made to accommodate attachments such as a reperforator or a transmitter–distributor within the hood. Special

Figure 1—*LO*-15-*B* page printer with built-in reperforator and tape transmitter mounted on a desk.

message blanks in folded strip form are provided for by guides to direct the paper to and from the platen and a base designed to collect the folded strip of paper.

2. Transmitter–Distributor

The new transmitter–distributor is designated the *LS-424B*. The perforated tape may be inserted from one end or at any other point along the tape. For convenience, the tape travels parallel to the front edge of the unit and the small plate that holds the tape in place may be released by pressing a button. The transmitter–distributor disconnects itself automatically if the tape is not in operating position. It also stops if the connected distant subscriber starts to transmit.

The transmitter–distributor is stopped whenever the mode of operation is changed. If tape is being run through the transmitter to produce a printed copy and a call is received, the stoppage due to the change in mode prevents the taped material from being transmitted to the calling subscriber.

Control keys are provided to operate or stop the transmitter and the perforator; the latter is electrically connected to the page printer by plugs and jacks.

3. Reperforator

The reperforator *ELO-514* has also been equipped to perform some additional functions. It can be switched on or off by the distant subscriber through the transmission of certain code combinations. When the tape is about to run out, both optical and acoustical warning signals are given.

A tape monitor has been added that will disconnect the reperforator if it has no tape. If a call is received, the teleprinter is switched to reception without reperforation. If the caller insists on reperforation by transmitting the designated code combination, the transmitter will be switched off entirely to indicate to the caller that there is no tape in the reperforator.

The reperforator is switched on and off by control keys. Special keys are also provided to eject the tape and to retract the tape to erase an error.

In other details and ratings, the *LO-15B* is the same as the previous model.

Figure 2—Page printer mounted on a pedestal showing magazine for handling message blanks in folded strip form mounted within the pedestal.

High-Speed Tape Perforator SL614

BY J. AUGUSTIN

Standard Elektrik Lorenz AG; Stuttgart, Germany

WHEN PERFORATED tape is used for permanent storage in data-processing systems, bookkeeping machines, and computers, the perforator used as an output device must have a much-higher speed than the conventional perforators used in teleprinter service.

High-speed perforator *SL614* (Figure 1) has been constructed for such high speeds, of which the following are presently available: 25, 31.25, 40, and 50 characters per second.

Tape can be perforated in any of the 5- to 8-unit codes. All varieties of this machine can be equipped to perforate two identical tapes if required. (A special design for perforation of card edges or McBee cards is under development.)

The high-speed perforator can be adapted to any of the above modes of operation at any time by simply changing the gears and control-magnet systems.

The main units of the machine are: the motor and drive, a set of contacts that generate orders and synchronizing pulses for the associated data-processing system, the tape reels with signaling contacts, the perforating mechanism with the tape driving system, the control magnets,

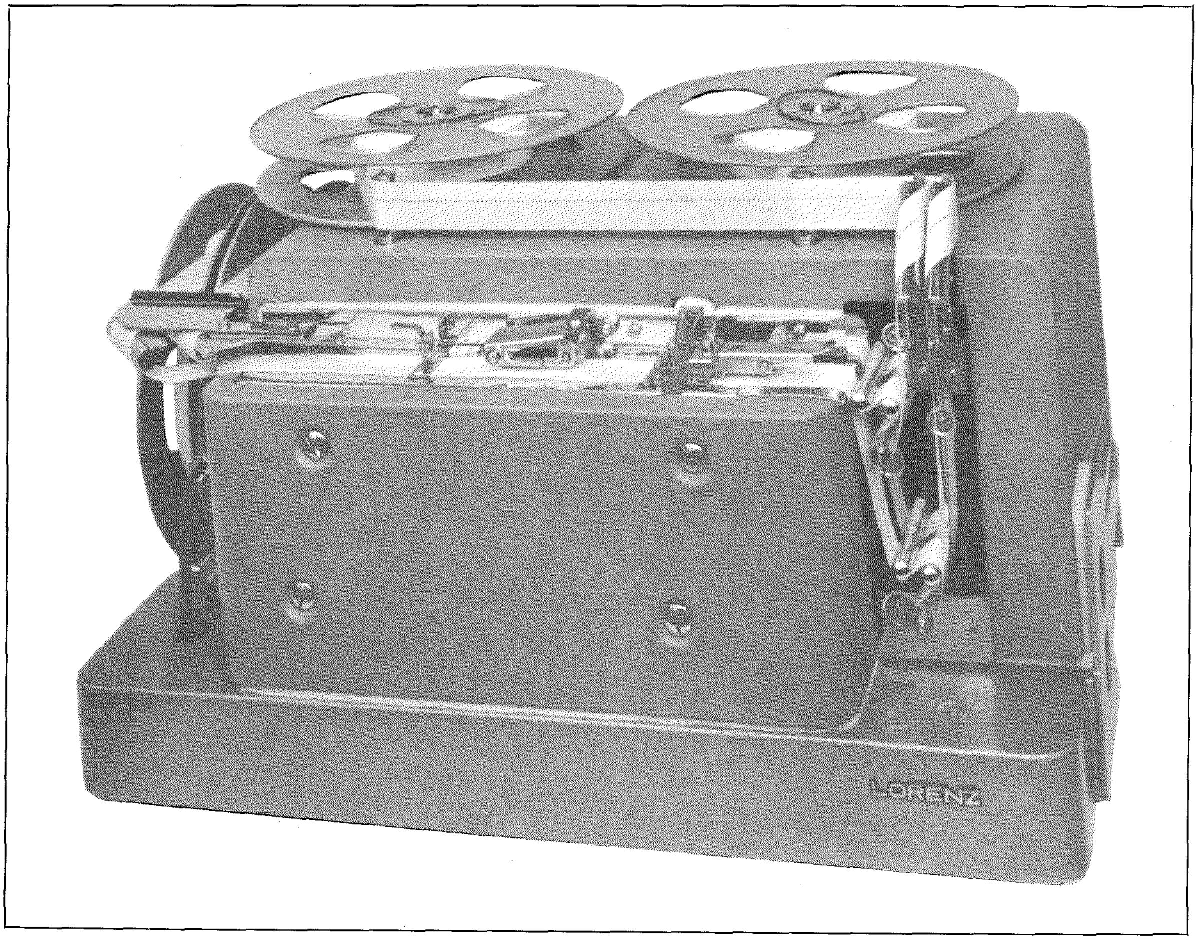


Figure 1—Front view of high-speed tape perforator.

the coupling magnet system, the revertive-signal contacts, the take-up reels, and the control relays. An additional cam-switching unit can also be mounted.

The standard equipment of the *SL614* comprises a universal motor for 220 volts at 50 cycles per second; the speed is regulated by a governor. If required, motors of other types (shunt-wound direct-current or asynchronous or synchronous motors for 50 or 60 cycles per second) can be provided. The power consumption of the drive is 200 volt-amperes. Rated speed is reached within 2 seconds after the motor is started.

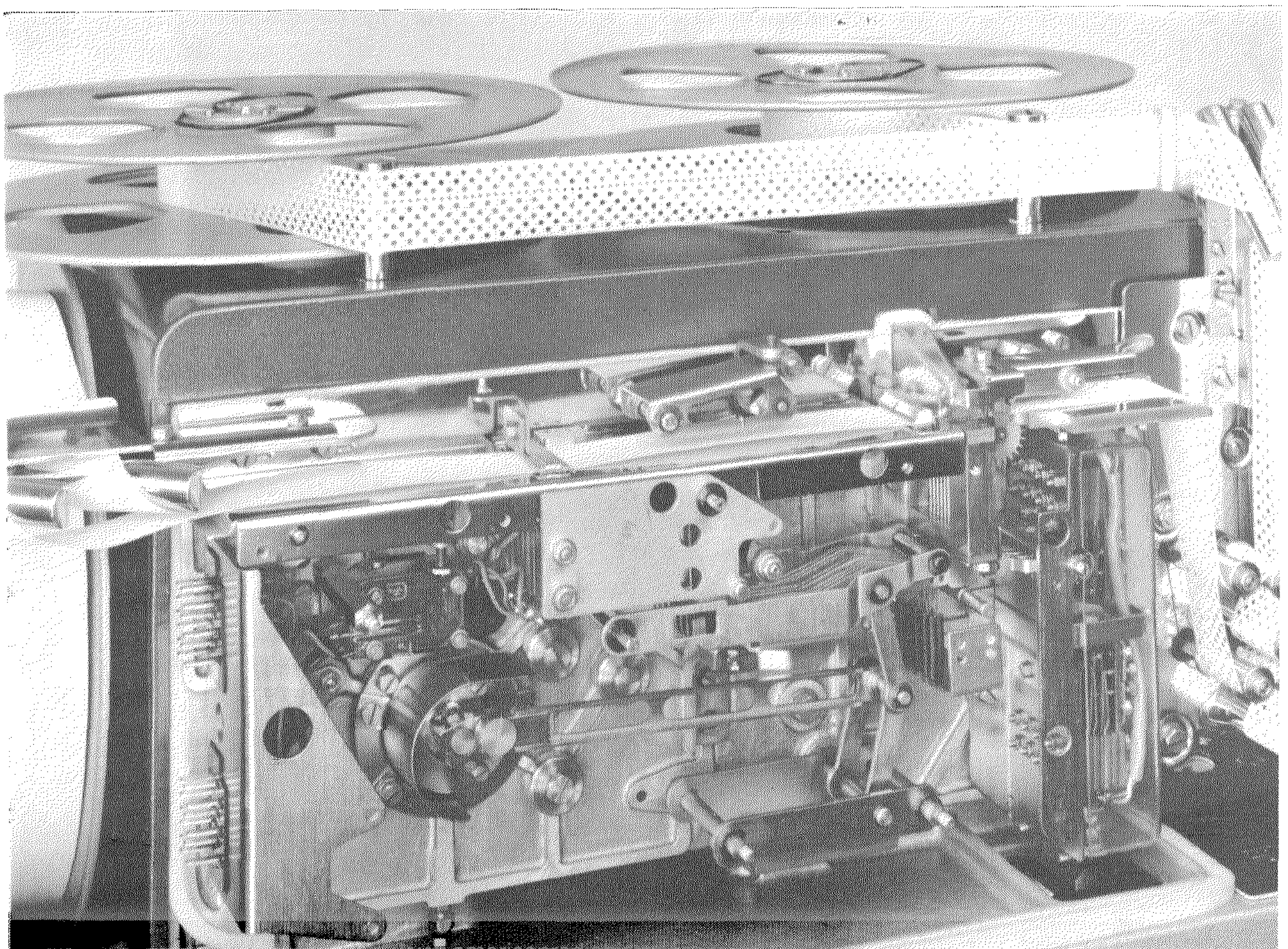
There are 7 control contacts mounted on the main-shaft cam disks. The first contact controls the coupling processes within the machine and the other 6 contacts generate order signals for the associated data-processing system. Cam disks for these contacts are supplied according to the particular application. The contacts are designed for bounce-free operation and an accuracy of ±1 millisecond. If more than 6 contacts are needed, the cam-switching unit mentioned above is added.

The left side of the perforator is equipped with journals holding the two reels of nonperforated tape. Each reel is monitored by a feeler lever operating a contact to actuate an alarm to warn the operator when the reel is nearly empty.

The tape is led around the corner to the tape guide, the tape-checking lever, perforating head, and to the tape driving system. The tape-checking lever operates a contact signaling whether or not tape is moving properly. This signal can control the output circuit of the data-processing system so that there is no output if tape is not present for perforation. The tape drive pulls the tape from the reel at the proper rate through the perforating unit.

At the output of the perforating unit, tape is pulled character by character by a sprocket wheel to go then to the take-up reels. Mounted on the same base plate is the coupling unit (Figure 2) connecting the perforating unit and the tape drive to the motor during the perforating process. The perforating mechanism comprises a gearbox of bell-crank levers operating the individual perforating pins. This system is controlled by intervening armatures of the control magnets so

Figure 2—High-speed tape perforator with cover removed.



that each perforating pin will operate only when its associated control magnet is energized. The feeding-hole pin operates at every step. At the moment of perforation, the revertive-signal contacts operate; they are interlocked with their associated perforating pins.

After passing through the perforating unit and pin-wheel drive, the perforated tapes are taken up on the reels on top of the machine. These reels are driven by the motor through separate friction clutches, ensuring tight winding of the tapes. Pins between the tape drive and the takeup reels guide the tape and act as brakes inasmuch as the pin wheel is not loaded by the pull of the reels.

Relays facilitate control of the perforator by the associated data-processing system. One relay switches the perforator on; the other operates when the tape runs out. Two relays are interlocked so that tape replacement is possible only when the machine is switched off and accidental depression of the control keys does not interrupt perforation.

The individual sub-units of the perforator are interconnected by plugs and jacks for convenient replacement and servicing. An interlock circuit ensures that the machine will not operate if any of the plugs are disconnected. The paper chaff box mounted at the right-hand side can contain the chaff from two perforated tape rolls.

The high speed of the *SL614* required special design for some units. Thus, the armatures of the control magnets are retained mechanically and either released or held magnetically, depending on the state of current. Moreover, the time constants are as small as possible; this necessitates higher power consumption in the control circuits.

Materials of best choice enable operation for long periods with minimum maintenance. Therefore, the life of the perforator depends greatly on the frequency of on-and-off switching of those parts involved in perforating. Hence, discontinuous modes of operation, where the output is individual characters, should be avoided. Operation should preferably be stopped only when the entire message is finished. Use of unsuitable paper can result in premature wear of perforating pins and dies.

The motor is conventional; its leads are equipped with fuses and are free of electrical interference. The current and voltage depend on local conditions.

All control terminals are brought to contacts on the rear side of the machine; from where connections can be made to the associated data-processing system. The high-speed perforator has no telegraph-current sources of its own.

The machine is 20.5 inches (520 millimeters) wide, 12.6 inches (320 millimeters) high, and 14.4 inches (365 millimeters) deep; it weighs 44 pounds (20 kilograms). Available speeds are 25, 31.25, 40, and 50 characters per second in codes from 5 to 8 units. The power requirement is 200 volt–amperes and suitable equipment can be provided to operate at the voltage and frequency of the power mains in the locality of use. The direct-current control power that must be supplied by the associated data-processing system is 60 volts at 1.2 amperes maximum.

Teleprinter for Reliable Transmission of Numbers*

By J. AUGUSTIN

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PRIVATE and public administrations often transmit numerical teleprinter information. When straight text is transmitted, an error consisting of a wrong letter is most often quickly apparent and the correct letter can be deduced from the context. It is quite different with numbers. The receiving subscriber usually has no means of perceiving erroneous numbers or replacing them by the correct numbers. Errors can be caught by complete retransmission of each individual message; visual inspection of the two versions of the message will reveal any errors. However, this method is rather costly, especially where much information is handled.

There was good reason, then, to develop a teleprinter for reliable transmission of numbers. Compatibility of such a machine with the public telex systems is of particular importance; in the majority of cases, maintenance of a private network is too expensive.

The machine developed has three shift positions instead of the conventional two. The first two shifts comprise the usual lower- and upper-case characters of any ordinary teleprinter. The third shift provides reliable transmission of 15 characters: The numbers 0 through 9, plus sign, minus sign, space, carriage return, and line feed.

Of the 32 possible combinations in the 5-unit code, only those 15 were selected that are characterized by 1 and by 3 marks in the 5 positions. Hence, at least 2 elements of a combination have to be wrong to cause an error. In principle, combinations having 2 and 4 marks in the 5 positions could also have been selected. However, the combinations for carriage return, space, and line feed also must be used in the first and second shifts. In telegraph systems operated according to Comité Consultatif International Télégraphique et Téléphonique recommendations, these three combinations are of the group having 1 mark in the 5 positions; this decided the case in favor of the 1-mark and 3-mark combinations.

Figure 1 shows a teleprinter providing reliable

Figure 1—Teleprinter for reliable transmission of numbers. The set of black keys used in the third-shift position is visible above the conventional keyboard.

* Originally published under the title "Fernschreiber zur gesicherten Ubertragung von Ziffern" in *SEG-Nachrichten*, volume 6, number 1, pages 58–59; 1958.

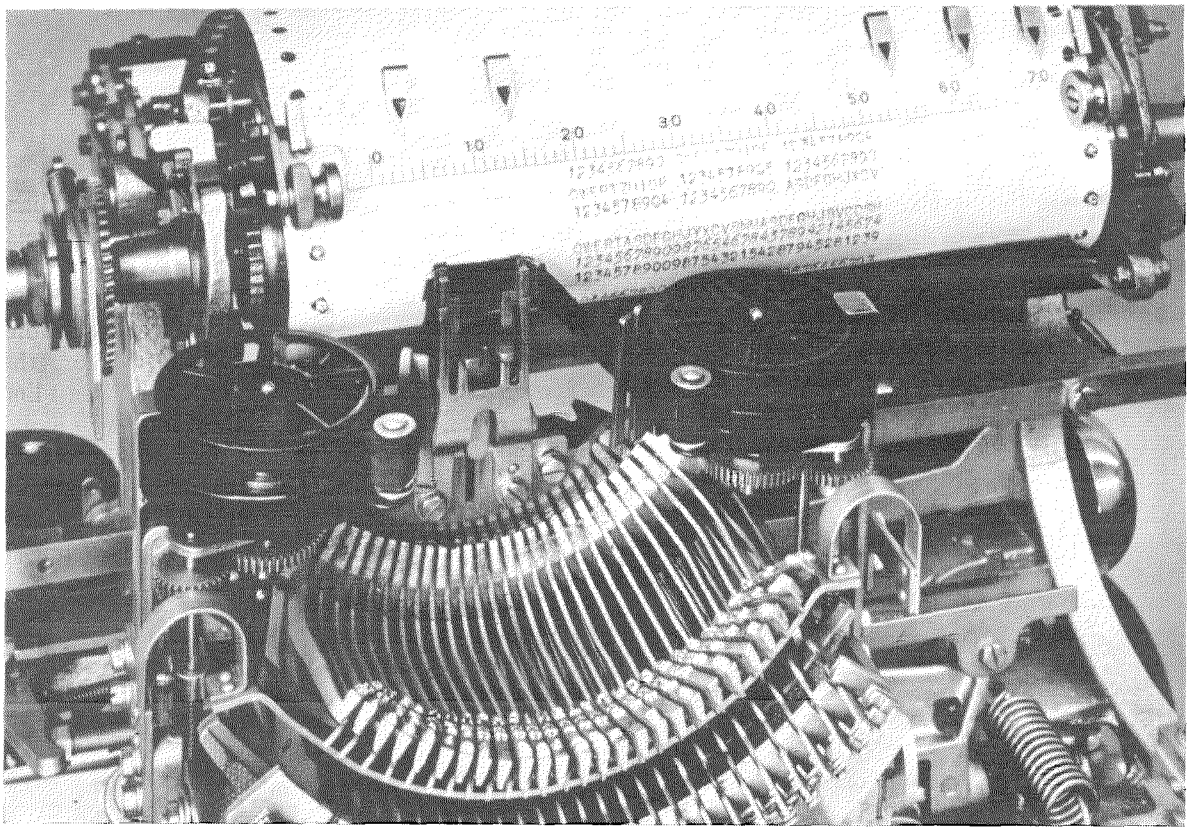


Figure 2—View of the type bars with three characters each and the teleprinted message in which the italics denote the high-reliability numbers.

transmission of numbers. The conventional keyboard has white keys. Above it are another 12 back keys for 0 to 9, plus, and minus. A locking bar releases the white keys only in the first- and second-shift positions; the black keys only in the third-shift position. Accidental operation of the wrong keyboard is thus precluded.

Each of the type bars carries three characters. (Figure 2). The position of the platen in any of the three shift positions determines which of the three characters of a type bar will print.

The numbers transmitted in the third shift are clearly distinguished from the conventional numbers by printing in italics. All type bars not equipped with the third-shift characters have an asterisk in the third-shift position. This asterisk signals that an erroneous number-combination has been received; as a result, no special electrical check measures are necessary.

If its only current impulse is dropped, each of the 1-mark-in-5-position combinations becomes combination *32* of International Telegraph Alphabet *2*; that is, the combination releasing the third shift. Since the receiver could not determine whether this code is transmitted on purpose or is just a faulty combination, a locking device has been attached to release the third-shift key only when the carriage position is at the beginning of a new line. In addition, a special type bar has been provided to print an asterisk each time the combination for the third shift is received (except at the start-of-line position). Thus, any faulty character converted into combination *32* can be identified.

The above arrangements offer the same protection against mistakes as message repetition in a conventional teleprinter circuit; however, they save the considerable cost that would be incurred by the doubled transmission time.

As mentioned before, this teleprinter operates in the conventional manner when in the first or the second shift and can be employed to communicate with every subscriber in a public telex network.

Teleprinter Synchronizing Set SYZ-634

By W. SCHIEBELER

Standard Elektrik Lorenz AG; Stuttgart, Germany

START–STOP teleprinters of any description may utilize the *SYZ-634* attachment. It is designed to bridge a time of disturbed transmission by supplying locally generated start and stop signals to the receiving teleprinter, thus maintaining the rhythm of the teleprinter transmission. It is therefore particularly useful for radio transmission and for the transmission of enciphered messages. Its application covers the whole field of automatic transmission in which the transmission speed is normally uniform.

There are two reasons for restoring start–stop pulses lost by disturbed transmission paths. One of these concerns the case of plain-language communication in which loss or distortion of one or more teleprinter signals and the associated start–stop pulses causes not only incorrect reproduction of that particular character but may also distort subsequent characters that were correctly transmitted.

If the receiving page printer fails to receive one or more stop pulses, the selector mechanism will run through and will not be synchronized until several characters later. This action is prevented by the use of the synchronizing set *SYZ-634*, by means of which the first character correctly transmitted after an interruption is also reproduced correctly. As a result, the number of faulty characters printed in conventional operation is considerably reduced by the *SYZ-634*.

The other case is that of enciphered messages in which the consequences of start–stop pulses being lost in transmission and the receiving deciphering set losing synchronization with the transmitting encoding set are even more serious than in plain-language communication. This is because the enciphering at the transmitter and deciphering at the receiver are on a character-by-character basis and a character lost in transmission throws all following characters out of proper deciphering order. The rest of the message, therefore, becomes illegible even though the following transmitted characters are correctly received.

The synchronizing set is designed to bridge interruptions of at least 3 seconds caused by distortion; this corresponds to 20 characters. It is adequate for the most-unfavorable case of such a disturbance. Bridging times of 20 seconds and more can be achieved for complete interruption of the transmission path.

The most important task of the synchronizing set is to generate start and stop pulses continuously at the correct frequency and phase established by the transmitter. In the synchronizing set, the received start and stop pulses are diverted to a pulse generator and the newly developed pulses are inserted into the received signal in the correct phase to actuate the teleprinter. As long as undistorted characters are received, the pulse generator is frequency controlled continuously by the mark-to-space transitions of the stop pulses. If these transitions are not received due to a disturbance in transmission, pulses continue to be generated by the local pulse generator at the last-received frequency, thus ensuring uninterrupted operation of the associated teleprinter receiver.

As will be seen in Figure 1, the teleprinter characters arriving from the line are applied to a line directional switch that removes the combined stop and start pulses and sends only the character pulses through the teleprinter directional switch to the teleprinter.

The removed stop pulses are differentiated, the negative pulses produced by the leading edges are suppressed and the positive pulses from the trailing edges go to a trigger circuit to control the frequency and phase of new pulses produced by a pulse generator under its control. From the generator, the pulses pass through the teleprinter directional switch to be combined with the character pulses received directly from the line directional switch, the combined signals actuating the teleprinter.

If the teleprinter signals being received from the line are interrupted, no positive pulses will be received by the trigger from the differentiator and line directional switch. Nevertheless, the pulse generator will continue to supply stop and start pulses through the teleprinter directional switch to the teleprinter. The latter will receive code groups consisting of all marks or all spaces depending on the condition of the line and which corresponds to "letter shift" or "ff". However, in the case of enciphered messages, for instance, the tapes continue to move synchronously at transmitter and receiver so that, when the disturbance disappears, correct text is again printed.

The principal unit of the synchronizing set is a free-running multivibrator, the circuit of which is shown in Figure 2.

Contrary to conventional multivibrators employing a double triode, two double triodes are used. One pair of triodes *V1* has the function of a conventional multivibrator. Cutoff pulses are alternately applied to the two grids so that one triode is always nonconducting while the other conducts heavily. After a certain time when the grid-circuit capacitors have been charged or discharged, respectively, the state is reversed so that the first triode conducts and the other does not.

The second pair of triodes *V2* controls the frequency of the multivibrator. A slightly positive or negative voltage is applied to the grids of *V2* which are connected together. This voltage also appears across capacitor *C1*. If this voltage is positive, the pulses are lengthened; if the voltage is negative, the pulses are shortened. At zero grid voltage, the pulse duration is 150 milliseconds, that is, it coincides with the duration of a teleprinter character.

The plate currents through *V2* are subject to pulse-shaped alternations similar to those passing through *V1*. However, the cutoff of triodes *V2* is not as complete and uniform as in *V1* but produces a sloping pulse crest. The plate-current pulse variations in *V2* are not controlled by the grids of these triodes, which are

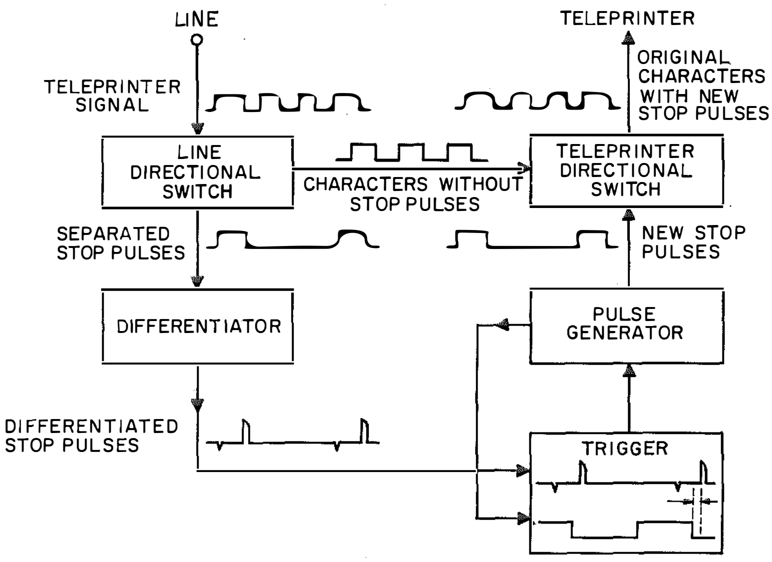


Figure 1—Operation of synchronizing set.

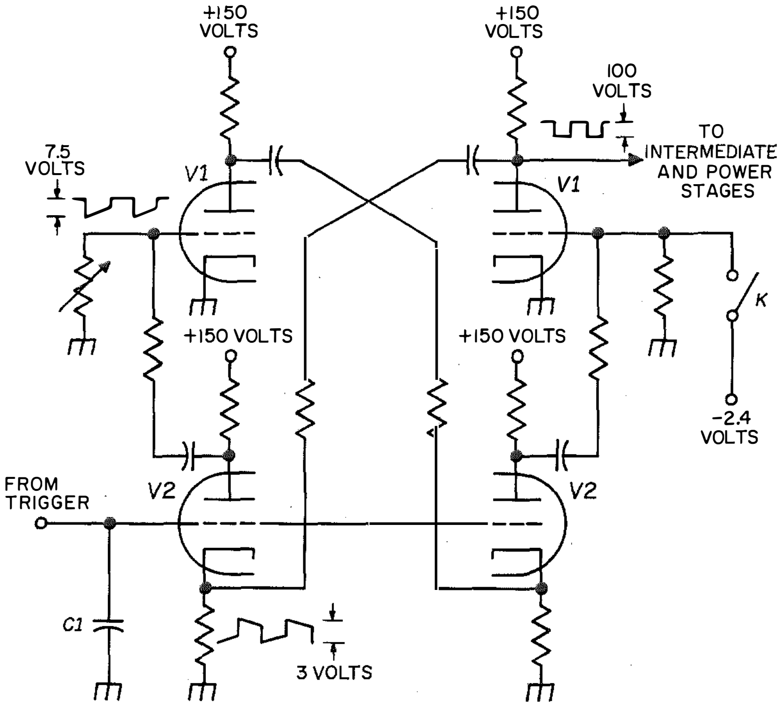


Figure 2—Multivibrator pulse generator. Signals from the trigger circuit to the double triode *V2* control the length of the pulses generated by the two triodes of *V1*. *C1* has a capacitance of 8 microfarads.

connected to the trigger circuit but by pulses obtained from the plates of *V1* and applied to the cathodes of *V2*.

The symmetric square-wave pulses from *V1* are applied to the grid of an amplifier employing a pentode connected as a triode to provide sufficient power to operate the receiving magnet of the teleprinter.

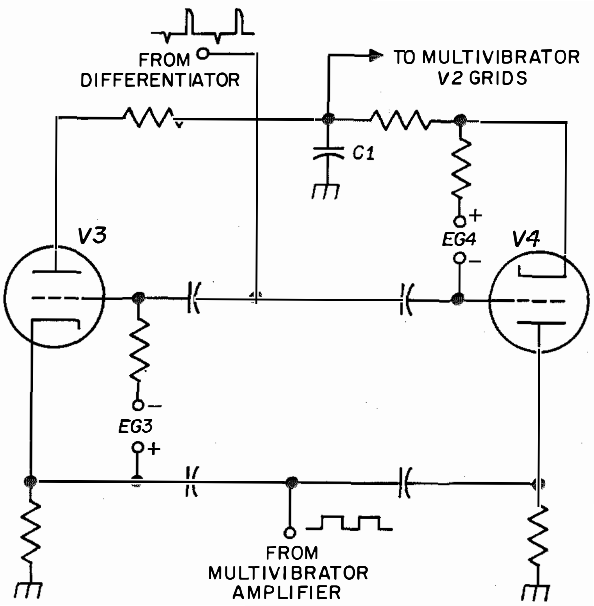


Figure 3—Trigger unit. The bias voltages *EG3* and *EG4* are adjustable.

In addition, the plate-circuit pulses of *V1* are applied through a buffer stage to the trigger unit. The buffer stage is provided to protect the multivibrator from overload.

The trigger unit employs the circuit shown in Figure 3. Its function is to compare continuously the positive control pulses obtained by differentiation from the trailing edges of the stop–start pulses of the received teleprinter code with the timing of the new stop–start pulses produced by the generator. If there is a difference in time between the trigger pulse and the new pulse, the grid voltage of the multivibrator tubes *V2* must be increased or decreased, depending on the sign of the difference. This ensures synchronization between the distant tape transmitter and the multivibrator.

The mode of operation of the trigger unit is as follows. Two triodes, *V3* and *V4*, are connected to capacitor *C1* of the multivibrator tube *V2* in Figure 2. The square-wave pulses from the multivibrator are amplified in an intermediate stage and applied as plate voltage to *V3* and *V4*. When *V3* and *V4* have no negative grid bias, that is, are not operating at plate-current cutoff, the positive half-wave of the pulse can flow only through *V4* and the negative only through *V3*. In this way, *C1* alternately receives positive and negative charges at the pulse-repetition frequency. The multivibrator connected to *C1* then operates somewhat slower during the positive half wave and somewhat quicker during the negative half wave of the pulse.

Generally, however, *V3* and *V4* are blocked against any flow of current by high grid bias voltages. Capacitor *C1* receives no charge and retains its voltage. Only when the positive pulses from the differentiator are applied to the grids of *V3* and *V4* do both tubes conduct for a short time, during which the charge across *C1* can vary. The sign of the charge variation depends on the phase of the multivibrator pulse instantaneously applied as plate voltage to *V3* and *V4*. If it is positive, a current pulse will flow through *V4* and charge *C1* positively. If the phase is negative, the current will surge through *V3* and produce a negative charge on this capacitor. The multivibrator reacts to this charge variation by a corresponding frequency variation.

The frequency of the multivibrator pulses and of the corresponding pulses of the teleprinter signals differ very little from each other. Hence, several sequential pulses from the multivibrator usually occur so as to overlap completely the short controlling pulses from the integrator. This causes the multivibrator to reduce the length of the pulses and increase the length of the spaces, corresponding to a higher frequency. This process is indicated in Figure 4. Through this action, the multivibrator pulses are gradually changed in their duration until their edges coincide with the narrow positive pulses, whose spacing is assumed to be constant. As soon as the positive pulses fall into the reverse phase of the multivibrator

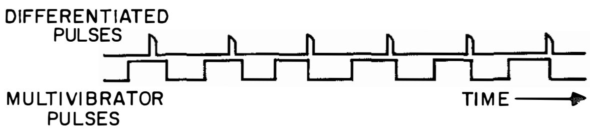


Figure 4—Synchronization of multivibrator pulses.

pulses, the controlling action starts in the opposite direction. After a moderate overshoot, the pulse edge is again pulled towards the positive spike and clamped. The multivibrator pulses show so small a variation about the correct timing set by the short pulses from the integrator that this is not operationally significant.

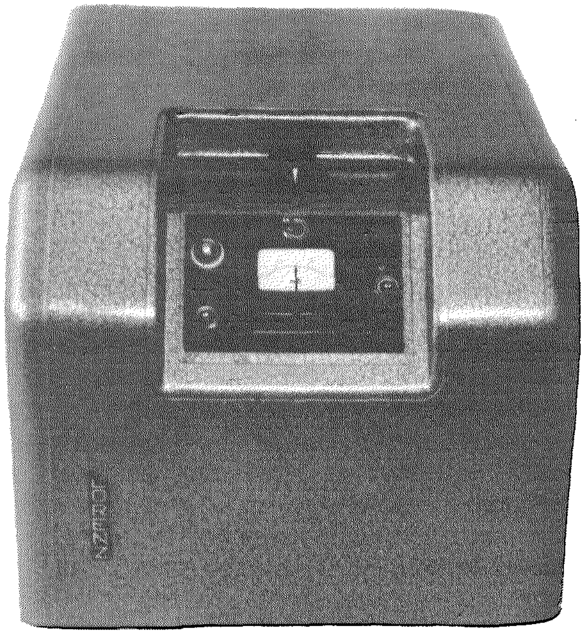


Figure 5—Synchronizing set in housing.

If the teleprinter characters and, hence, the positive spikes do not arrive at the trigger unit, the plate current of *V3* and *V4* is cut off by the negative grid bias; the charge on capacitor *C1* is no longer varied, and the multivibrator oscillates at the last-adjusted frequency. When the teleprinter signals are received again, synchronization will be resumed provided the time difference between the positive integrator peaks and the multivibrator-pulse trailing edges has not exceeded +20 or −30 milliseconds. If it has, the control pulses will no longer be applied to the triggering circuit because the line directional switch will have already disconnected the line from the differentiating stage. In this case, no proper deciphering can be expected.

The two directional switches, one connected to the line and the other to the teleprinter, consist of two transfer contacts of a relay operated by a univibrator in the synchronizing set. Operation is such that, as soon as the stop pulse begins, the line is electrically connected to the differentiating unit and, at the same time, the generator is connected to the receiver magnet of the teleprinter. Towards the end of the start pulse, these connections are opened again and the line is connected to directly the teleprinter receiver magnet.

Contact *K* of the multivibrator, shown in Figure 2, is initially closed to place a negative bias on the grid of the output triode of *V1* and stop the multivibrator from oscillating. Hence, the multivibrator amplifier stage supplies continuous current to the receiver magnet of the teleprinter. As soon as the first start pulse arrives from the toll line, contact *K* is opened, the multivibrator begins its first oscillation, disconnects the continuous current to the receiver magnet by cutting off the final-stage tube, and allows the teleprinter to start operating.

The sychronizing set is shown in Figures 5 and 6. It has a meter on its front panel to indicate the voltage across *C1*. If the voltage differs substantially from zero, this is a sign that the fundamental frequency of the multivibrator deviates from the teleprinter frequency. The multivibrator frequency may be readjusted by a control knob on the front panel. The set can be switched to telegraph speeds of 50 and 45.5 bauds.

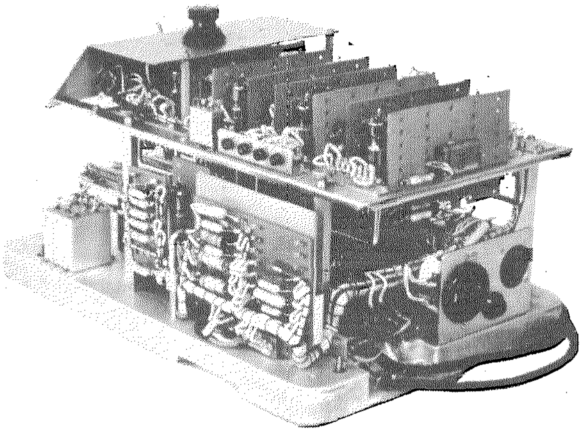


Figure 6—Synchronizing set with cover removed.

High-Speed Gas Tubes for Switching*

By A. H. BECK and T. M. JACKSON

Standard Telecommunication Laboratories Limited; London, England

DURING the past ten years a series of cold-cathode gas tubes has been developed specially for applications in telephone and telegraph switching and for use in computers. A very-brief account of the earlier tubes is given. Two recent advances described in more detail are: A tube designed to pass voice-frequency currents, which is easily incorporated in a switching matrix, and a high-speed triode trigger tube capable of operation at speeds up to at least 1 megacycle per second. The physical obstacles to further progress are discussed.

• • •

For the past decade an extensive programme of investigating the properties and uses of cold-cathode discharge valves in telephone and telegraph switching has been in progress. A short account of the major results obtained is given here.

All the tubes to be described have a common feature:—they use pure metallic cathodes of nickel or molybdenum. This permits the addition of a deionisation agent, usually hydrogen, which greatly increases the operating frequency range. For precise discharge-initiation timing, it is necessary to include a permanent electron source known as the primer. The high work function of the cathode materials precludes use of priming by external illumination and requires an internal priming source. Early experimental work established the principal mechanisms involved in the main discharge; (ionisation coupling, dynamic breakdown, and deionisation) upon which later tube developments were based. This work extended the study[1] of the Townsend discharge by use of modern pulse techniques. Based on these studies, the first counter tube made (1947) was of the progressive-glow-discharge type, based on the variation of ionisation coupling with spacing. This tube had two disadvantages, the multiple glow condition and the need for resetting the tube at the end of its cycle.

* Presented before the German Society of Telecommunication Engineers at Aachen, Germany; February, 1957.

[1] G. H. Hough, Thesis, University of London; 1950.

The next step was to eliminate the multiple-glow condition; other possibilities were examined. These are shown in Figure 1 and are all based on extinguishing a conducting gap by initiating

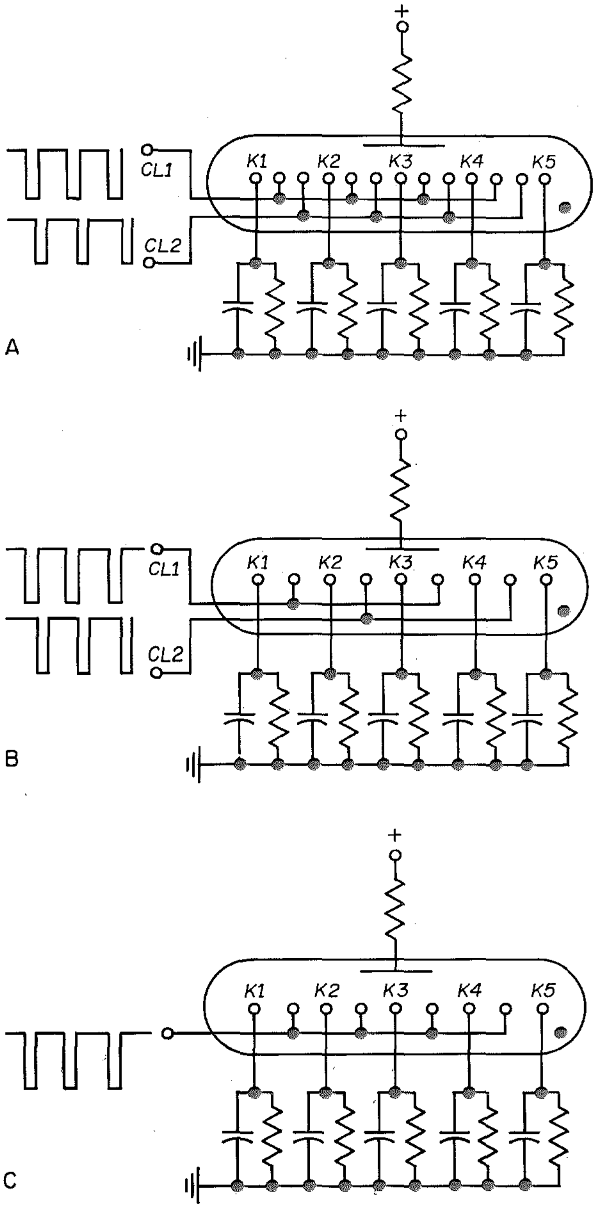


Figure 1—Some electrode arrangements for multicathode gas counting tubes.

a new discharge on an intermediate electrode, causing a reduction of the anode voltage below the level necessary to maintain the previous discharge. In type *A* (Figure 1), the pulse on *CL1* extinguishes *K1* and the overlapping pulse on *CL2* transfers ionisation to the proximity of *K2*. Type *B* has input pulses switched alternately between *CL1* and *CL2*, the order of the pulses deciding the direction of glow movement along the array. The arrangement of type *C* was adopted and became the basis[2] of a decade counter tube, the Nomotron.

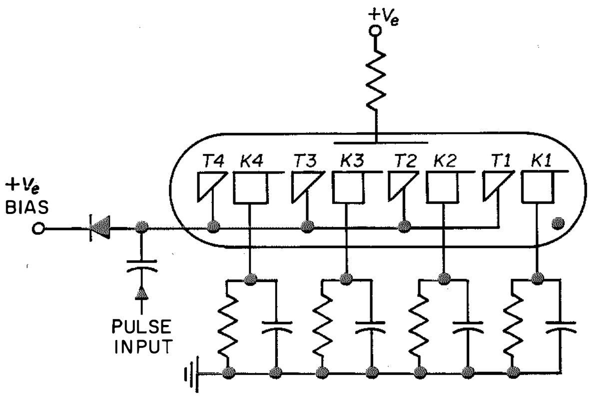


Figure 2—Diagramatic representation of electrode structure and basic circuit connections of Nomotron. The count progresses from right to left.

1. Nomotron Operation

Directional glow stepping in this tube is based on the glow characteristic of a geometrically asymmetric cathode. It is seen from Figure 2 that the cathode has two distinct regions: —the plate and the thin edge referred to as the tail. Apart from the properties of the cathode material and the gas, the voltage required for maintaining discharge also depends on diffusion of charges. This factor is largely controlled by the ratio of peripheral to total area, consequently the maintaining voltage V_m of the tail portion is higher than that of the cathode body and the glow discharge will be more easily maintained on the plate.

Having established this condition, it follows that the transfer electrode adjacent to the glow will be preferentially primed. All transfer electrodes are commoned as shown in Figure 2; the transfer system is directional when the time constant of the cathode circuits is long compared with the input pulse width, because bias remains on the extinguished cathode when another cathode breaks down. The cathode current is nominally 3·5 milliamperes and the tube operates at frequencies up to 20 kilocycles with a negative 10-to-15-microsecond input pulse. Trouble was experienced in some applications of this tube due to changes in characteristics after long periods

[2] G. H. Hough and D. S. Ridler, "Cold-Cathode Glow Discharge Tubes," *Electronic Engineering*, volume 24, pages 272–276; June, 1952.

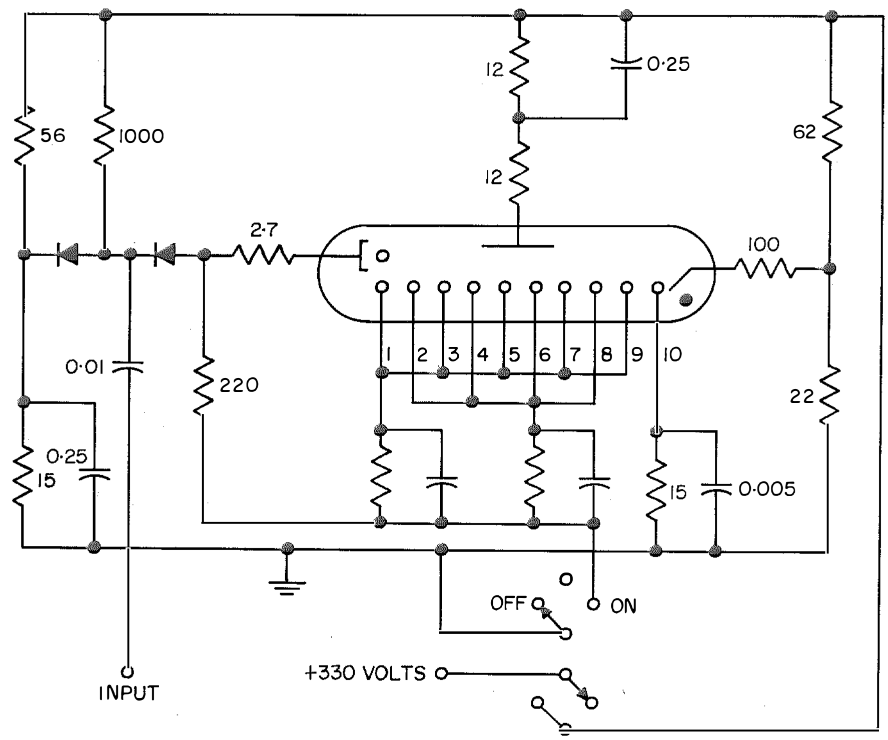


Figure 3—Nomotron circuit for operation up to 5-kilocycle input recurrence frequency. Resistances are in kilohms and capacitances in microfarads. The input is a series of negative square pulses of 120±15 volts amplitude, 16±4 microseconds wide.

of conduction to a particular cathode. These changes were attributed to cataphoresis, by which gaseous impurities were concentrated on electrodes adjacent to a conducting electrode. This phenomenon has now been eliminated[3] and satisfactory operation obtained. A complete circuit for operation at input recurrence frequencies up to 5 kilocycles is shown in Figure 3. All 10 cathodes are brought out to enable the tube to be used as a distributor when required.

2. *Trigger Tube G1/371K*

To complete gas-tube counter systems, it was necessary to have trigger tubes covering a similar frequency range and the *G1/371K* trigger tube was developed for this purpose. The salient points follow. Photo-electrons are produced in the trigger–cathode gap by a beam of photons from a priming discharge. To obtain close limits on trigger–cathode breakdown voltage, the dimensions are arranged so that the gap is equal to the physical length of the cathode fall region; that is, at the pressure times distance value for the minimum of the Paschen curve. With the anode–cathode gap chosen, the glow will transfer

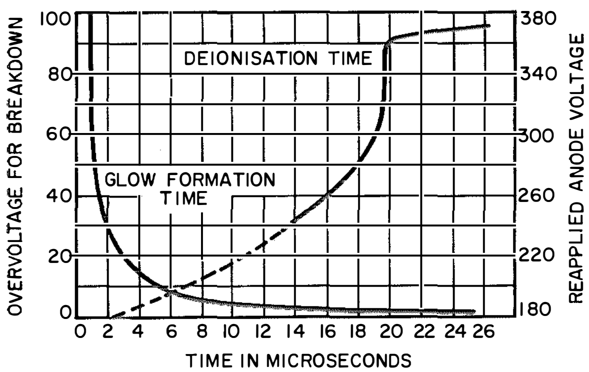


Figure 4—Breakdown voltage and deionisation time for *G1/371K* trigger tube.

from trigger to anode over a wide range of applied voltages. Metallic channels joining the anode and cathode support micas restrict the total diffusion volume and assist deionisation. Figure 4 shows typical breakdown and deionisation time characteristics. The tubes are made to close mechanical limits and are individually tested on most parameters. Large numbers of tubes are in use and have given good performance with little change in characteristics after 25 000 hours of use.

[3] T. M. Jackson, "Improved Circuit for Reliable Operation of Nomotron Counter Tubes," *Electronic Engineering*, volume 29, pages 324–326; July, 1957.

3. *Speech-Gap Tube*

Voice-frequency electronic switching technique requires a device allowing the design of a completely electronic telephone exchange by provision of switched speech paths. The broad requirement for such a device can be stated as a relatively noise-free audio-frequency path of sufficiently low impedance and suitable power-handling capacity to allow the necessary number of series elements in a two-wire circuit to be used without amplification. Also, the static and dynamic characteristics should allow the switching operation to be performed by pulses. Clearly, gas-discharge tubes were capable of performing the switching functions but known tubes were unsuitable for speech transmission.[4] The first task was therefore to examine the transmission characteristics of gas discharges.

Examination of diode gaps of various forms showed the impedance to be made up of resistive and inductive components. The parallel inductance depends on current and frequency in the audio-frequency range. The result was that the impedance was not simply related to the slope of the static characteristic and was of the order of 1000 ohms at practical current levels.

An investigation of the impedance of the various regions of the glow discharge was carried out by means of moveable probes, on the basis that plasma regions of high density should show low resistance. With balanced probes biased to the space potential, hence not conducting current, it was found that the impedance between the probes was high in all regions of the discharge. It was also observed that noise was present in the positive column but completely absent in the Faraday dark space. Further tests made with probes adjusted to conduct current showed that the impedance was high in the positive column when conducting either ion or electron current but in the Faraday dark space, a low impedance was obtained when the probes drew electron current.

[4] A. H. Beck, T. M. Jackson, and J. Lytollis, "Novel Gas-Gap Speech Switching Valve," *Electronic Engineering*, volume 27, pages 7–12; January, 1955: also, *Electrical Communication*, volume 32, pages 179–189; September, 1955.

3.1 Double-Anode Tube

When the circuit was modified so that all the cathode current was taken by the probes, the main anode of the experimental tube could be

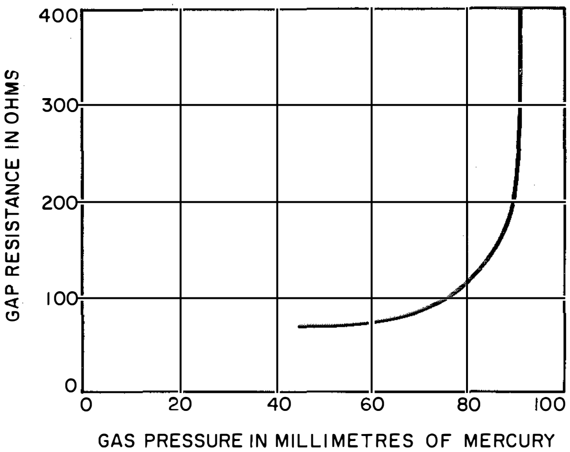


Figure 5—Variation of anode-to-anode resistance with helium gas pressure; anodes-to-cathode spacing is 2 millimetres (0·08 inch).

disconnected. The path between the two probes (now anodes) was found in this way to be substantially independent of their spacing, provided the anodes were situated on the cathode side of the outer edge of the Faraday dark space. A plot of resistance against pressure is shown in Figure 5 for pure helium at a spacing of 2 millimeters (0·08 inch). The resistance shows a rapid increase

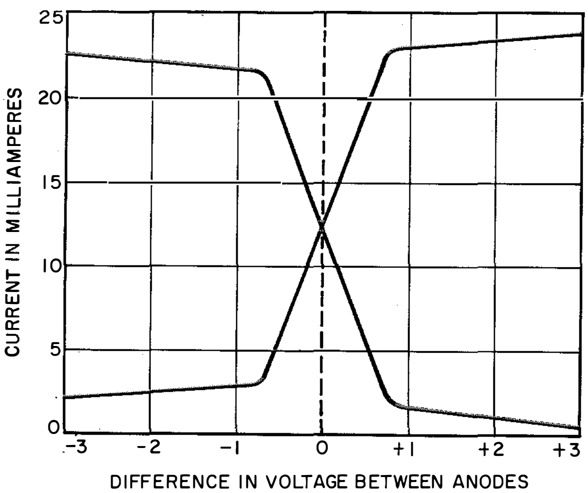


Figure 6—Static characteristics of speech gap. Resistance between each anode and the cathode is 1500 ohms; speech-gap resistance between anodes is 100 ohms.

when a positive column is formed and a space-charge sheath surrounds the anodes. The static characteristic is illustrated in Figure 6. The central intersection, at which the anode currents are equal, corresponds to no signal input. The slope of the characteristic gives the anode-to-anode (speech-path) resistance. It follows that a cathode current of 20 milliamperes can be switched entirely to either anode by a change of ± 1 volt; the modulation is 10 milliamperes. The mechanism of the low gap impedance was established by probe and other measurements that showed the plasma potential following closely the highest anode potential. Consequently, when the potential of one anode is raised, the other anode becomes negative with respect to the plasma and its current decreases. Similarly, when one anode voltage is lowered, the plasma remains at the potential of the higher anode. Figure 7 shows the variation of alternating-current re-

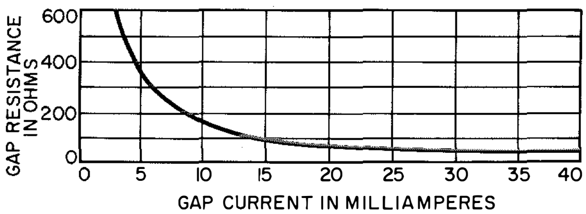


Figure 7—Variation of gap resistance with current.

sistance with current, which controls the plasma density and therefore the conductivity. Probe measurements indicate an electron density of 10^9 electrons per cubic centimetre. Figure 8 shows the alternating-current resistance to be independent of frequency. The curves go up only to 20 kilocycles but, in fact, the independence is maintained up to very-high frequencies, as it is solely an electronic effect. The points involved are best considered with direct reference to the design of the tube adopted. The tube itself is shown in Figure 9. It consists of a flat molybdenum cathode between two forked anodes in a subminiature bulb. A connection is made by a clip and metallised strip to the inside wall of the bulb. Simplicity of design and low cost are likely to be a major consideration in a device of this type because of the large numbers needed in a practical system. The various tube parameters will now be described.

3.1.1 *Switching Characteristics*

The requirements for use as a switch are:—Stable and reproducible static characteristics, minimum excess voltage for pulse breakdown, and, also, a minimum reduction in breakdown immediately after a conducting gap is extinguished.

3.1.2 *Maintaining Voltage* V_m

Because of the long-term stability required, it was decided to adopt the sputtered molybdenum technique, which was known from other work to be satisfactory. By this process, the cathode is cleaned by high-density ion bombardment, the sputtered material being deposited on the wall of the tube and thus preventing release of impurities from the glass during life. A connection is made to this sputtered layer to control wall charges that otherwise would cause variations in tube characteristics. By this means, stable 120-volt values of V_m were obtained.

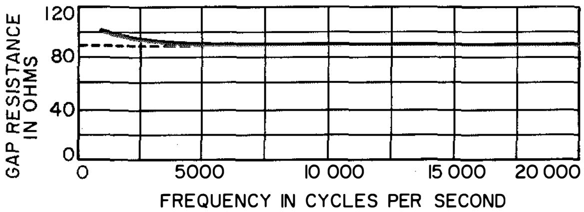


Figure 8—Variation of gap resistance with frequency.

3.1.3 *Breakdown Voltage* V_b

The aim in this case is to obtain the maximum value consistent with low resistance; consequently, the pressure times distance product was fixed just below the knee of the graph of Figure 5, giving a V_b of 270 volts.

3.1.4 *Pulsed Breakdown*

It is well known that when a limited time is allowed for breakdown, the breakdown becomes subject to statistical variation and, furthermore, in the absence of these variations a glow formation time is observed that is related to the dark current and the excess applied potential. The former effect is eliminated and the latter reduced by priming. The possibility of using radioactive priming was considered and, in particular, the use of tritium was investigated. It appeared very attractive from many aspects, especially the low particle energy ($\approx 10^4$ electron–volts) and half-life of about 12 years. Tests showed tritium to be reasonably effective but difficuties were encountered in admitting known quantities into the tube and effectively controlling the process.

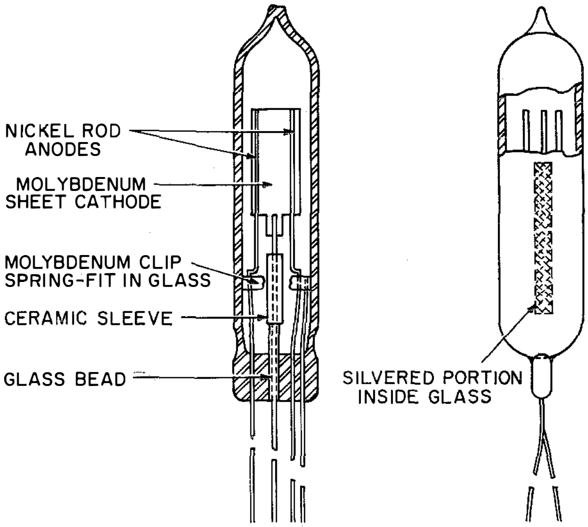


Figure 9—Construction of gas-tube switch.

Tritium was eventually replaced by an auxiliary glow discharge. This was arranged by dual use of the internal clip connection—which was connected to a negative voltage through a high resistance, A subnormal glow discharge of about 10^{-6} ampere was maintained to a selected point on the clip lead. This was found to be extremely efficient in priming the main discharge, giving pulse breakdowns with only a small overvoltage.

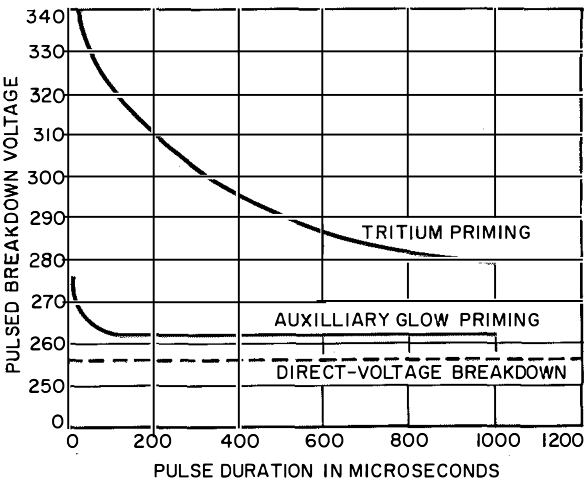


Figure 10—Pulsed breakdown voltage versus pulse width.

Dynamic breakdown characteristics are shown in Figure 10 for tritium and auxiliary primed tubes.

3.1.5 Reapplied Anode Voltage

Reapplied anode voltage is the highest voltage that can be reapplied to the gap after a specified time without causing a further breakdown. It is related to static breakdown by deionisation time and other residual effects. In the present tubes the time allowed is an order of magnitude greater than the deionisation time and the value obtained is largely controlled by cathode surface effects and depends on the magnitude of the current before extinction. The reapplied voltage may be 15 volts below the static breakdown voltage.

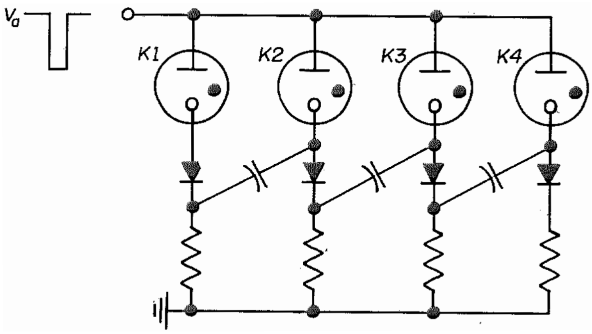


Figure 11—Diode switching circuit.

It is important to note that, though static breakdown can be controlled to within close limits, the dynamic limits are much wider. This is particularly true in the case of coordinate switching using double pulses superimposed on standing bias, since a single pulse applied immediately after extinguishing a gap may cause it to refire. Furthermore, standing voltages must be appreciably higher than the tube drop to permit sufficient current stabilisation, which is important in balancing the system. The characteristics obtained from a tube of this design provided satisfactory operating conditions for model exchanges.

Early life tests showed that characteristics changed and it was established that these changes were a direct result of the method of processing the tube. To minimise overall pumping time, the tubes were processed in neon to take advantage of the high sputtering rate. The neon was then pumped away and the final gas filling of helium admitted. It was found that neon was released from the cathode during life and causing the change in characteristics. Methods were subsequently developed for processing and filling only with helium as this was the most-suitable gas filling for the required characteristics. Satisfactory life was then obtained. In the application envisaged for this device, 4000-hour life with continuous conduction is equivalent to 25 years of normal service.

4. High-Speed Triode-Switch Operation

Attention was now directed towards the possibility of higher-speed operation. Initial studies showed that individual elements with external coupling showed more promise than multi-electrode tubes operating on the ionisation glow transfer principle. A new technique that largely removed the circuit limitation was evolved and permitted the use of tubes to the limit of their characteristics.[5] The basic principle is to extinguish a conducting gap by a negative pulse on the commoned anodes and to apply the resulting change in cathode voltage to initiate a discharge in another gap.

4.1 Circuit Description

This technique is suitable for either diode or triode elements. It was established, using *G1/371K* trigger tubes, that operation was feasible at input frequencies in excess of 100 kilocycles —limited by the deionisation time of 10 microseconds at low current level. The simplest case, using diodes, is shown in Figure 11. With anode voltage V_a set below breakdown voltage V_b, an output voltage $V_o = V_a - V_m$ appears on conducting cathode *K2*. This is the steady-state condition in which *K3* is restored to earth through the low forward resistance of the crystal diode. Application of a negative impulse voltage greater than $V_a - V_m$ extinguishes the glow on *K2* and the output voltage decays towards earth. This results in a similar change in voltage appearing at *K3*, negative with respect to earth and developed across the high reverse resistance of the crystal diode. Providing the sum of the negative change on *K3* and V_a is greater than the instantaneous V_b of gap *3*, discharge will be initiated on *K3*. For a given deionisation time, the maximum frequency is doubled by use of biphase

[5] British patent 33115/54.

waveforms connected to alternate anodes. The maximum operating frequency then occurs when the pulse spacing equals the deionisation time.

It was concluded that provided deionisation time, glow formation delay time, and build-up time could be substantially reduced, an element capable of working in the 0·5-to-1-megacycle range could be made. Independent investigations of the individual parameters involved were made.

4.2 Glow Initiation

Published work shows that short time lags for spark breakdown are obtained with low overvoltage (10^{-6} second with 1-per-cent overvoltage for 1-centimetre gap). These workers have invariably used high values of pressure times distance, greater than 200 millimetres of mercury times centimetres; a region where considerable controversy still exists regarding the exact mechanism of breakdown. The present experiments have been restricted to low products, approximately 20 millimetres of mercury times centimetres, where the validity of the Townsend model is undisputed. From earlier work, some difficulty was anticipated in obtaining a substantial reduction in glow formation time. Here this time is defined as the interval between application of a specified overvoltage and the establishment of a self-sustained glow. This requires field distortion by the formation of a positive-ion space charge in front of the cathode. For uniform fields, increase in priming current reduces the dynamic breakdown potential but does not greatly alter the overvoltage for a given time delay because the static breakdown voltage is also lowered. Only when priming is sufficient to reduce V_b to V_m is the formation delay eliminated. In this condition the space charge is already formed and occurs when a primed electrode is situated inside the cathode dark space of a priming discharge. Such conditions, however, are not useful when the breakdown is required to be considerably higher than the maintaining voltage. For this condition, therefore, with parallel plane electrodes, the field strength at the cathode is sufficient to produce breakdown at very-low values of ionisation current and short glow formation time can only be obtained with relatively large overvoltage. The obvious requirement is to provide a stable condition with a relatively high degree of ionisation.

Alternative geometries were examined and it was established that suitable conditions were provided by both concentric and spherical structures operating with positive wires and points. The field strength near the anode is greatly enhanced and the field at the cathode is reduced in the ratio of the radii. This results in high gas-current amplification and produces an ionised state in the anode region. The current density near the cathode, however, is comparatively low, resulting in a low secondary coefficient. The overall result is that high values of prebreakdown current are obtained without breakdown This stabilises the breakdown of the gap. If a concentric diode with a source of priming charges along its axis is considered, the high field strength in the anode region will produce ionisation there, resulting in the formation of a partial space charge and only a small amount of additional energy is required to cause breakdown. This results in a short glow-formation time. Typical curves are shown in Figure 12 for concentric and spherical gaps. Two methods have been devised for producing the required conditions.

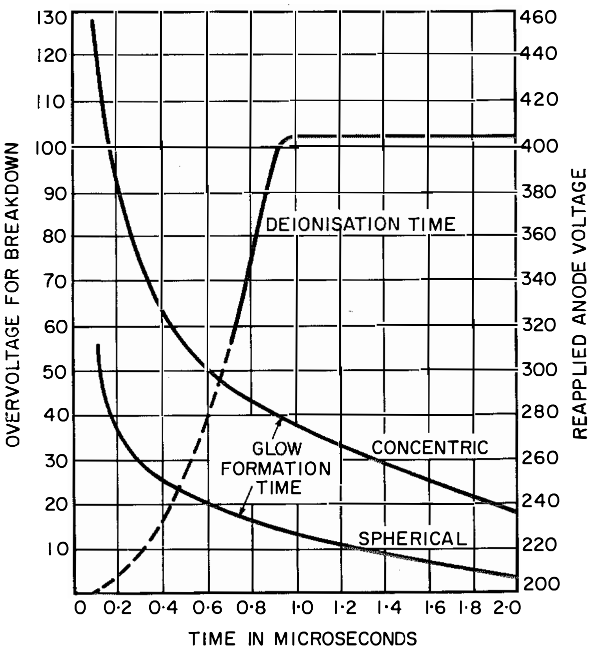


Figure 12—Typical breakdown and deionisation time characteristics for diode gaps. Priming current $\approx 10^{-3}$ ampere.

4.2.1 *Corona Discharge at Positive Wire*

A complete volt–ampere characteristic of a tube with a fine anode wire is shown in Figure 13, the initial electrons being obtained by using the anode wire as a source of light. The plot shows that the current increases with applied voltage up to point *B* where corona occurs and then increases sharply to point *C*, thereafter it increases at approximately 2 microamperes per volt to point *D*, where the gap breaks down into a self-sustained discharge. A similar effect is observed with anode cold, though at a higher anode voltage, depending on the wire radius and the dielectric strength of the gas. The important point here is that the formation delay times above and below point *B* are enormously different. With V_a set above *B*, delays are very short but below this point, time lags become very long. The successful application of corona priming would produce a very-simple and attractive device but some troubles have yet to be overcome. The main difficulty is the small voltage difference between corona breakdown and breakdown proper, a substantial differential being required to ensure fast-enough corona reformation time after application of an extinguishing pulse. This becomes important in application and involves the state of the next tube following a discharging tube. It will be remembered that the extinguishing pulse is applied in common to all anodes. If corona is extinguished in the tube concerned it must reform in a time less than the pulse width, otherwise the tube will be inadequately primed. The details are still being examined.

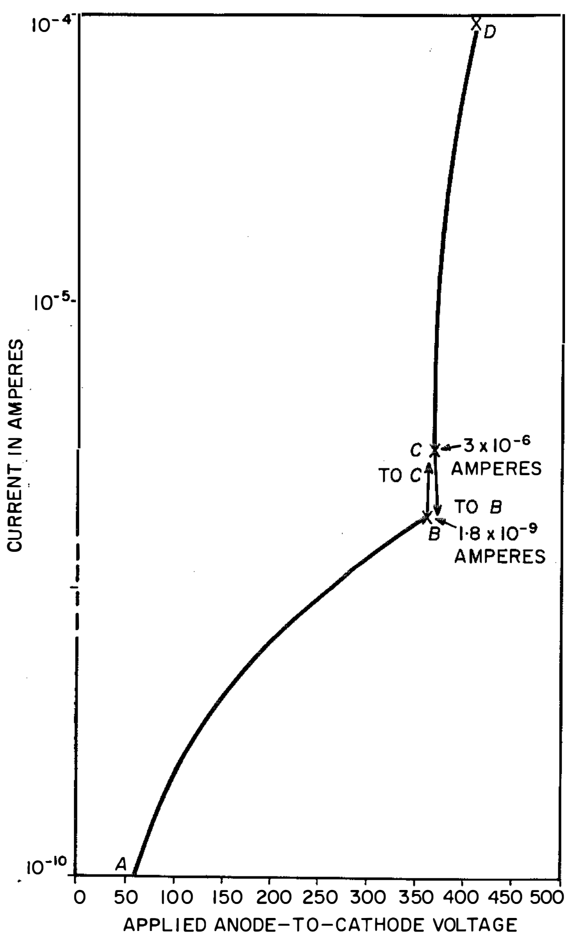


Figure 13—Voltage–current characteristic showing corona priming.

4.2.2 *Central Priming Discharge*

Similar stable prebreakdown currents are obtained in the constructions shown in Figure 14, which also shows a plot of the prebreakdown current against anode voltage.

4.2.2.1 *Concentric Geometry*

A fine tungsten wire is stretched along the axis of an apertured cylindrical anode inside a cylindrical cathode. By maintaining a direct-current discharge with the centre wire as cathode, some ionisation occurs in the main gap producing the breakdown characteristic plotted.

4.2.2.2 *Spherical Geometry*

Two fine wires are beaded on a ceramic disc which fits into a domed cup. This is a very-simple construction giving stable characteristics.

4.3 Deionisation

Considering first a discharging gap, the potential distribution is such that almost the entire applied voltage appears across the cathode dark space because the field is distorted by the presence of the positive-ion space charge. Within a certain specified time, the charge density must be reduced below a certain level, which, when the field is reapplied, will be insufficient to reform the discharge. Several factors involved in this process are listed below:

A. Gas filling and pressure (involving mobility K and pressure p).

B. Spacing d.

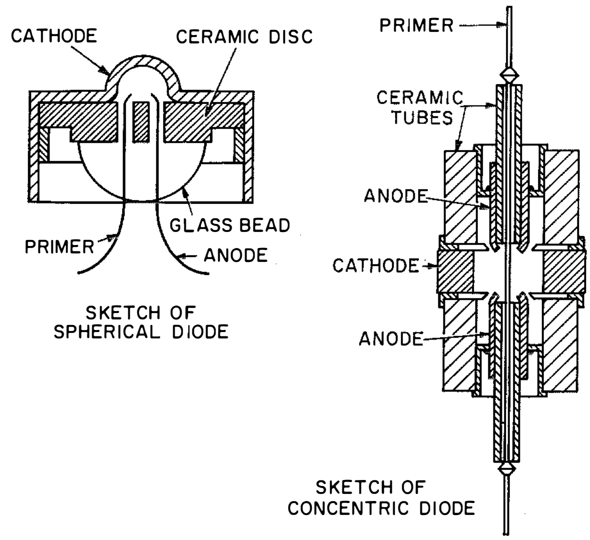
C. Sweeping field X.

D. Geometry.

E. Current.

F. Diffusion volume.

Before considering these points in detail, it should be mentioned that previous work had established the use of admixtures of hydrogen as a deionising agent. Metastable states of long lifetime are eliminated by the introduction of hydrogen and deionisation time can be reduced from milliseconds to microseconds. Any further reduction in time must be related to the mobility of the particles involved and, as electron mobility is at least 10^3 times greater than ion mobility, it is with the elimination of ions that we are concerned.



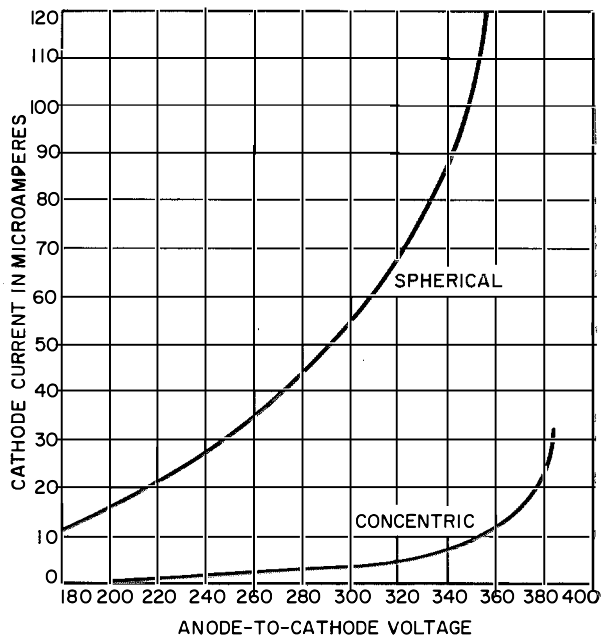


Figure 14—Prebreakdown current versus gap voltage for two tube structures. Priming current $\approx 10^{-3}$ ampere. Sketches are about 2·5-times actual size.

It has been found that the deionisation time τ is approximately equal to d/KX, the drift velocity being KX with $X = V/d$ being the field applied across the gap during the deionisation period and set just below the maintaining potential V_m. Voltage V_m is relatively independent of p and d over the range involved. Breakdown voltage V_b is set by circuit requirements and is a constant times p times d for a given gas.

$$\tau = d^2/KV_m$$

But $K = C_1/p$ and $p = C_2/d$. Thus,

$$K = dC_1/C_2$$

and

$$\tau \doteq dC_2/C_1V_m$$
$$\propto d.$$

In practice, as d is made smaller and p increased, the current density, which is inversely proportional to p^2, becomes very high resulting in contraction of glow and requires a small cathode area to maintain glow stability. Reduction of the cathode area magnifies edge effects and therefore conditions have to be optimised to the extent of obtaining usable values of deionisation time, V_m, output voltage, and current. The relevant factors controlling the selection of gas filling are positive ion mobility and dielectric strength, the latter permitting a reduced gap for a given breakdown. Hydrogen best satisfies these requirements but has a maintaining potential of about 300 volts, which involves high operating voltages. Mixtures of inert gases have therefore been used in which the hydrogen content is optimised to be most suitable for producing the required operating conditions.

4.4 Geometry

Most early work was carried out using parallel plate electrodes but it was apparent that the field strength at the cathode would influence the reapplied voltage level; a lower field would permit a larger amount of residual ionisation without refiring. Fortunately, therefore, structures that have been shown suitable for rapid glow initiation are also most suitable for fast deionisation time. In a conducting gap, the nonuniform field goes over to roughly parallel-plane conditions due to the presence of the positive ion space charge. On extinction, the important process is the collapse of the space charge, the field thereafter reverting to radial form. The net result is that a higher reapplied voltage is permissible after the collapse time, which can be interpreted as a shorter time for a reapplication of a given percentage of V_b. It is not claimed that geometry is the major factor in the reduced deionisation times (factor of about 100 overall), but rather the general scaling down. It is important to note, however, that the deionisation time becomes very long if the gap is reversed by using $-V_e$ wire. A typical deionisation time characteristic is shown in Figure 10 and may be compared with the corresponding curves for the *G1/371K*, shown in Figure 4, which is, at present, the highest-speed cold-cathode trigger tube available.

Tubes of the above designs have been demonstrated to operate at input recurrence frequencies up to 1 megacycle, but their final design has not yet been fixed.

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Transition-Rate Discriminators

By J. D. HOLLAND

Standard Telephones and Cables Limited; London, England

TELEGRAPH signalling waveforms consist of a series of mark–space or space–mark transitions that represent points at which the rate of change of information is a maximum. The number of transitions that occur over a given period for an error-free message depends on the signalling speed. If the message is disturbed by noise the transition rate increases, and for certain types of interference the rate may tend to become zero.

The difference in transition rate between an error-free message and one perturbed by noise can be detected, and the information used to provide a utilization voltage. Circuits have been developed for this purpose, known as transition-rate discriminators; they have a number of applications.

1. *Transition Rate of a Telegraph Signal*

Examination of the 5-unit start–stop telegraph code shows that 6 characters have 6 mark–space or space–mark transitions, 6 characters have 2 transitions, and the rest have but 4 transitions of either type. In contrast, the 7-unit error-detecting code has between 1 and up to 6 transitions per character.

With the 5-unit code the transition arrangement is shown in Table 1.

TABLE 1

5-UNIT-CODE TRANSITION ARRANGEMENT

Number of Characters	Number of Transitions	Characters
6	2	M, O, T, V, LETTERS key, and BLANK
6	6	D, F, J, R, S, and Y
20	4	All the rest

The transition rate can conveniently be expressed in terms of frequency as:—

$$F = N/2D$$

where

F = transition rate in cycles per second

N = number of mark–space or space–mark transitions per character

D = duration time of a character in seconds.

Hence, for the 5-unit code at 50 bauds the rate for 2-, 4-, or 6-transition characters is closely equal to 6·6, 13·3, and 20 cycles per second, and the average number of transitions per character taken over a large number of randomly chosen characters is 4, that is, at a rate of about 13·3 cycles per second.

For example, in the standard test sentence, "The quick brown fox jumped over the lazy dog's black," the average rate is about 14·3 cycles per second.

An analysis of encoded or clear transmissions or a mixture of both infers that, at 50 bauds, the minimum and maximum rates are closely 10 and 18 cycles per second respectively. This does not include reversals (RYRY ...) that provide no semantic information and are discarded. (The rate for this signal type is about 21 cycles per second.)

2. *Frequency of Fluctuation Noise in a Given Bandwidth*

Rice[1] has shown the noise currents in a narrow-band filter behave like a sine wave of frequency $(f_1 + f_2)/2$ and of an amplitude that fluctuates at an irregular frequency of $(f_1 - f_2)/2$, where f_1 and f_2 represent the band limits of the filter. The output from the filter therefore resembles a single frequency, equal to the midband frequency, modulated in amplitude and phase. Since the amplitude modulation frequency approaches one half the filter bandwidth it seems reasonable to suppose that the phase modulation frequency does also.

If this spectrum is limited, applied to a demodulator and then to a frequency counter, the counter will register one pulse per cycle and the noise count will be closely equal to one half the bandwidth in cycles per second.

[1] S. O. Rice, "Mathematical Analysis of Random Noise," *Bell System Technical Journal*, volumes 23 and 24, numbers 1 and 3; July, 1944 and January, 1945.

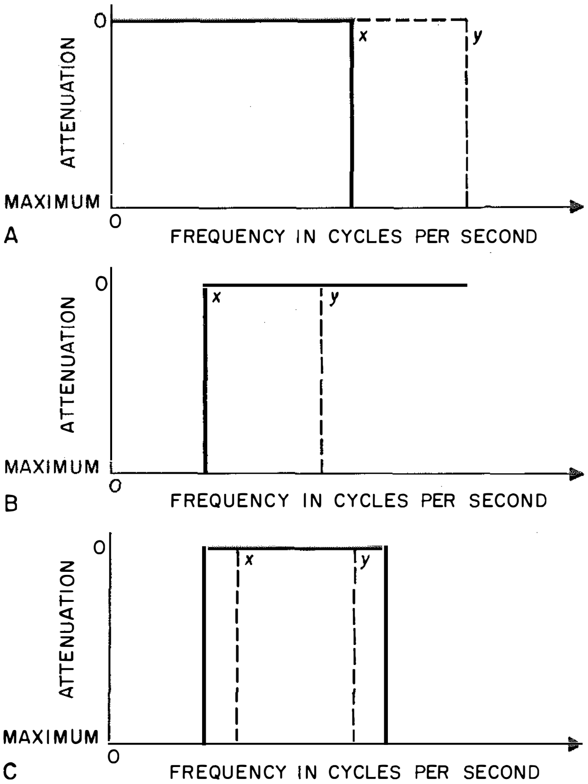

Figure 1—Characteristics of transition-rate discriminator.

3. Discrimination Between Noise and Signal Message Frequencies

The bandwidth in cycles per second required before demodulation of a telegraph signal of quasirectangular form is greater than the message transition rate in cycles per second. For example,[2] at 50 bauds, a bandwidth of about 100 cycles per second is required and hence the noise count will be about 50 cycles per second whereas the maximum rate due to a meaningful message is closely 18 cycles per second.

This difference in frequency can be detected by a transition-rate discriminator, and the type of action obtained is shown in Figure 1A, 1B, and 1C. In Figure 1A, at x, maximum attenuation is obtained to frequencies exceeding those that could occur in an error-free message and point y is another arbitrary cut-off point that would be used at higher keying frequencies; necessitating increased bandwidth.

In Figure 1B, maximum attenuation is obtained to frequencies lower than the lowest message frequency at the arbitrary cut-off points x

[2] L. J. Heaton-Armstrong and J. D. Holland, "Direct-Printing Receiving Systems at Low Radio Frequencies," *Electrical Communication,* volume 35, number 3, pages 202–208; 1958.

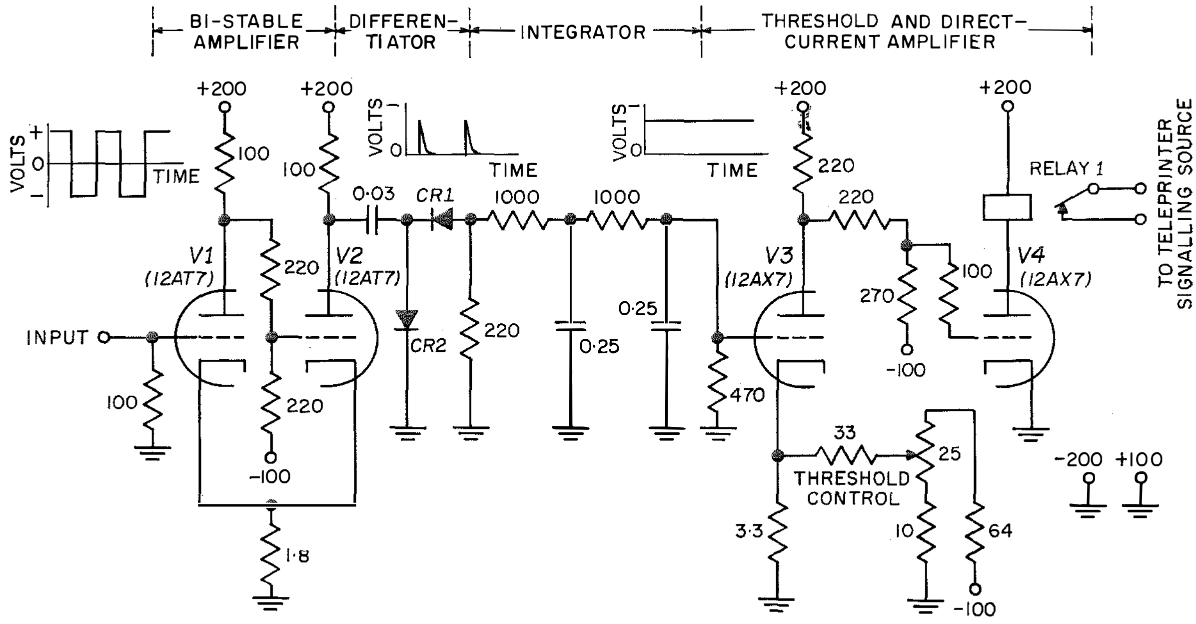

Figure 2—Transition-rate discriminator for 50-baud operation. Values of resistance are in kilohms and values of capacitance are in microfarads.

and y. Figure 1C shows the type of characteristic that can be obtained by the use of two discriminators of the characteristics shown in Figure 1A and 1B. In this case, frequencies above or below the band occupied by an error-free message are suppressed.

4. Transition-Rate Discriminators

One form of circuit[3] and the functions obtained is shown in Figure 2. Due to the use of a bi-stable amplifier the positive-going output from *V2* can be short-circuited by use of diode *CR2* since the transition history of either polarity of the wave is identical.

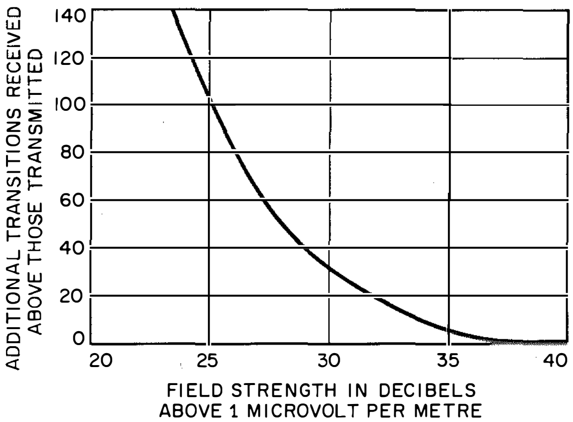

Figure 3—Spurious extra transitions received as a function of the field strength of the received radio signal.

This form of circuit is suitable for connection to the output of a demodulator. For example, the number of extra transitions received above those originally sent in a random message against field strength is shown in Figure 3. These characteristics were taken at the output of the demodulator of an *RV.14* receiver. Error-counting tests showed that the error rate was about 1 in 1000 at a field strength level of 30·3 decibels relative to 1 microvolt per metre.

Figure 4 shows the results obtained after demodulation, using reversal-type signals or signals of random character from a high-frequency receiver. The signal-to-noise ratio of the receiver was approximately 12 decibels for an input of 0·1 microvolt in 75 ohms.

Valves *V1* and *V2* (Figure 3) can be dispensed with and connection made to the output terminals of the telegraph receiver, providing the in-built circuits are bi-stable. Similarly valves *V3* and *V4* can be omitted if the negative-going utilization voltage across the 470-kilohm resistance is required for other applications. Some form of threshold circuit should be retained.

For the circuit shown in Figure 2, the characteristic of the output level versus frequency across the 470-kilohm resistor is shown in Figure

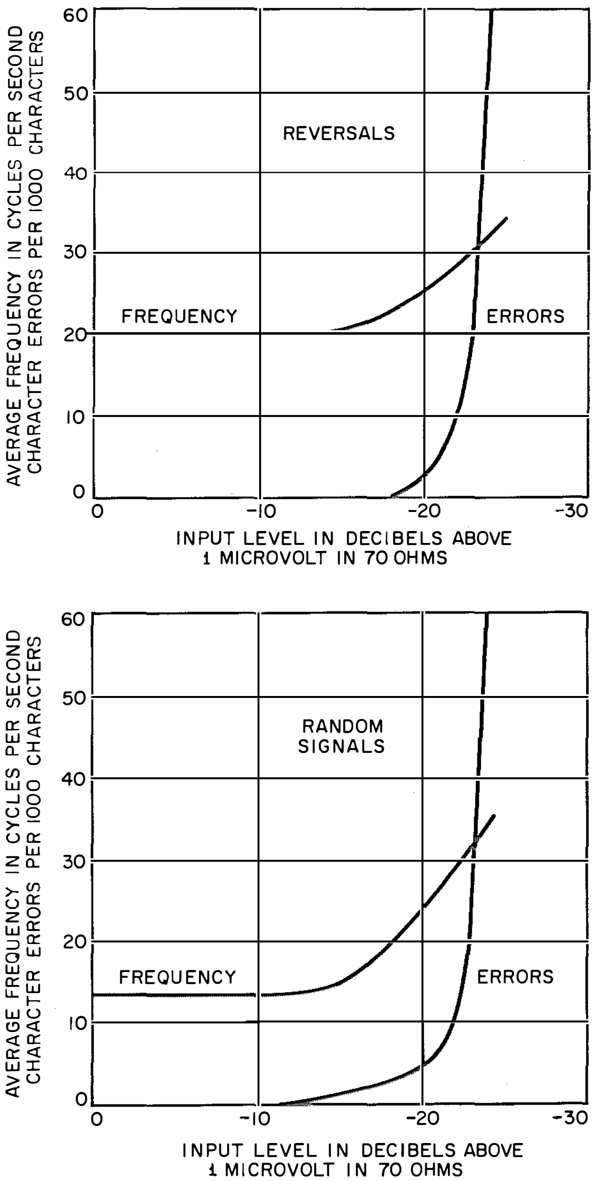

Figure 4—Average message frequency in cycles per second or character errors per 1000 characters as a function of input signal level for two types of signal. Signalling speed = 50 bauds; bandwidth = 120 cycles per second.

[3] British Patent Application 7077/57.

5 and the input required at the grid of *V1* should not be less than ±5 volts. The threshold control is effective to within a change of ±1 cycle per second of the transition frequency.

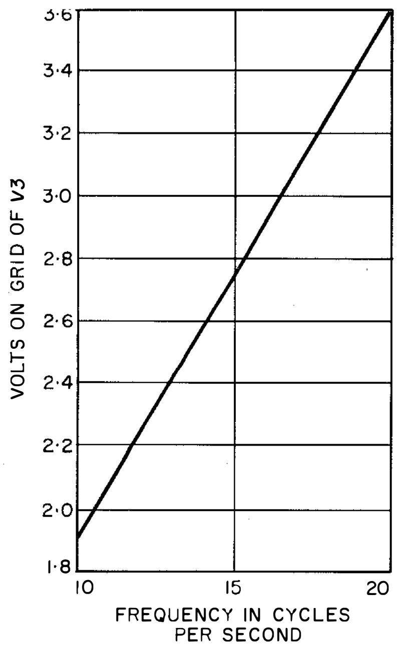

Figure 5—Voltage appearing on the grid of *V3* as a function of frequency with an input of not less than +5 and −5 volts on the grid of *V1*.

5. Application

5.1 Prevention of Printing Random Symbols

At 50 bauds, the threshold control of the circuit shown in Figure 2 is set so that the relay contacts are opened at 19 cycles per second.

If the keying speed is not known, correct operation is obtained by monitoring on a reversal signal (RYRYRY...) and adjusting the threshold control until repetition of these characters is eliminated.

5.2 Signal Combination

When signalling currents carrying the same information, but arriving over different channels or radio paths, are combined, the effective output after combination will be mutilated if any of the sources are perturbed by noise of a higher level than the other contributions. This has been checked in the field on a dual diversity system. It was found that without transition-rate discriminators in each channel the output, with severe noise from one path, was completely mutilated, whereas with the discriminators in circuit before the combining process, the error rate was of the order of 1 in 2000 characters.

5.3 Dual-Diversity Switched-Antenna Receiving System

Dual-diversity spaced-antenna systems are of necessity complex in that separate high-grade receivers are used requiring periodic alignment checks to verify that equal weighting is being given to each path over the frequency range.

To reduce equipment investment, systems have been proposed whereby one receiver is switched between antennas in such a manner that the receiver is always connected to the antenna offering the highest signal level.

Tests show[4] that the error rate is reduced as compared to reception on a single antenna, and although this rate can never be as low as a system employing separate receivers and proper combining methods, a considerable saving in equipment is obtained.

The difficulty with the switched-antenna scheme is that since the switching information is obtained on an amplitude basis, the action required is likely to be indecisive when high noise conditions exist at one antenna site and a usable signal exists at the second antenna.

A system is proposed[5] in which the switching information is obtained from a transition-rate discriminator, having characteristics as shown in

[4] "Report on the Testing of an Antenna Diversity Equipment," internal report of Standard Elektrik Lorenz A. G., Pforzheim; 1956.

[5] British Patent Application 24649/58.

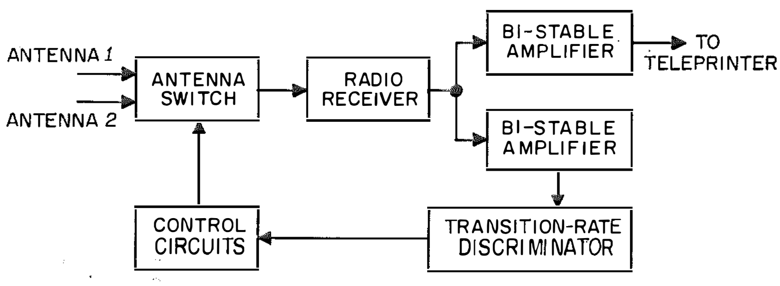

Figure 6—Diversity system using antenna switching to select best signal.

Figure 1A, in such a manner that the receiver is switched to a second antenna before reception on the first antenna becomes poor enough to cause appreciable errors.

A block diagram of the system is shown in Figure 6, and the detailed connection of the control circuits are shown in the reference.

The demodulated signal from the receiver goes to two bi-stable amplifiers. The triggering level of the amplifier actuating the teleprinter is so arranged that it will change state close to the steady-state condition of the wave whereas the amplifier forming part of the transition-rate discriminator changes state close to the zero level datum of the demodulated wave.

It can be shown[6] that, if the triggering level is the same as the steady-state condition of the wave, the value of the signal-to-noise ratio before a mark is turned into a pseudo space is −6 decibels; that is, the noise must exceed twice the momentary signal level. Alternatively, the ratio approaches unity for triggering levels close to the zero voltage datum line; and therefore the decision to switch should occur before a mutilization is recorded, provided that the fading rate is not too rapid.

[6] D. A. H. Johnson, "Laboratory Report 10," New Zealand Post, Telegraph, and Telephone Administration; 1957.

6. *Summary*

The probability of the presence of a signal in noise depends on the degree of *a-priori* knowledge of its precise form. Usually the maximum amplitude or the total received energy is examined and an existence probability is determined. Transition counting may represent a useful aid, since additional information, in terms of transition rate, is derived from the waveforms apart from an examination of their amplitude or energy content.

Probability Theory in Telephone Transmission*

By B. B. JACOBSEN

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TELEPHONE switching problems mean endeavour to get a call connected through a desired channel with a certain specified expectation of success.

In long-distance telephone transmission the endeavour is to transport speech information over long distances with a specified expectation of the impairment that the speech signal will suffer.

A common endeavour in these two fields is to carry out the separate purposes in the most economical way and in doing this it is usually necessary fully to exploit "the specified expectation" since a design that gives too good a performance will usually be more expensive than one that only just meets the requirements.

One very important cause of impairment in speech transmission is noise. In long-distance telephone circuits there are several mechanisms that produce noise. The most important ones are:—

(A) Thermal agitation.

(B) Nonlinear (intermodulation) distortion in multichannel amplifiers.

(C) Disturbances from other circuits.

The magnitude of noise in the circuit due to these mechanisms can generally speaking be reduced to almost any desired degree but such reduction strongly affects the cost of the equipment required and it is therefore very important to find out how much noise can be tolerated. This meets with two difficulties:—

(A) The tolerable noise power at any time depends on the volume of the speech signal that is available at the end of the circuit at that time and this volume varies from call to call due to many factors, see for instance reference 2, page 29.

(B) The amount of noise present in the long-distance circuit will fluctuate with time, for instance, that part of the noise that arises from nonlinear distortion in a multichannel carrier system will fluctuate with the signal loading of the amplifiers[1–3] and in a radio link the thermal agitation noise will fluctuate due to fading of the radio path. The fluctuation involved is that of the mean power taken over a few hundred milliseconds and not the faster fluctuation that is present in thermal agitation noise.

One way of specifying the acceptable amount of noise is shown in Figure 1A. This specification amounts to allowing a certain maximum value (−50 decibels relative to 1 milliwatt) of noise power at a particular point in the long-distance circuit but with the escape clause that for 1 per cent of the time (in general during the busy hour when high noise is expected) no limit is placed on the amount of noise.

This way of specifying the noise is very simple but it does not take into account the full facts of the actual situation. It is not a very complete statement of the amount of noise that can be tolerated without seriously disturbing telephone connections. If in a particular design the noise statistic should be a straight line with a certain slope, then, for maximum noise of this type, the noise statistic line would have to meet the specification at the 1-per-cent point; and everywhere else the noise would be much smaller than would be acceptable.

This is an unnatural restriction and the result of the simple way in which the specification in Figure 1A is framed.

In Figure 1B is shown a more-recent proposal for a specification for the permissible noise, which is now limited in two ways.[5]

* Reprinted from *Teleteknik* (Copenhagen), volume 1, number 1; 1957. Presented at First International Congress on Application of the Theory of Probability in Telephone Engineering and Administration; Copenhagen, Denmark; June 20–23, 1955. The purpose of this paper is to point out that probability theory is used not only in telephone switching but also in telephone transmission. This is done by giving an example.

[1] All references will be found in the bibliography.

(A) The mean power of the noise over the whole hour shall not exceed -50 decibels relative to 1 milliwatt.

(B) The statistical distribution of the noise shall fall below the line shown.

This specification has been derived by taking account as far as possible of the known fact that the speech volume will fluctuate from call to call and the likelihood that the noise in actual systems will also fluctuate. The sloping line in Figure 1B is part of a log-normal distribution curve; that is, one in which the logarithm of the variant is normally distributed. This type of distribution is very appropriate for telephone transmission and is followed fairly accurately by a number of variables. The speech volume at a certain point in the telephone network has been found to be distributed in this way with a standard deviation of the order of 6 decibels. Now Figure 1B is in some cases easier to satisfy than Figure 1A but at the same time the actual nuisance value of the noise is no greater than for the specification in Figure 1A. Figure 1B is a more-accurate statement of the noise that actually can be tolerated. While Figure 1A might be called a convention, Figure 1B takes more account of the actual situation. The specification in Figure 1B has the advantage that families of analytical curves can be derived from it, each of which meet the overall specification (Figure 1B) in two points—they have the correct total mean power over the hour (average power *not* the logarithm of power (decibels)) and the extreme peaks meet the peak specification at 0·03 per cent. In Figure 1C is shown two extreme curves *a* and *c* which meet the specification—one suitable when the actual noise is very constant (*a*), and the other (*c*) suitable when the noise is highly fluctuating. A third curve *b* is shown as a representative of a whole family of similar curves that may be drawn—each with the required mean power over the test hour and all meeting the peak part of the specification in Figure 1B at its highest point. This particular family of curves has been named "augmented log-normal distribution curves"; they are in fact log-normal curves to which a constant power has been added. The mean power of the log-normal curve itself is in logarithmic measure: Median $+ 0{\cdot}115\ \sigma^2$ where σ is the standard deviation in decibels.

The sum of the mean power of the log-normal distribution and the added constant power is kept constant at the specified value for the family of curves.

Both the statements Figure 1A and Figure 1B are intended to apply to circuits of lengths of 2500 kilometres and now comes a problem: how are we to specify the requirements for a shorter length of circuit (partial circuit) such that when a number of partial circuits are put together in tandem to produce a circuit 2500 kilometres long the resultant noise will meet the specification.

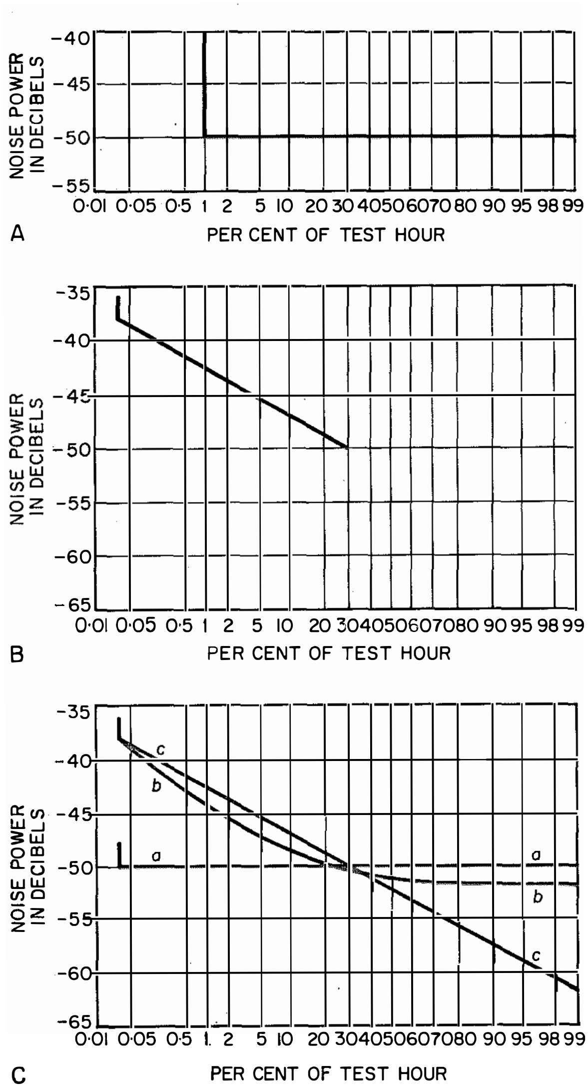

Figure 1—Noise specifications for a long-distance circuit of 2500 kilometres.

This is a difficult statistical problem and work is still proceeding[6] but a little can be said about two parts of the specification for a partial circuit.

(A) For the part of the curve that restricts the incidence of very-loud noise it is necessary that the frequency of occurrence in a partial circuit shall be proportionately smaller than for the full circuit, but in testing a single partial circuit, the test period may be lengthened in inverse proportion to the relative length of the partial circuit (see appendix 1).

(B) For the mean power over the test hour it is necessary that in the partial circuit this should be proportionately smaller than in the overall circuit but here it is also permissible in the case of the partial circuit to judge the mean noise power over a proportionately longer time. It would, of course, still have to be judged during times when a high noise level would be expected to occur. (This assumes that the busy hour et cetera will be the same for all regions through which the circuit passes. This will not always be true but is a safe assumption.)

It is however much-more difficult to find a proper answer for the main part of the curve which is not covered by (A) and (B). What we would like is to be able to specify the statistical properties of the noise that could be tolerated in for instance 1/9th of the total circuit which is specified in Figures 1B and 1C. The specification should be such that when 9 equipments having this performance are connected together the overall noise should just satisfy the specification in Figures 1B and 1C.

In combining the noise produced in the separate partial circuits it should be borne in mind that the variant to be summed up is power —not the logarithm of power (decibels).

For the present however it is probably necessary to "reverse" the problem and to propose a noise statistic for the partial circuit and then by computation find the noise performance that would be obtained if such partial circuits were connected in tandem to the length of 2500 kilometres and check if this meets the overall target.

1. Appendix

The lengthening of the test period for a partial circuit is justified by the following considerations.

The performance of the full circuit for a test hour may be considered to be the outcome of a repeated sampling process. The size of the sample must be proportional to the number of (equal) partial circuits required to make up the full circuit.

The population that is sampled contains the "hourly performance of the partial circuits" (for a period of the day when high noise is expected).

When only a single partial circuit is available and it is permissible to assume that further partial circuits will be similar, than the requisite size of sample may be taken from the hourly performance population of this circuit, or in other words, the test period for a partial circuit should be proportionally longer than the test period for a full circuit.

The individuals in a sample should be combined as if each represented the performance of one of the transmission sections that together make up a full circuit.

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Wide-Band Ultra-High-Frequency Over-the-Horizon Equipment*

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WIDE-BAND over-the-horizon radio equipment in the 680-to-900-megacycle-per-second band suitable for use in toll-quality multichannel telephone or television circuits is described. The system comprises 60-foot (18.3-meter) parabolic antennas fed with dual polarization horns, 10-watt drivers, 10-kilowatt amplifiers using 6-cavity klystrons, and receivers that permit dual or quadruple diversity by combining the received signals at the intermediate frequency. The over-all 1-decibel bandwidth is 15 megacycles, and the time delay distortion characteristics are suitable for interconnection with existing toll-quality radio links.

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The broad-band tropospheric communication system described can transmit between 120 and 600 telephone channels with trunk-circuit quality or a television program, up to at least 200 miles. The equipment also interconnects with common-carrier telephone networks.

For the 185-mile (298-kilometer) link between Florida and Cuba, with a path loss of 197 decibels, this equipment, employing space diversity, was expected to provide a mean carrier-to-noise ratio of about 42 decibels and a received-signal level above the frequency-modulation threshold at least 99.9 percent of the time. With 120 telephone channels using frequency-division multiplex, the noise on each channel was expected to be less than 26 decibels adjusted at the −9-decibels-below-1-milliwatt point 99.9 percent of the time. Actual measurements showed the channel noise level to be about 22 decibels adjusted at zero level. There were definite indications that the terminal equipment rather than the tropospheric scatter equipment contributed largely to the channel noise.

Quadruple diversity, improving performance by about 3 decibels, can be obtained by combining space and frequency diversities. This combination is possible if an active stand-by system is operated on separate carrier frequencies. The system can also be used as two dual-diversity systems, permitting preventive maintenance on one system while the other is in service. It is also possible to transmit messages and television simultaneously, each on a dual-diversity reception basis.

This over-the-horizon equipment is terminated at the 70-megacycle intermediate frequency, at which frequency the diversity combining takes place. It is thus possible to connect to either line-of-sight or over-the-horizon repeaters at intermediate instead of base-band frequency. Table 1

TABLE 1
SYSTEM SPECIFICATIONS

Features	Requirements
Transmitter Power	10 kilowatts
Antennas	60-foot (18.3-meter) paraboloidal reflectors (two at each end) with dual-polarization horns
Diversity	Quadruple (space and frequency); may be used as two dual diversity systems; intermediate-frequency (70-megacycle) combining used
Service	System input and output 70-megacycle frequency-modulation signal; capable of handling at least 120 telephone channels; can also be used simultaneously for television transmission
Frequency Band	680–900 megacycles
Transmission Lines	*WR-1150* waveguide for transmitting line, pressurized; $3\frac{1}{8}$-inch (7.94-centimeter) Styroflex cable for receiving line, pressurized
Bandwidth	15 megacycles, 1 decibel down; 19 megacycles 3 decibels down: with suitable terminal equipment, 6 megacycles, 1 decibel down, video-to-video

* Reprinted from *Communication and Electronics*, number 35, pages 86–93; March, 1958. Presented, American Institute of Electrical Engineers' Winter General Meeting; February 2–7, 1958.

TABLE 1—*Continued*

Features	Requirements
Envelope Delay Distortion	Suitable for multihop, multichannel telephone or wide-band television transmission
Noise Loading Distortion Products	More than 50 decibels below signal level
Safety Features	All equipment fully interlocked and protected; key interlocks used in power amplifier
Alarm and Metering Circuits	All equipment fully equipped with alarm circuits and contacts for external alarms; metering and test points incorporated for alignment and maintenance
Environmental	Continuous operation in tropical climates

lists the specifications for the system, and Figure 1 shows its major elements and the switching facility provided.

The first application of this broad-band equipment is the microwave communication system between Miami and Havana jointly sponsored by American Telephone and Telegraph Company and International Telephone and Telegraph Corporation. The system includes two *TD-2* line-of-sight hops in Florida between Miami and Florida City where it connects with the over-the-horizon hop to Guanabo, Cuba. A *TD-2* line-of-sight hop completes the system from Guanabo to Havana. This system supplements the present cable system between Florida and Havana to provide many more message channels. It also provides a two-way television channel between the United States and Cuba.

1. *Converter–Amplifier*

The converter-amplifier accepts a 70-megacycle frequency-modulation signal, converts it to frequencies in the 692-to-880-megacycle band and amplifies it to provide sufficient driving power for the 10-kilowatt klystron power amplifier. Specifications are as shown in Table 2.

A block diagram of the converter–amplifier is shown in Figure 2. Local-oscillator frequencies are selected in the range from 51.8 to 59.4 megacycles to minimize spurious responses in the 70-megacycle band in the intermediate-frequency

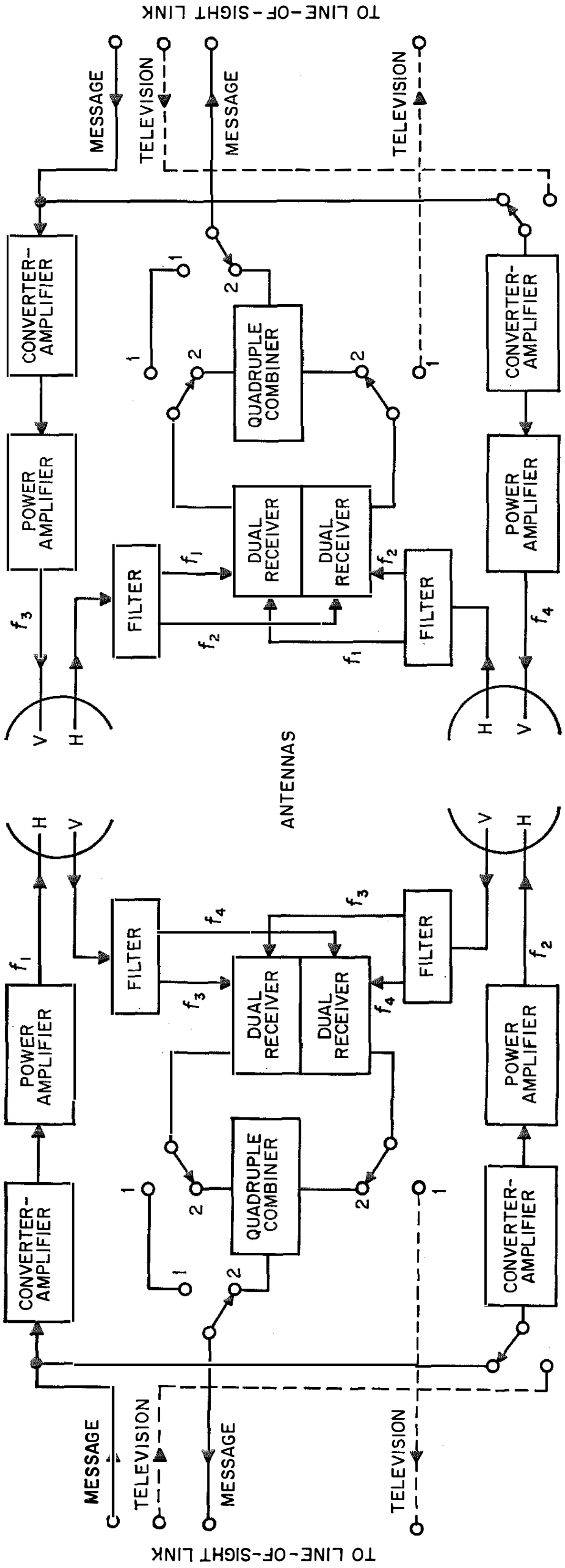

Figure 1—Block diagram of complete wide-band over-the-horizon system showing major components.

arm of the mixer. Local oscillator frequencies are multiplied and amplified by the multiplier chain and applied to the mixer at a fixed level of approximately 4 watts. The 70-megacycle input frequencies are amplified and applied to the mixer at an adjustable level of approximately 0.5 watt. The beat frequency obtained in the mixer is then selected by tuned circuits and amplified to the required power level.

The stability of the output frequency of this system is a function of the stability of the local oscillator and of the incoming 70-megacycle signal. To reduce the total carrier frequency drift to a value acceptable to the system, the local oscillator has been designed for a total frequency change of one part per million per day. The multiplied local-oscillator frequency is above that of the carrier for output frequencies of 840 and 880 megacycles and below that of the carrier for 692 and 740 megacycles.

The converter–amplifier consists of four 19-inch (48.2-centimeter), rack-mounted chassis in a 7-foot (2.1-meter) cabinet with front and rear access doors, as shown in Figure 3. The top half contains the radio-frequency chassis, with the

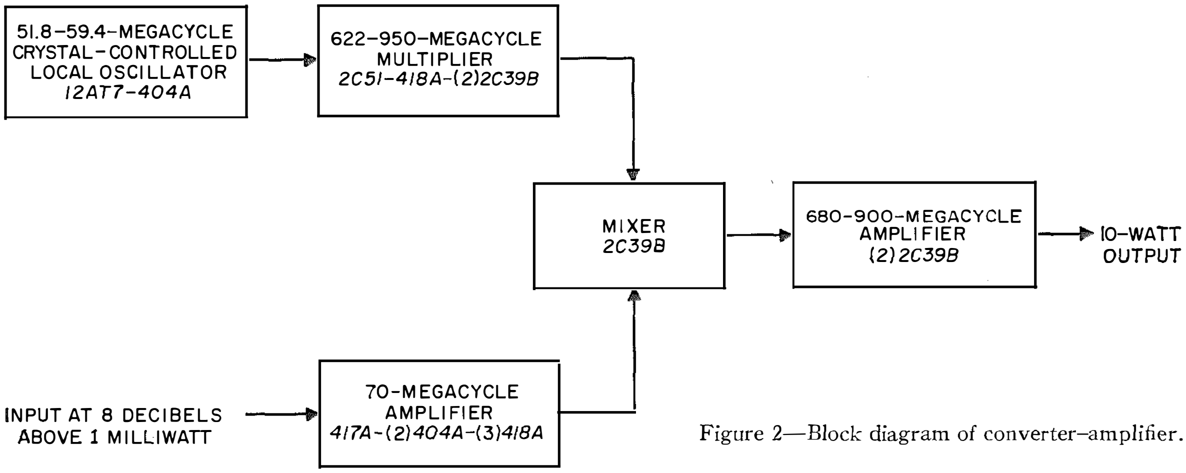

Figure 2—Block diagram of converter–amplifier.

TABLE 2

Specifications for Converter–Amplifier

Features	Requirements
Input	
Power	+8 decibels above 1 milliwatt minimum
Frequency	70 megacycles
Bandwidth	20 megacycles
Impedance	75 ohms
Output	
Power	10 watts
Frequency	680–900 megacycles
Bandwidth	20 megacycles at $\frac{1}{4}$-decibel points
Impedance	50 ohms
Envelope Delay	<4 nanoseconds* for ±4 megacycles

* Nanosecond = 10^{-9} second.

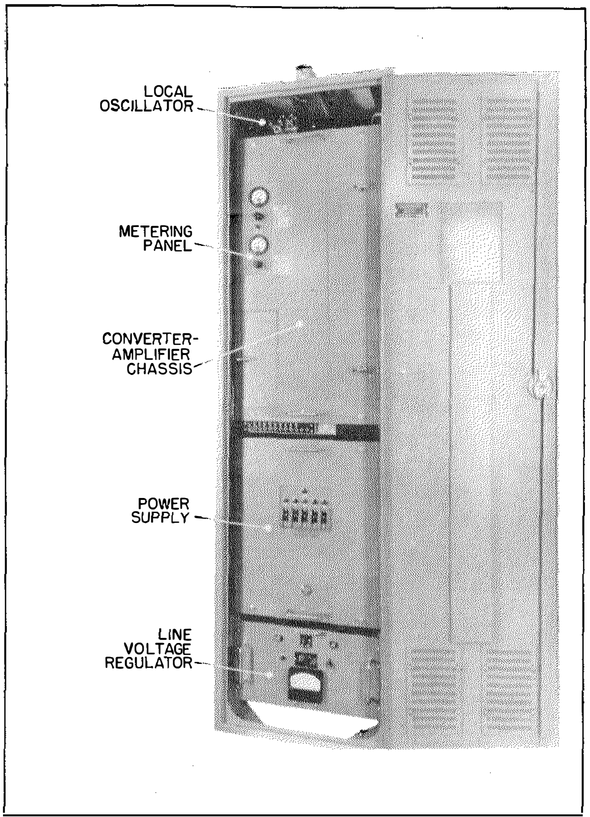

Figure 3—Front view of converter–amplifier.

power supply and voltage regulator in the lower half. The operating facilities on the front that are accessible without interrupting the interlocks are: the metering panel, tuning of local oscillator and multipliers, and tuning of mixer amplifier. The last two are available through doors in the protective dust cover. Power control and circuit breakers and a line-voltage meter are available on the front panel. A ventilating fan, with replaceable filter, is mounted on the inside of the rear door. The rear door is interlocked with the high-voltage-controlling relays and when opened the high voltage is turned off.

1.1 Radio-Frequency Chassis

The radio-frequency chassis, Figure 4, in addition to providing a mounting surface for the detachable intermediate-frequency power amplifier and radio-frequency multiplier subchassis, also contains the mixer-amplifier chain, metering facilities, monitoring circuits, fan and manifold, interlocks, and alarm circuits.

Three types of metering and monitoring circuits are available.

A. Two panel meters to monitor certain points that indicate satisfactory operation of the equipment.

B. Test points, on a jack panel accessible without turning the equipment off, with the proper test equipment to determine transmission characteristics of the equipment.

C. Each vacuum tube is connected to a test jack for measuring the cathode current by one of the panel meters.

The ventilation system is on the rear of the chassis and consists of a 115-cubic-foot- (3.3-cubic-meter-) per-minute blower with a distribution manifold and individual air ducts leading to each tube to be cooled. Failure of the blower is

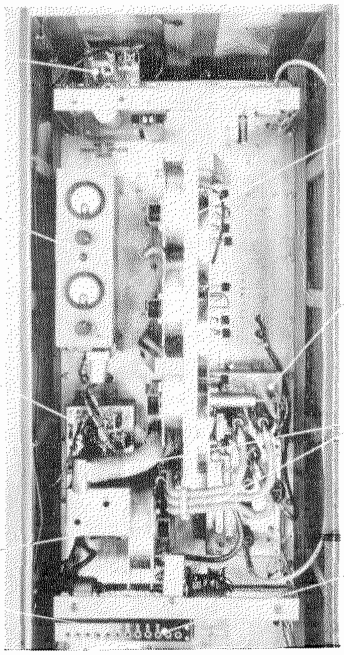

Figure 4—Detailed front view of radio-frequency chassis, converter–amplifier.

indicated by a relay and alarm buzzer activated by an air-flow switch, as well as by a lamp on top of the cabinet.

Two devices are used for protection against exposure to high voltage. The first device employs interlock switches in the front dust covers and rear cabinet access door. The second employs gravity-operated short-circuiting bars, operated by removal of the front dust covers and the rear access door.

1.2 Stable Oscillator

The oscillator consists of a stable crystal-controlled oscillator and a buffer amplifier. The tubes are one *12AT7* and one *404A*. The crystal is temperature-controlled and operates in its fifth overtone from 51.8 to 59.4 megacycles with a stability of 1.0 part per million per day. The output is about 2 volts into a 90-ohm load.

1.3 Radio-Frequency Multiplier

A chain of four tubes multiplies the frequency received from the local oscillator and amplifies the power to a level sufficient to drive the high-level mixer. The multiplication factors are 12 for output frequencies of 622 to 670 megacycles and 16 for output frequencies of 910 to 950 megacycles. The tubes used are one *2C51/396A*, one *418A*, and two *2C39B*'s. The relative power levels and frequency distributions along the multiplier chain are shown in Figure 5. One volt across the 90-ohm input impedance is required for a 3-watt output.

The coupling circuits used between the first three tubes are conventional lumped-constant tuned circuits with a radial transmission-line cavity terminating the fourth tube. An adjustable coupling loop in this cavity couples the power to the input of the mixer.

The envelope temperature is kept below approximately 100 degrees centigrade by air cooling on all tubes except the *2C51* and the total power dissipation in any tube is below 50 percent of the rated dissipation for the tube.

1.4 Intermediate-Frequency Power Amplifier

The intermediate-frequency amplifier is a broad-band 6-tube 5-stage unit that amplifies the 70-megacycle signal. It will deliver an output power adjustable between 0.1 to 1.5 watts with an input signal of 7 milliwatts across 75 ohms. The tubes used are one *417A* grounded-grid input stage, two *404A*'s and one *418A* interstage, and two *418A*'s in a push-pull output stage; see Figure 6.

The input coupling network is provided with a matching adjustment consisting of bias control of the input tube and a single-tuned broad-band cathode inductance. A matching adjustment is necessary because the input to this amplifier may be from a relatively long transmission line, and also to minimize distortion due to reflections.

Most interstage coupling networks are adjustable equivalent-T transformers with a passband amplitude response of 0.1 decibel over a 20-megacycle band centered at 70 megacycles. Exceptions are the coupling between the fourth and fifth stages and the output transformer, where fixed coupled transformers are used.

Three methods are used to minimize variations in amplifier gain from tube aging and supply voltage fluctuations. These are: (A) Regulation

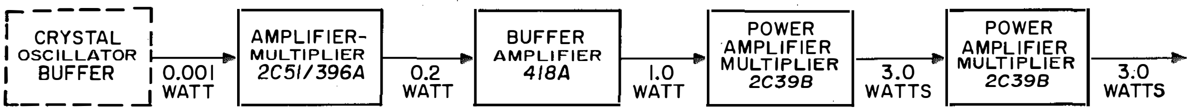

Figure 5—Block diagram of radio-frequency multiplier in converter–amplifier. Total frequency multiplication is 12 for the low band and 16 for the high band. Oscillator frequencies are 51.833 to 55.833 megacycles for the low band and 56.875 to 59.375 megacycles for the high band.

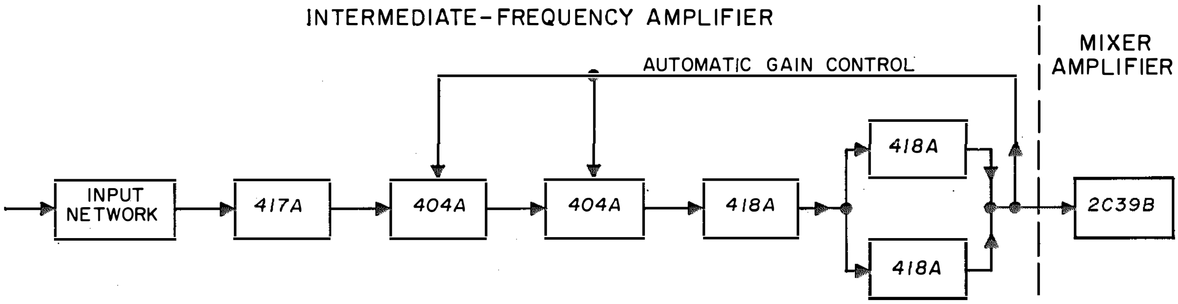

Figure 6—Block diagram of intermediate-frequency amplifier in converter–amplifier.

of the filament power supplies by a 12-volt filament supply and appropriate dropping resistors. (B) Positive grid bias and large cathode resistors. (C) Automatic-gain-control feedback from the intermediate-frequency output to the grids of the second and third stages.

1.5 Mixer–Amplifier

The mixer–amplifier consists of three *2C39B*'s in double-tuned radial transmission line cavity-coupling networks. The required bandwidth is obtained by using stagger-damped interstage coupling and single-tuned broad-band output coupling, giving an over-all 3-peaked response having a peak-to-valley ratio of approximately $\frac{1}{4}$ decibel over a 20-megacycle bandwidth.

The mixer–amplifier, Figure 4, consists of a mixer, two ultra-high-frequency amplifiers, and radial transmission line cavities. Each tube has one cathode cavity and one plate cavity. Broad bandwidth is obtained by adjusting the coupling between the plate cavity of one stage and the cathode cavity of the following stage. The local-oscillator power is coupled to the cathode of the mixer tube by a radial cavity, and the intermediate-frequency power is introduced through a broad-band lumped-constant network. The plate cavity of the mixer is loop-coupled to the cathode cavity of the first amplifier and proper coupling is obtained by the relative values of two tuning capacitors in the amplifier cathode cavity.

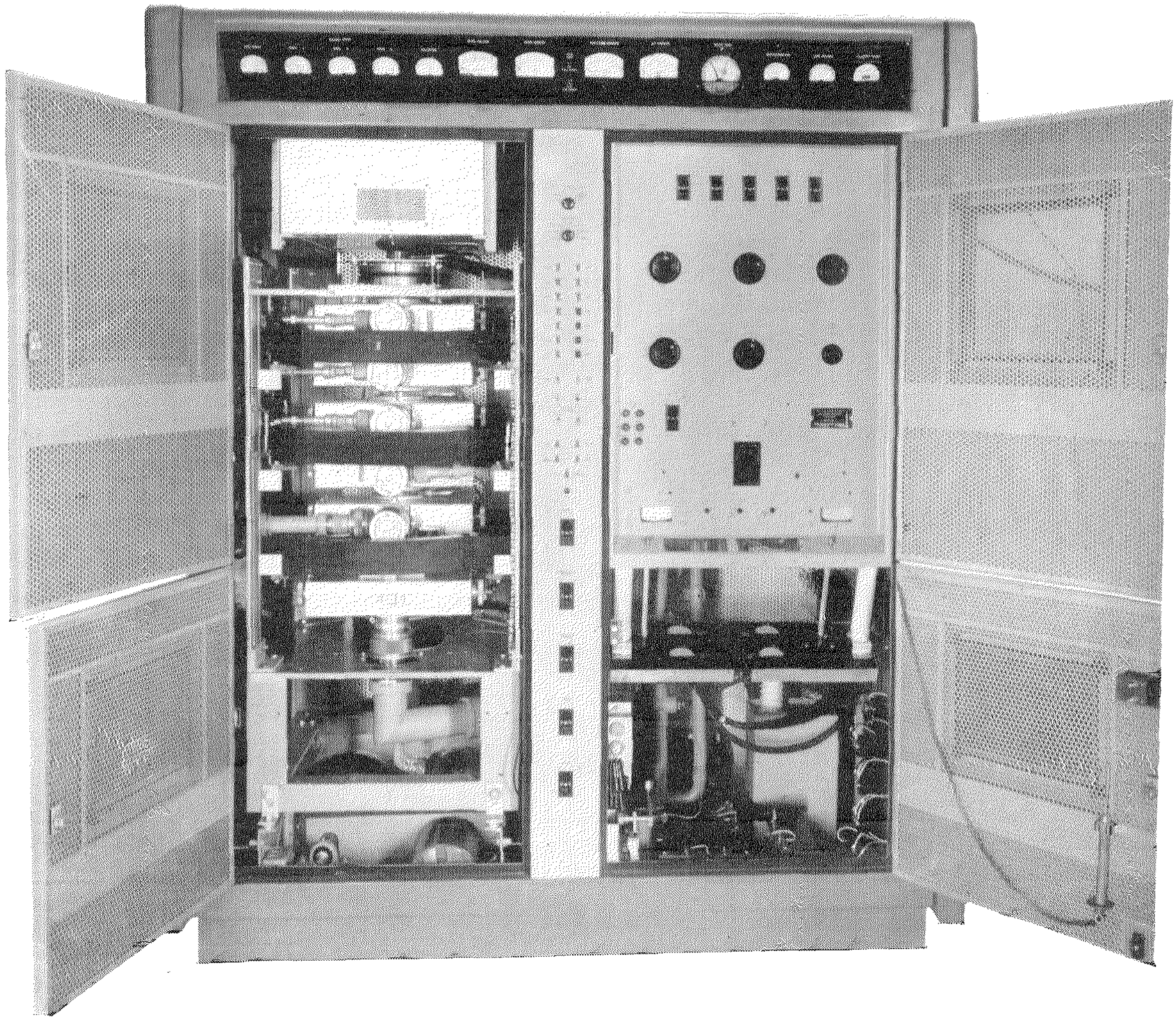

Figure 7—Front view, 10-kilowatt power amplifier.

A similar coupling arrangement is used between the first and second amplifiers. The second amplifier plate circuit is a single tuned cavity, loaded to the required Q by the 50-ohm load on the output loop. Capacitive-tuning plungers, inserted parallel to the axis, tune each cavity.

1.6 Power Supply

The power supply is nonregulated and supplies all direct and alternating voltages. Voltage regulation of ± 1 percent is obtained from an automatic line-voltage regulator mounted below the power supply.

A 60-second time delay between the application of filament power and plate power allows for filament warm-up. All high voltages are connected through relays on the primary side of the transformers, which are operated by the interlocks on the rear cabinet door and front dust covers.

2. *Power Amplifier*

The converter–amplifier output at approximately 10 watts is amplified to 10-kilowatt level by the power amplifier. The required 30 decibels of gain is obtained with a 6-cavity klystron (type *X631*, by Eitel-McCullough, Incorporated). This klystron and a set of tuned circuits covers the band from 680 to 900 megacycles.

2.1 Amplifier

Since the velocity modulation action at the gaps and the density modulation in the drift spaces are frequency-sensitive mechanisms, the klystron is inherently a narrow-band device. However, by utilizing broad-band tuned circuits at the klystron gaps, bandwidths of 3 to 4 percent of center frequency can be achieved.

The amplifier assembly is shown in Figure 7. The klystron assembly is on the left and the controls for the focusing-coil power supplies are on the upper right panel. The lower right-hand section houses the klystron filament transformer and the semiconductor rectifier for the focus supplies. Circuit breakers and overload indicators are aligned along the center panel and meters indicating the klystron parameters are in the top panel.

The klystron assembly, Figure 8, shows in greater detail the double-tuned circuits surrounding each klystron gap. The circuits across the first five klystron gaps consist of coaxial lines coupled to rectangular waveguide cavities. A waveguide section surrounds the klystron gap. To handle the high power level, the output circuit is formed by two waveguide cavities coupled together with an iris in the common wall between the guide sections. A lumped-circuit equivalent of these tuned circuits is shown in Figure 9.

The desired bandwidth is achieved by loading each circuit with an external resistor. In the input circuit, the waveguide section is loaded and the input signal is coupled to the coaxial section of the double-tuned assembly. Coupling and loading are adjusted to obtain maximum power transfer and maximal flatness of response (equal primary and secondary Q).[1] The interstage assemblies are loaded on one side only (the secondary coaxial section) for maximum impedance across the klystron gap at the given bandwidth.

The output primary tuned circuit is loaded by the klystron beam impedance; the secondary is loaded by the antenna impedance coupled through a capacitive probe. The loading resistors used in the first three tuned circuits are air-cooled. The fourth- and fifth-circuit resistors are $1\frac{5}{8}$-inch (4.13-centimeter) coaxial-line water-filled loads.

To simplify alignment of the amplifier, the six tuned circuits are synchronously aligned and centered about the carrier frequency. Each circuit is initially adjusted for transitional coupling with the klystron unenergized. A sweep signal generator is coupled to one circuit at a time and the response observed at the test point in the circuit. Then, under normal drive and power input conditions, a slight readjustment of the circuit parameters will result in the desired response. It should be noted that increased efficiency and gain can be obtained by "stagger damping" but at the expense of considerably greater adjustment difficulties.[2]

[1] M. J. Hellstrom, "Design Charts for Tuned Transformers," *Electronics*, volume 29, page 182; November, 1956.

[2] G. E. Valley, Jr. and H. Wallman, editors, "Vacuum Tube Amplifiers," McGraw-Hill Book Company, Incorporated, New York, New York, Massachusettes Institute of Technology Radiation Laboratory Series, volume 18; 1948: page 221.

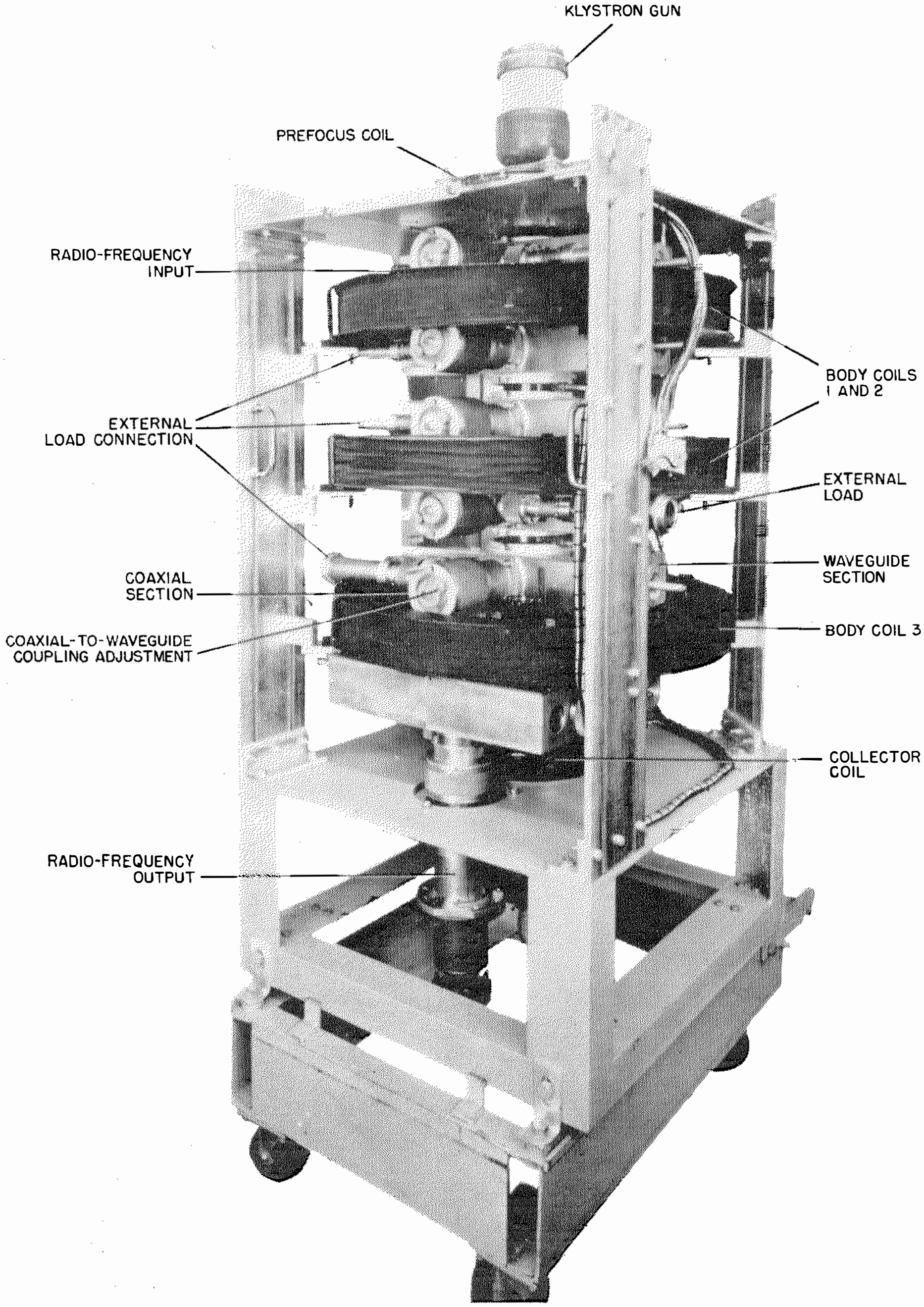

Figure 8—Klystron assembly and magnetic structure mounted on dolly.

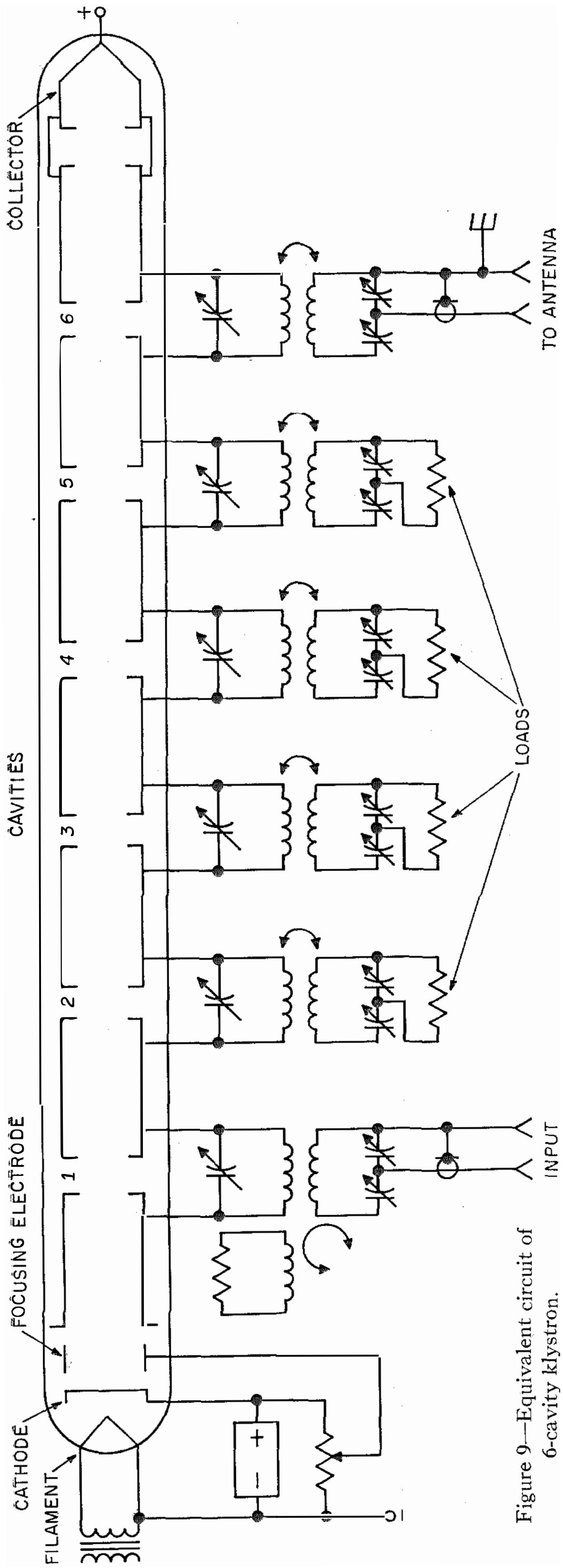

Figure 9—Equivalent circuit of 6-cavity klystron.

2.2 Power Supplies

The primary power to all filaments, to the klystron bombarder supply, and to the electromagnetic focusing supplies is regulated by an automatic motor-driven induction regulator to assure maximum tube life and minimum variation in klystron operating conditions.

All high-voltage magnetic components are nonflammable-oil-filled. The bombarder supply and beam supply are unitized assemblies; the magnetic components are contained in one oil-filled tank.

The bombarder supply utilizes xenon-filled rectifiers, whereas the beam supply has high-vacuum diodes to minimize the possibility of rectifier arcback. The circuits of both supplies are the 3-phase bridge type. The beam supply delivers 20 kilovolts at 2.5 amperes. Filtering of this unit is adequate to limit the incidental frequency modulation to less than 100 root-mean-square cycles per second. Stepless control of the beam voltage is possible at the amplifier control panel, over the range from 10.5 to 20 kilovolts at the nominal line voltage ±10 percent.

2.3 Controls

Control and overload monitoring circuits provide proper sequential application of the various voltages and protect the equipment against progressive damage resulting from component failures or overloads.

All power-line circuits are monitored and switched by magnetic-trip circuit breakers with an appropriate tripping-time-versus-current characteristic that will accommodate starting current surges but also allow rapid tripping under large overload conditions.

High-speed direct-current relays monitor the following parameters: beam current, drift-tube current, fifth- and sixth-cavity voltage, excessive output transmission line mismatch, and low transmitter output. All these relays except the low-output relay will trip the beam power supply circuit breaker in less than 50 milliseconds. Pilot-light indications and an alarm call attention to a fault. In addition to electrical interlocks, the necessary mechanical interlocks such as liquid coolant flow, air flow, and high-voltage grounding switches are incorporated.

2.4 Cooling

Since the collector and drift tubes require liquid cooling, a closed system (coolant-to-air or coolant-to-water unit) is provided for each amplifier. Where an adequate supply of well water is available, the latter type of cooler is preferred. The air-cooled unit is shown in Figure 10 and the water-cooled unit in Figure 11. The coolant, a mixture of ethylene glycol and distilled water (60 to 40 parts per volume) with an acidic inhibitor, is also used as the dissipative element in the klystron fourth- and fifth-cavity loads.

2.5 Performance

Over-all transmission response of the converter–amplifier plus power amplifier is shown in Figure 12; the oscillogram shown was taken at 840 megacycles. Note that the 3-decibel bandwidth is 19 megacycles, and the 1.0-decibel points approximately 15 megacycles. Saturation power is 10.5 kilowatts at the upper end of the band (880 megacycles) and it falls to 8 kilowatts at the low end of the band (690 megacycles).

Gain at the level 0.5 decibel below saturation power is 32 decibels at the high-frequency end of the band and 28 decibels at the low end for the bandwidths shown. For smaller bandwidths, the gain can be increased to as much as 48 to 50 decibels before instability problems arise.

Envelope delay versus input frequency of the system, including the wide-band terminal equipment, converter–amplifier, power amplifier, and receiver is shown in Figure 13. The envelope delay curve is resolved into three components to obtain a standard basis of comparison; the slope component is found by drawing a straight line between the frequency points of interest, such as 66 megacycles and 74 megacycles of Figure 13; the difference in time delay noted between the two points is the slope. The change in envelope delay remaining after subtracting the slope component is the parabolic component, provided ripple is negligible.

If appreciable ripple is present, this must be averaged about the slope component and subtracted from the total envelope delay; the remainder is the parabolic component. Thus, the

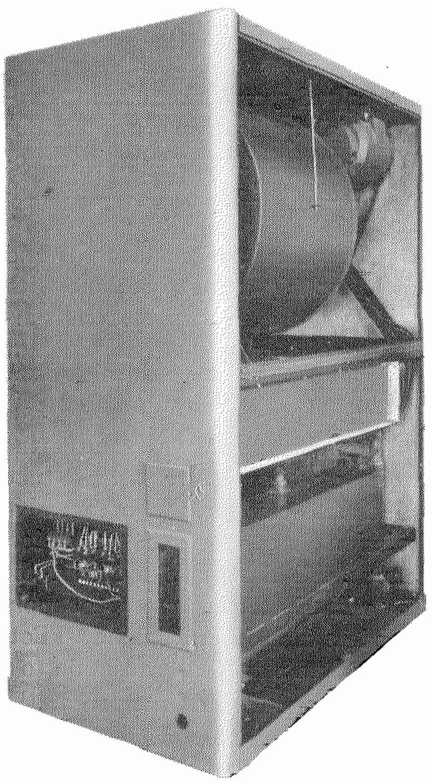

Figure 10—Heat exchanger for coolant to air.

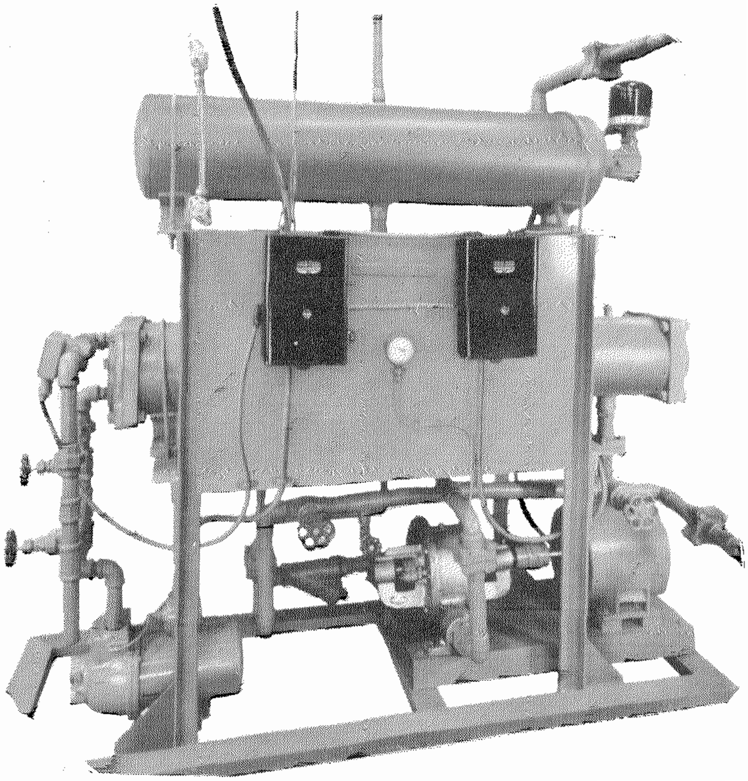

Figure 11—Heat exchanger for coolant to water.

sum of the three components at each frequency should equal the recorded curve. The components of time delay are given in Table 3.

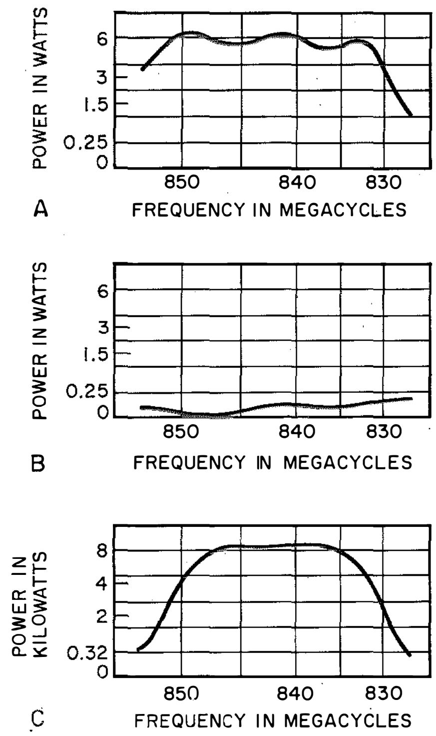

Figure 12—Amplitude response versus frequency. *A* = converter–amplifier. *B* = reflected from klystron input. *C* = 10-kilowatt amplifier driven by converter–amplifier.

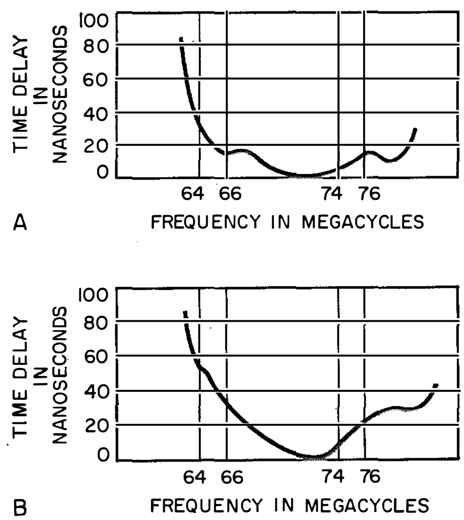

Figure 13—Envelope delay versus frequency. *A* = modulator, demodulator, and converter–amplifier. *B* = modulator, demodulator, converter–amplifier, and 10-kilowatt power amplifier.

Over-all response from intermediate-frequency input to the converter–amplifier to intermediate-frequency output at the receiver is shown in Figure 14 and the video-to-video response from the transmitting terminal through the radio system and the receiving terminal is shown in Figure 15. Intermodulation distortion is 50 decibels down, as measured through the system with a 500-kilocycle-bandwidth noise source producing a root-mean-square deviation of 0.4 megacycle.

TABLE 3

TIME DELAY COMPONENTS

Component	Time Delay in Nanoseconds	
	At ±4 Megacycles	At ±6 Megacycles
Parabolic	15	22
Slope	25	30
Ripple	±2	±2

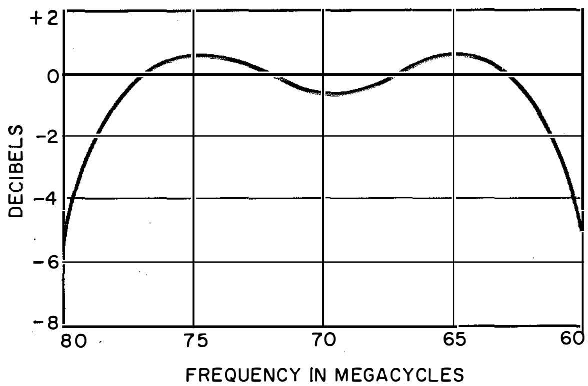

Figure 14—Over-all amplitude response versus frequency through converter–amplifier, power amplifier, and receiver.

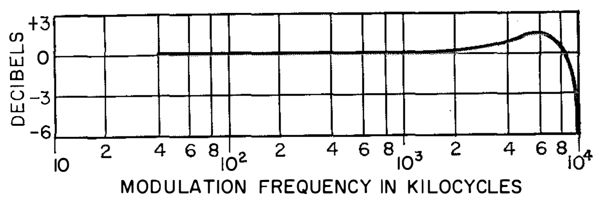

Figure 15—Over-all video-to-video amplitude response, including wide-band modulator and demodulator.

3. *Antenna System*

The antenna system consists of two 60-foot (18.3-meter) paraboloidal reflectors at each site, fed by dual-polarization horns, each of which is connected to one transmitter and two receivers. The transmission line connecting the transmitters to the antennas is *WR-1150* waveguide, and the receiver connection is by $3\frac{1}{8}$-inch (7.94-centimeter) 50-ohm Styroflex cable.

3.1 Reflectors

The reflectors and horn-support structures are shown in Figure 16.

3.2 Transmission Lines

WR-1150 copper-clad steel waveguide is employed for the high-power connection between the horn and the transmitter. The waveguide runs include both rigid bends and flexible waveguide bends at appropriate points on the tower and horizontal lines. Waveguide is used because of its low loss and its high power-handling capacity.

The receiver connection is made through $3\frac{1}{8}$-inch (7.94-centimeter), 50-ohm polyethylene-jacketed Styroflex cable. This large cable is used primarily because of its low loss and lower cost compared with waveguide.

3.3 Horn

The use of dual orthogonally polarized horns (greater than 50-decibel decoupling between polarizations) permits a wide variety of diversity system arrangements to be employed. In particular, the polarizations are selected in such a way that no high-power diplexers are required in the transmitter lines. This is accomplished by transmitting one frequency and receiving two frequencies at each horn. Received signals are separated by the receiver preselecting filters arranged as a diplexer.

The dual polarized horns for the 60-foot (18.3-meter) paraboloidal reflectors were designed to direct a signal at the reflector edge at approximately −10 to −12 decibels relative to the signal directed toward the reflector vertex.[3] Typical data are given in Figure 17. Since the horns were designed for a specific application, no particular effort was directed toward broad-band development.

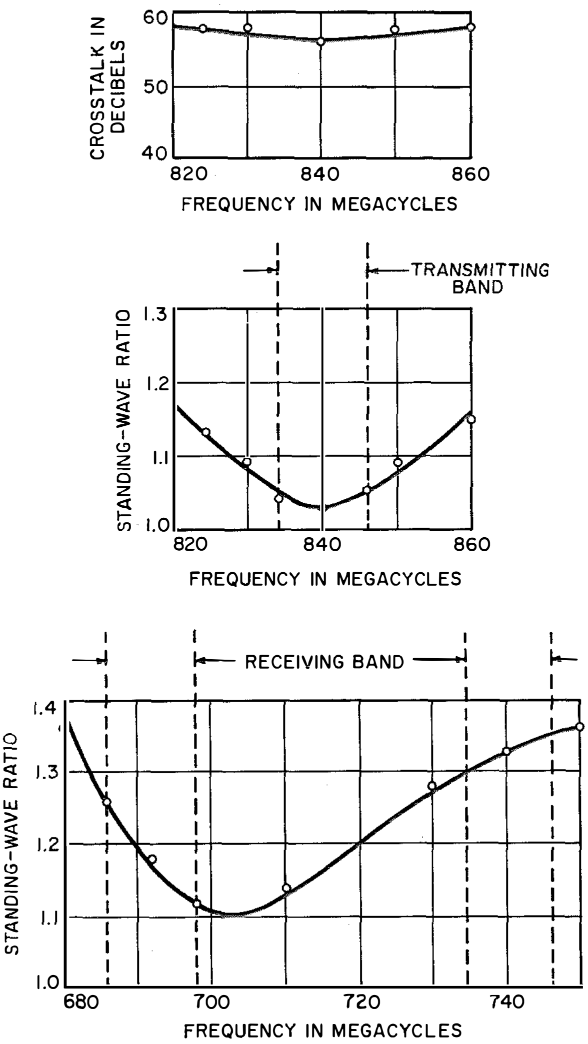

Figure 17—Standing-wave ratio and crosstalk of a typical dual-polarization horn.

An outline drawing of the horn is shown in Figure 18. Short capacitive posts match the horn at the individual transmitter frequency bands. This was necessary to obtain the required low voltage standing-wave ratio, less than 1.08, over the transmitter band. The receiver connections

Figure 16—On the facing page is shown a 60-foot (18.3-meter) paraboloidal antenna at Guanabo, Cuba.

[3] D. J. LeVine and W. Sichak, "Dual-Mode Feed Horn for Microwave Multiplexing," *Electronics*, volume 27, page 162; September, 1954.

are matched by proper selection of probe length and diameter and by the fin position relative to the probes.

The horns have been satisfactorily tested with 10-kilowatt continuous-wave power applied to the waveguide connection. The transmission lines and horns operate with approximately 0.5 pound-per-square-inch (35-grams-per-square-centimeter) (gage) internal dry air pressure and have been tested with over 2 pounds (140 grams) internal pressure.

The horn aperture barrier is a gasketed $\frac{1}{8}$-inch (3.2-millimeter) thick teflon-fiberglass sheet clamped to the aperture flange. Figure 19 shows a completed horn, ready for installation. The horn weighs approximately 150 pounds (68 kilograms). The maximum dimensions are approximately 14 by 14 by 46 inches (36 by 36 by 117 centimeters).

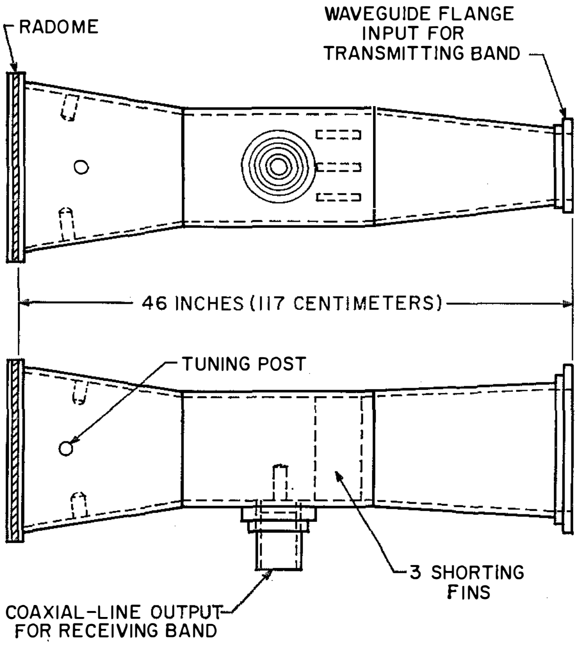

Figure 18—Outline drawing of dual-polarization horn.

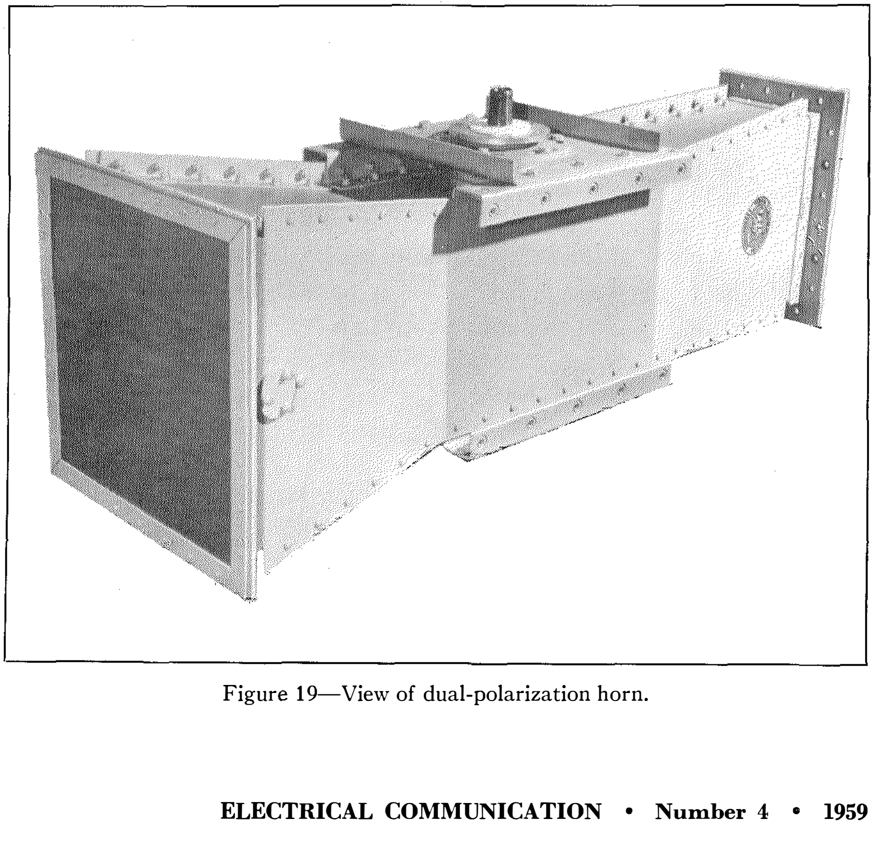
Figure 19—View of dual-polarization horn.

The antenna system performance can be described as follows: The free-space gain at midband is 41 decibels, not including line loss, and the midband half-power beamwidth is 1.4 degrees.

4. *Diversity Receiver*

The quadruple-diversity receiver set comprises two dual-diversity receiver sets operating at different frequencies in the range 680 to 900 megacycles with means for combining at the 70-megacycle intermediate frequency into one 70-megacycle output.

Figure 20 shows the receiver with dust covers removed. Two 9-foot (2.74-meter) racks (joined together at the center) hold the various 19-inch (48.3-centimeter) subchassis panels. Available components such as intermediate-frequency amplifiers, equalizers, and receiver control units, are used wherever possible.[4] Preselectors, mixers, local-oscillator–multipliers, combiners, diversity switching panel, and power distribution and alarm panels are designed specifically for this system.

[4] A. A. Roetken, K. D. Smith, and R. W. Friis, "TD-2 Microwave Radio Relay System," *Bell System Technical Journal*, volume 30, pages 1041–1077; October, 1951.

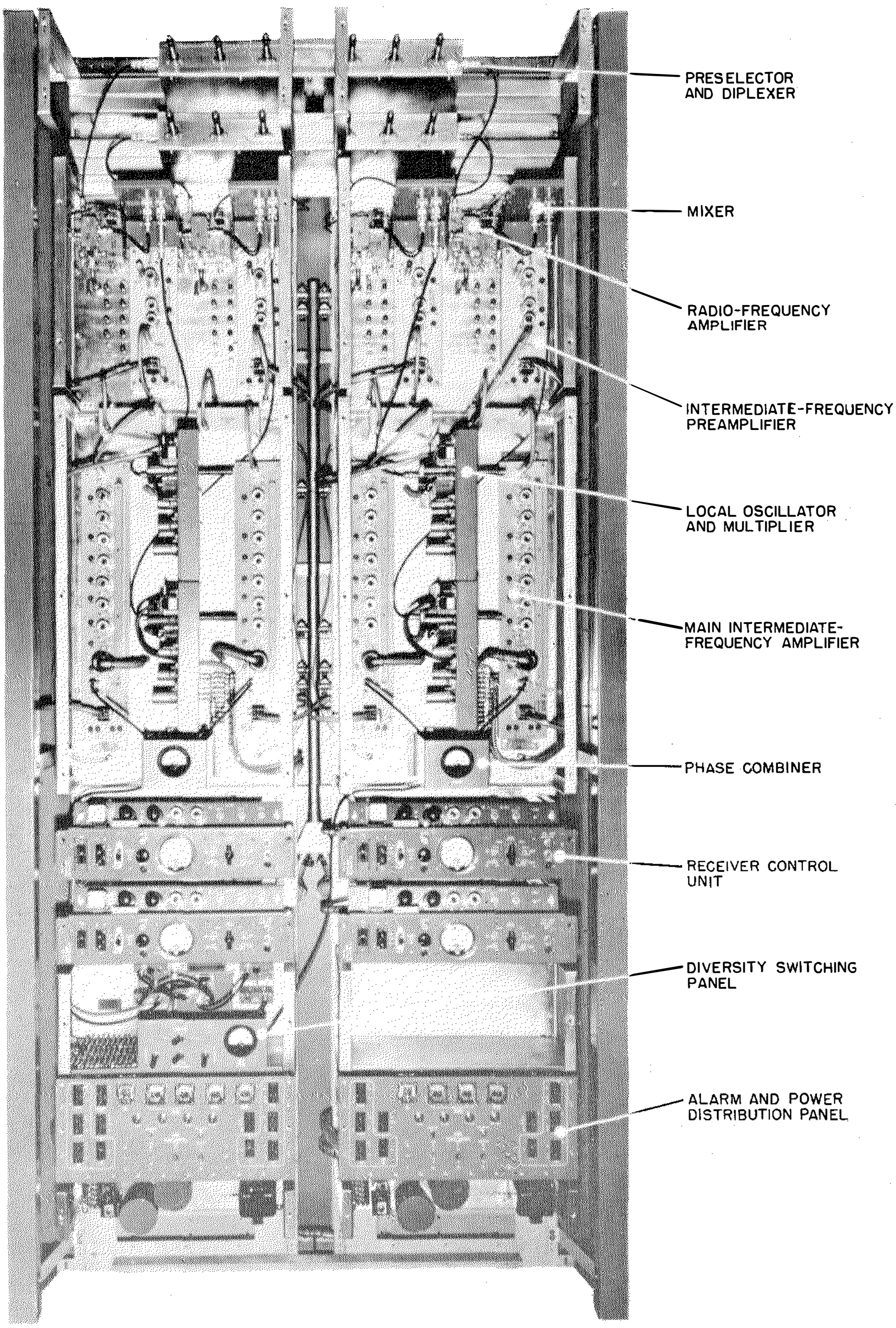

Figure 20—Front view of quadruple-diversity receiver.

Figure 21 is a simplified receiver block diagram. Preselectors separate the two frequencies from each antenna line for each of the four receiving channels. Each signal passes through a radio-frequency amplifier and mixes with the output of a crystal-controlled local-oscillator–multiplier to produce a 70-megacycle intermediate frequency. The intermediate-frequency preamplifiers, equalizers, attenuators, and main amplifiers deliver the proper level for phase combining. There are three phase combiners, one for each dual-diversity channel, and one to combine the dual-diversity outputs for quadruple combining. To meet the over-all delay distortion and amplitude response requirements, each major component of the receiver is so designed that the 0.1-decibel bandwidth is about 20 megacycles.

4.1 Diplexer Filter

The diplexer is composed of two 3-cavity maximally flat amplitude response filters, centered at 692 and 740 megacycles, respectively, (or 840 and 880 megacycles), coupled to a T junction. The branch legs to each filter are a quarter wavelength at the center frequency of the other filter so that short circuits at the filter inputs are reflected as open circuits at the diplexer junctions. The correct leg is a matched line (standing-wave ratio under 1.1). The filter specifications are given in Table 4.

4.2 Radio-Frequency Amplifier

A single *416B* tube is used in a grounded-grid amplifier stage. Figure 22 is a top view. The

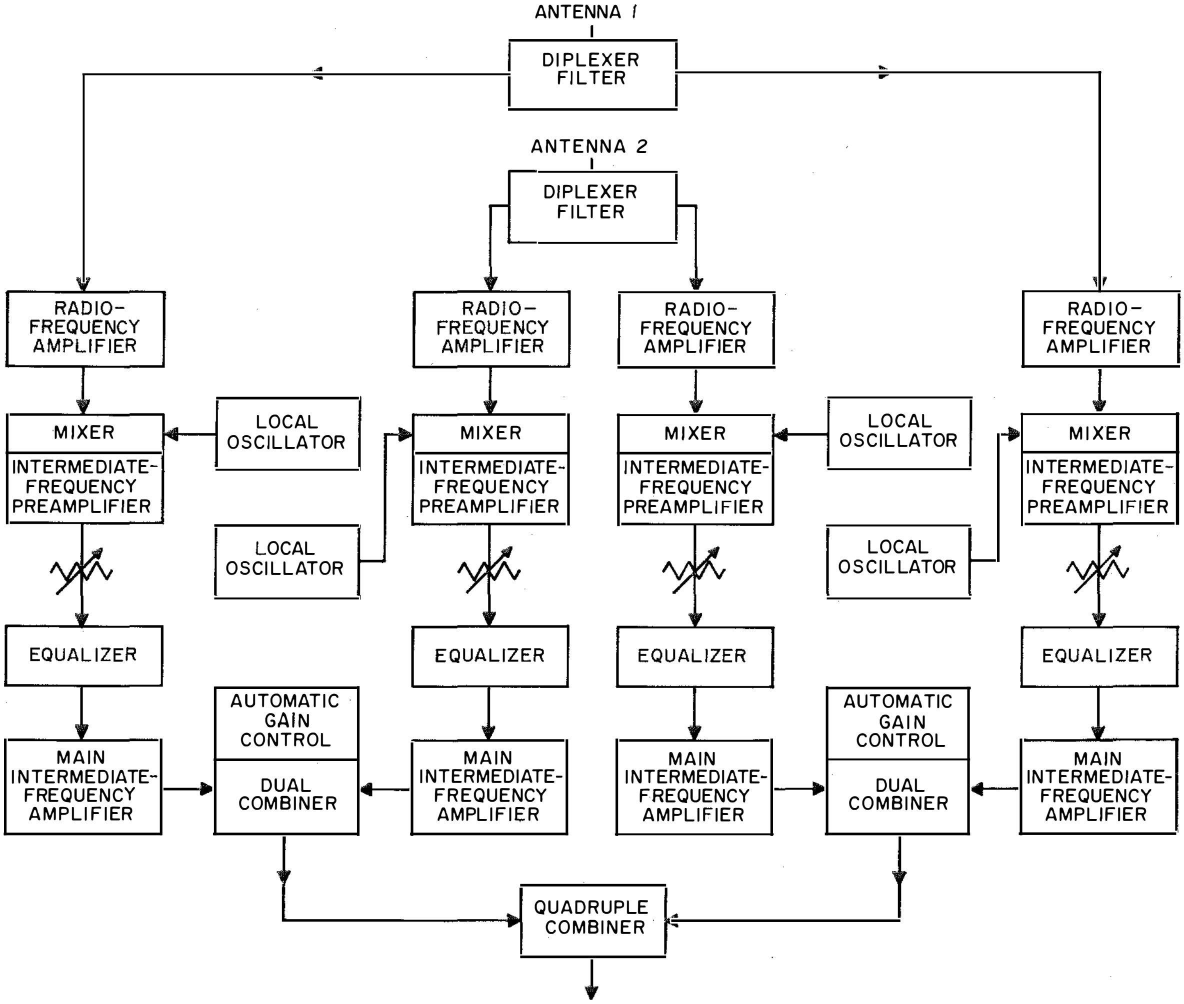

Figure 21—Block diagram of major receiver components.

input standing-wave ratio is less than 1.3 over a 20-megacycle bandwidth. The plate circuit has two resonant transmission lines provided with adjustable coupling and an output impedance of

TABLE 4

RECEIVING FILTERS

	Frequency in Megacycles	Decibels Bandwidth	Insertion Loss in Decibels
Center Frequency	18.5	0.1	0.5
Other Received Frequency	30	1.0	30
Transmitter Frequencies	38	3.0	50
	175	40	

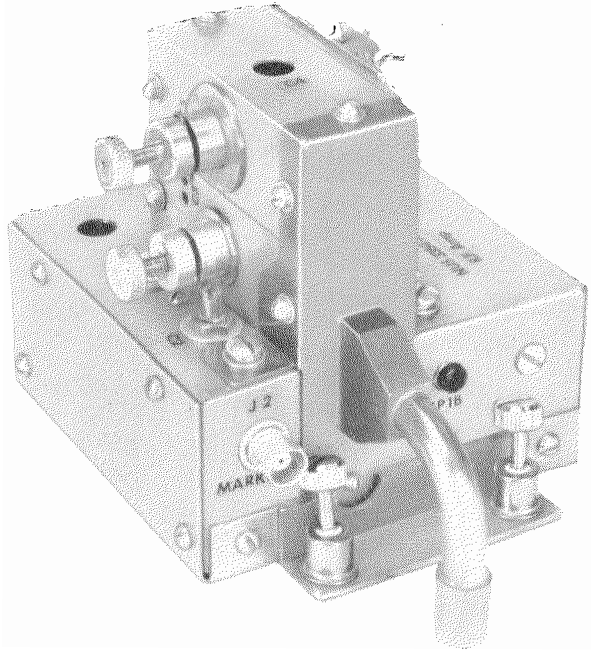

Figure 22—Low-noise radio-frequency amplifier.

50 ohms. Gains of 14 decibels and noise figures under 8 decibels are obtained.

4.3 Mixer Preamplifier

Microstrip construction is used in the balanced mixer.[5] The crystals (types *1N21B* and *1N21BR*) are each matched to a 50-ohm line by a short-circuited stub of proper length and position. Decoupling between the signal input and the local-oscillator arm is greater than 20 decibels.

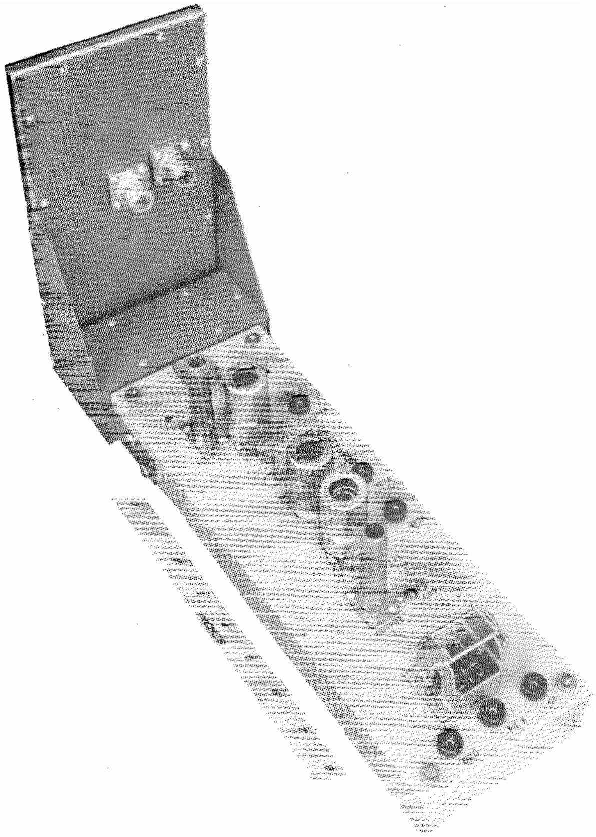

Figure 23—Microstrip mixer and 70-megacycle preamplifier.

The signal input standing-wave ratio is under 1.3 for a 20-megacycle bandwidth. The mixer is attached directly to the intermediate-frequency preamplifier, Figure 23, and drives the cathode of the first of three cascaded stages.

4.4 Main Intermediate-Frequency Amplifier

A wide-band intermediate-frequency amplifier, 8 tubes in cascade, provides 60-decibel gain at 70-megacycle center frequency.[4] The 3-decibel bandwidth is 32 megacycles and the 0.1-decibel bandwidth is 20 megacycles.

4.5 Local Oscillator–Multiplier

The crystal-controlled local oscillator, Figure 24, uses a *12AT7* twin triode in a cathode-

[5] D. D. Grieg and H. F. Engelmann, "Microstrip—A New Transmission Technique for the Kilomegacycle Range," *Proceedings of the IRE*, volume 40, pages 1644–1663; December, 1952: also, *Electrical Communication*, volume 30, pages 26–35; March, 1953.

coupled circuit at frequencies in the 50-megacycle range. The output triode section is tuned to twice the crystal frequency. Two *404*-type tubes follow in cascade, the first as a 100-megacycle amplifier and the second as 300-megacycle tripler. This drives the grounded-grid *2C39* output tube either as a doubler or tripler in the frequency range 620 to 950 megacycles.

4.6 Phase Combiner at 70 Megacycles

Diversity reception is made possible by locking two 70-megacycle intermediate-frequency signals and adding them in phase. The phase detector supplies a direct control voltage to the local oscillator, which in turn adjusts their frequencies to produce phase lock. Except for two semiconductor diodes, the combiner uses only passive circuit elements. Part of the combined output voltage is used for automatic gain control and controls the main intermediate-frequency amplifiers. The 70-megacycle outputs of the dual combiner connect to the diversity switching panel.

4.7 Diversity Switching Panel

Outputs from the dual combiners are switched by coaxial relays to separate jacks or to a third combiner for quadruple operation. Automatic-gain-control voltages are simultaneously switched for either type of service. Figure 21 shows the system diagram for a quadruple receiver.

4.8 Power Distribution and Alarm Panels

Each dual-diversity receiver is separately powered by external lines from a commercial power supply. Incoming power is controlled by master switches, circuit breakers, and fuses. The alarm circuits monitor blower operation, radio-frequency signal level, and intermediate-frequency output; 3-phase lock conditions actuate an audible alarm whenever their operation is abnormal.

5. Acknowledgments

The successful development of this equipment is due to the teamwork and co-operation of a large number of people. The authors are grateful to the following engineers for their contributions: A. G. Kandoian, for the initial planning and over-all supervision; W. Sichak, for over-all technical supervision; J. Gulack for system layout; W. Glomb, H. E. Lott, and J. Simkovich for the converter–amplifier; L. Gray, H. Goldman, and W. Jacobus for the power amplifier;

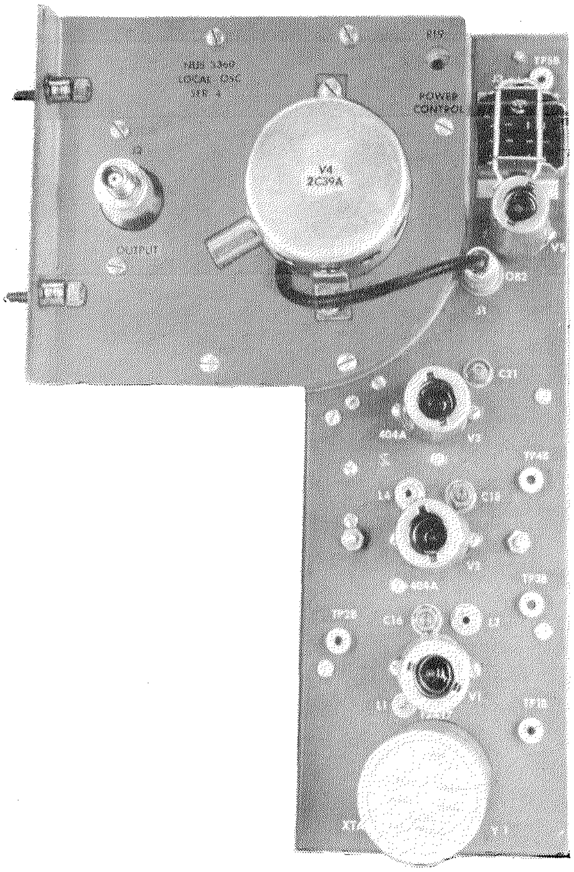

Figure 24—Crystal-controlled local oscillator.

R. T. Adams, A. T. Brown, and R. Bresk for the receiver; and L. Juhas for the feed horn.

The requirements and equipment specifications for the system were set up by a joint committee of representatives of Bell Telephone Laboratories, the Long Lines Department of American Telephone and Telegraph Company, Western Electric Company, International Telephone and Telegraph Corporation, and ITT Laboratories. Special acknowledgment is due to the following members of the committee: K. P. Stiles and A. A. Bottani of Long Lines Department, American Telephone and Telegraph Company; and H. E. Curtis, N. F. Schlaack, and H. A. Wells of Bell Telephone Laboratories.

United States Patents Issued to International Telephone and Telegraph System; May-July 1958

Between May 1, 1958 and July 31, 1958, The United States Patent Office issued 72 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

H. H. Abelew, Mackay Radio and Telegraph Company, Signal Selector Device, 2 834 003.

P. R. Adams and R. I. Colin, Federal Telecommunication Laboratories, Aircraft Radio Navigation System, 2 836 815.

R. T. Adams and R. E. Altoonian, Federal Telecommunication Laboratories, Microstrip Switch, 2 842 637.

P. R. R. Aigrain, Laboratoire Central de Telecommunications (Paris), Circuit Element Having a Negative Resistance, 2 843 765.

P. R. Aigrain, Laboratoire Central de Telecommunications (Paris), Semiconductor Crystal Rectifiers, 2 845 370.

D. F. Albanese, Federal Telecommunication Laboratories, Pulse Modulator 2 837 719.

J. L. Allison and B. Alexander, Federal Telecommunication Laboratories, Airborne Pictorial Navigation Computer, 2 836 816.

M. Arditi and J. Elefant, Federal Telecommunication Laboratories, Microwave Transmission Line, 2 833 995.

M. Arditi, Federal Telecommunication Laboratories, Traveling-Wave Electron Discharge Devices, 2 833 962.

R. J. Arndt, Kellogg Switchboard and Supply Company, Negative-Impedance Repeater Having Gain Controls, 2 844 669.

R. P. Arthur, Kellogg Switchboard and Supply Company, Electromagnetic Counting Device and Contact Bank, 2 844 686.

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T. H. Clark, Federal Telephone and Radio Corporation, Radio Direction Finder, 2 840 813.

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L. C. Pocock, Standard Telephone and Cables (London), Telephone Subscriber's Instruments, 2 838 612.

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K. Wernick, Arrangement for Telephones with an Amplifying Device Connected in Series with the Receiver Capsule, 2 844 658.

E. P. G. Wright, D. A. Weir, and G. R. Phillips, Standard Telecommunication Laboratories (London), Totalisator Equipment, 2 837 281.

E. P. G. Wright and J. Rice, Standard Telecommunication Laboratories (London), Methods of Recording and/or Modifying Electrical Intelligence, 2 838 745.

H. O. Wolcott, Frequency-Marker-Pulse Circuits, 2 845 533.

S. Yasuda, Nippon Electric Company (Tokyo), Electron-Tube Magnetic Focusing Device, 2 843 775.

Methods of Recording and/or Modifying Electrical Intelligence

2 838 745

E. P. G. Wright and J. Rice

A data-processing system is described that includes a magnetic device on which information is stored in binary code on a plurality of tracks. The system provides an arrangement for reading the information from the storage device successively for the different items recorded, modifying this information in accordance with changes that should be made, and then re-recording the modified item in the same track as that from which it was originally read. The control of the system is effected by various control pulses operating through a system of coincidence gates and flip-flop circuits.

Omnirange-Beacon Antenna

2 836 820

S. Pickles, P. R. Adams, and C. Lucanera

An omnidirectional-beacon antenna for use in tacan has radiators in a vertical array to reduce the vertical angle of transmission. With each radiator, there are two sets of reflectors mounted on an insulating drum for modifying the pattern radiated from the antennas to produce a multilobe directive beacon pattern.

Aircraft Radio Navigation System

2 836 815

P. R. Adams and R. I. Colin

A radio beacon system producing a rotating directive pattern, such as is used in tacan, is covered. It provides for the transmission on this rotating pattern of reference and bearing signals, mobile-unit-identifying signal pulses, and message signal pulses together with controls for applying these various types of pulses in sequential order. The beacon can be used to cooperate simultaneously with a plurality of mobile units, such as aircraft.

Telephone Subscriber's Instruments

2 838 612

L. C. Pocock

A telephone subscriber's set is described that employs a transistor amplifier to permit the use of a resistive bridge network in place of an inductance coil for reducing side-tone effects. The transistor amplifier is associated with the resistive bridge network and serves to offset losses that would otherwise occur in this bridge network. The amplifier is in conjugate relationship with both the transmitter and receiver.

Composite Radiation Amplifier

2 837 660

R. K. Orthuber and L. R. Ullery

The radiation amplifier is of the type in which light or X-ray radiation images may be amplified by control of the voltage across an electroluminescent substance through the medium of photoconductive elements illuminated by the radiation forming the image. An arrangement is provided in which the electroluminescent particles and the photoconductive particles form a single composite layer either by a coating of photoconductive material on the electroluminescent particles or by intermixing of the two substances so that the photoconductive material is effectively in shunt with the electroluminescent particles.

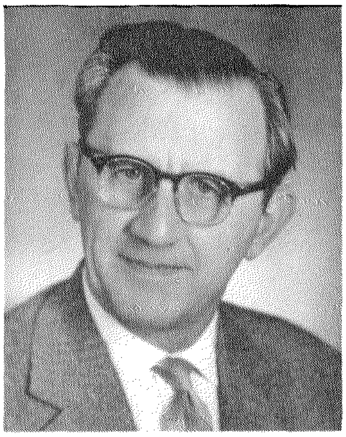
Circuit Element Having a Negative Resistance

2 843 765

P. R. R. Aigrain

An electric circuit taking advantage of the properties of semiconductor diodes, such as germanium diodes, was developed to provide a circuit element having a negative-resistance characteristic. A direct-current source and an alternating-current source are applied in series across a semiconductor rectifier. The alternating-voltage is of higher amplitude than the difference in voltage between the terminals of the direct-current source. The frequency of this alternating current is chosen so that the irregular carriers of electric charge injected into the semiconductor at a particular point in the cycle of the alternating voltage do not disappear completely during one period of the alternation.

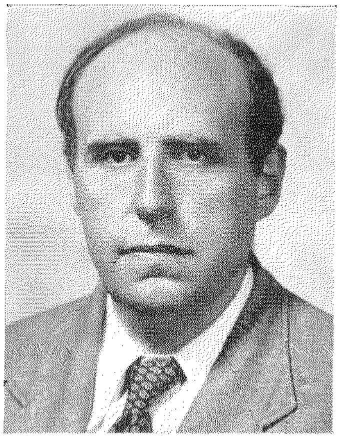
Contributors to This Issue



JOHANN AUGUSTIN

JOHANN AUGUSTIN was born in Berlin, Germany, on October 8, 1909. After practical training as a mechanic, he entered a public technical college from which he received an engineering degree in 1935.

Mr. Augustin has been employed by Standard Elektrik Lorenz since 1935. Prior to the second world war, he was a designer in the Berlin plant. After the war, he was placed in charge of a group concerned with the development of teleprinters including page and tape printers as well as tape perforators and readers. He is author of three papers in this issue on such equipment.



A. H. BECK

A. H. BECK was born in Norfolk, England, in 1916. He received a bachelor of science degree in engineering from University College, London.

After two years with Henry Hughes and Sons, he was assigned to the Admiralty Signals Establishment until the end of the war. In 1947, he left Hughes to establish the Enfield valve laboratory of Standard Telephones and Cables, which later became the valve division of Standard Telecommunication Laboratories. He is coauthor of the paper in this issue on gas tubes. In 1958, Mr. Beck joined the academic staff of Cambridge University as a lecturer in engineering. He is a member of Queens' College.

Mr. Beck is an Associate Member of the Institution of Electrical Engineers and a Fellow of the Institute of Radio Engineers. He is the author of three books on electronics, "Velocity Modulated Thermionic Tubes", "Thermionic Valves", and "Space-Charge Waves".

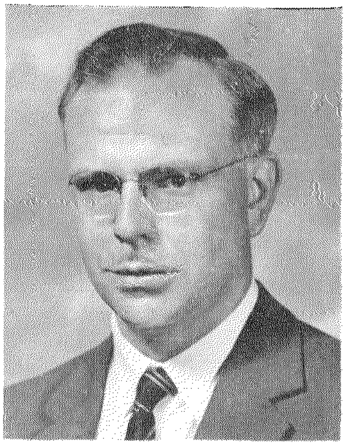
• • •

WILLIAM F. S. CHITTLEBURGH was born in London on October 2, 1923. He studied electrical-communication engineering at the City and Guilds College, London, where he received in 1943 an Associateship of the City and Guilds Institute together with a bachelor of science (engineering) degree.

On graduation, he joined Standard Telephones and Cables, and was initially associated with voice-frequency telegraphs. In 1948, after a period of working on coil and transformer design, he returned to the voice-frequency telegraph and signalling section, where he is at present in charge of such developments. He is the author of an article on voice-frequency telegraph equipment in this issue.

• • •

ROBERT A. FELSENHELD was born on February 15, 1910, in East Orange, New Jersey.



W. F. S. CHITTLEBURGH

He has been active in the broadcast field since 1927. In 1941, he joined ITT Laboratories and is now a senior project engineer in the radio communication laboratory. He has been active in the development of very- and ultra-high-frequency components, particularly antennas, transmission lines, and receivers. He reports in this issue on some over-the-horizon equipment.

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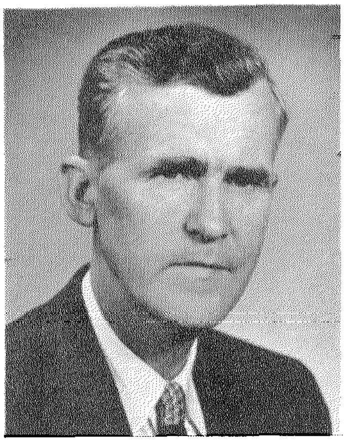
H. HAVSTAD was born on August 15, 1907. He received the degree of bachelor of science in electrical engineering from the College in Oslo, Norway, in 1931.

ROBERT A. FELSENHELD



H. Havstad

Mr. Havstad was the radio engineer for the Veslekari north-pole expedition in 1928. From 1931 to 1939, he did radio field engineering, installation, and operation for Tropical Radio Company and several merchant marine organizations.

He did research and development work at Bell Telephone Laboratories from 1939 to 1946. During the next two years, he was self-employed.

Mr. Havstad came to ITT Laboratories in 1948. His main line of work has been on pulse-modulation microwave links and on beyond-the-horizon systems. In this issue, he is coauthor of a paper in the latter field.

He is a member of the Institute of Radio Engineers.

• • •

J. D. Holland. A photograph and biography of Mr. Holland, author of the paper on transition-rate discriminators, appears in the September 1959 issue of *Electrical Communication.*

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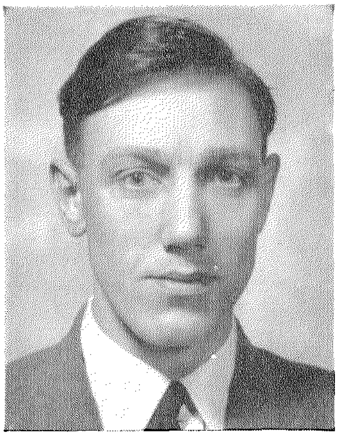

T. M. Jackson

T. M. Jackson was born in Merioneth, Wales in 1921.

He joined the Royal Air Force in 1940 and spent four years in experimental work on radar systems at the Government Telecommunication Research Establishment.

In 1946, he joined the Enfield valve laboratory of Standard Telephones and Cables and was transferred to Standard Telecommunication Laboratories in 1954. For several years he has been concerned with gas discharge phenomena and has been responsible for the development of special gas discharge devices, on which subject he is coauthor of a paper in this issue. Mr. Jackson is now doing microwave work, particularly on the problems of power generation at millimetric wavelengths.

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Bent Bulow Jacobsen

Bent Bulow Jacobsen was born in Kolding, Denmark, in 1906. He graduated in 1928 from the City and Guilds of London Engineering College.

He joined the International Telephone and Telegraph Laboratories, London, in 1929 and later transferred to Standard Telephones and Cables Limited. He has worked for many years on carrier system design and more recently also on microwave radio links. He has speciallized in the study of noise in long-distance transmission. Mr. Jacobsen is the author of the paper in this issue on probability theory in telephone transmission.

He is a Member of the Institution of Electrical Engineers and of the Institution of Danish Civil Engineers.

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Jack L. Jatlow was born on April 7, 1902. He received a bachelor of science degree in electrical engineering from Rensselaer Polytechnic Institute in 1924.

From 1924 to 1931, he was associated with the Conner Crouse Corporation. From 1931 to 1932, he was with Wired Radio Corporation, and for the following three years served as assistant chief engineer of F.A.D.Andrea Radio Corporation. From 1935 to 1940, he was employed by Photo Positive Corporation for research on photographic emulsions. He was chief engineer of Republic Engineering Products from 1941 to 1942.

In 1942, he joined Federal Telephone and Radio Corporation. Since his transfer to ITT Laboratories in 1954, Mr. Jatlow's main duties have been those of project manager for various beyond-the-horizon radio link systems. In this issue, he is coauthor of a paper reporting on one of these systems.

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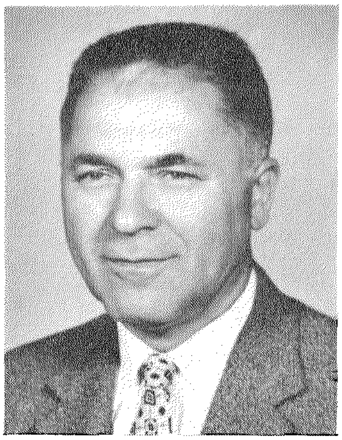
Donald J. LeVine was born in Brooklyn, New York, on October 10, 1921. He received two degrees in electrical engineering, the bachelor of science from the College of the City of



Jack L. Jatlow

Donald J. LeVine

New York in 1943 and the master of science from the Polytechnic Institute of Brooklyn in 1952.

From 1943 to 1946, he served in the Army Signal Corps working with radar and pulse-time-modulation microwave relay communication equipment. From 1946 to 1948, he was with the Microwave Research Institute of the Polytechnic Institute of Brooklyn, with a primary interest in microwave power-measuring devices.

In 1948, Mr. LeVine joined ITT Laboratories, where he has been active in the development of microwave systems, components, and antennas. He is coauthor of the paper in this issue on beyond-the-horizon radio equipment.

Mr. LeVine is a Senior Member of the Institute of Radio Engineers, an Associate of the American Institute of Electrical Engineers, and a registered professional engineer in the state of New York.

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Alexander William Montgomery

Alexander William Montgomery received the B.Sc. Tech. degree from Manchester College of Technology after having served for four years with the army during World War I.

In 1921, he joined the Western Electric Company in London and became closely associated with the early development in Great Britain of repeaters and carrier systems. He was active also in the field of voice-frequency telegraphy. He is now joint general manager and a director of Standard Telephones and Cables, Limited, and vice-chairman of Standard Telecommunication Laboratories, Limited. During World War II, he was actively engaged on the development of the British defense teleprinter network and of a wide range of communication equipment. He served on a number of government committees and, in 1944, visited the U.S.A. as a member of a Ministry of Supply mission. For his war work, he was made an Officer of the Order of the British Empire.

Mr. Montgomery represents the British telecommunications industry on a number of governmental and other committees.

Mr. Montgomery reports in this issue on the past and future of coaxial-cable communication systems.

He is a Member of the Institution of Electrical Engineers and of the American Institute of Electrical Engineers, and a Fellow of the Institute of Radio Engineers.

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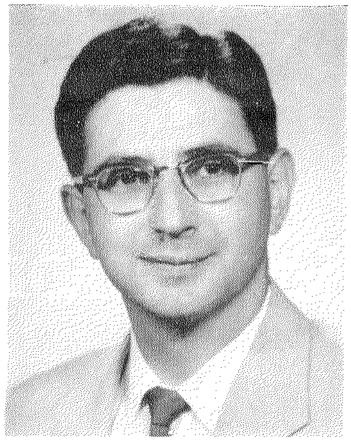

L. Pollack

L. Pollack was born in New York City on November 4, 1920. After receiving his bachelor of science degree in electrical engineering from the College of the City of New York in 1941, he worked at Fort Monmouth Signal Development Laboratory and in the Alaska Defence Command on the installation and maintenance of radar and communications equiment.

In 1943, Mr. Pollack joined ITT Laboratories, where he is now an executive engineer. His main course of work has been in the design of very- and ultra-high-frequency receivers and high-power transmitters. He is a coauthor of the article in this issue on beyond-the-horizon radio equipment.

Mr. Pollack is a Senior Member of the Institute of Radio Engineers. He is a member of the relay committee and of the television transmitter committee of the Electronic Industries Association.

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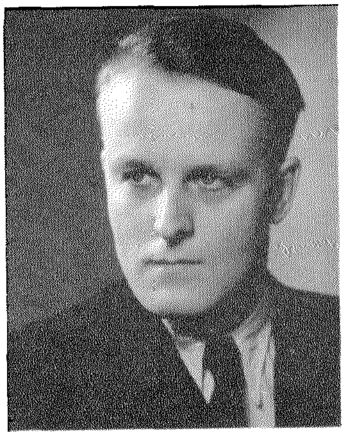

Werner Schiebeler

Werner Schiebeler was born in Bremen, Germany, on March 17, 1923. He received a diploma in physics in 1952, from Göttingen University. Three years later, he was awarded a doctorate in natural sciences from the Max Planck Institute in Göttingen.

Dr. Schiebeler has been on the staff of Standard Elektrik Lorenz in Pforzheim since 1955 and is employed in the development of electronic teleprinters and electronic input and output devices for data-handling systems. He reports in this issue on a teleprinter synchronizign set.

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Companhia Telefônica Nacional, Curitiba and Pôrto Alegre

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